ELECTRICAL CIRCUITS FOR CALUTRONS

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Electrical Circuits for Calutrons

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The existence of adequate experimental equipment, in particular the two operating cyclotrons of the Radiation Laboratory and the large 184-inch unit under construction, together with a nucleus of trained personnel, made it inevitable that work in connection with the war effort would be prosecuted vigorously at the University of California. Prior to the fall of 1941, studies of the properties of the transuranic elements were carried out and artificial radioactive materials were produced in the cyclotrons for use in various laboratories. This work was done informally and primarily on university funds. The importance of the studies of transuranic elements cannot be overestimated since the results formed a basis for the Plutonium Project.

Although the mass spectrographic method of separating uranium isotopes had been under consideration prior to the fall of 1941, there was no unanimity of opinion among physicists regarding the ultimate success of the method, owing to the space-charge effects. The feeling prevailed in the Radiation Laboratory of the University of California that in spite of this uncertainty the method should be pushed vigorously. The first concrete step in this direction was taken in November 1941, when a group was assigned to convert the 37-inch cyclotron to study this method of separating uranium isotopes. At about the same time two other groups started work on other electromagnetic separation schemes, namely, the ionic centrifuge and the radial magnetic separator. All this work was undertaken with the full support of the Uranium Committee but under no formal contract. The first formal contract designed to further work along these lines was entered into between the university and the Office of Scientific Research and Development in late December 1941, with the Laboratory Director as Project Leader.

The work on the mass spectrographic method, now called the “calutron process,” proceeded so satisfactorily that by the early fall of 1942 plans were being formulated for a production plant. Also, owing to the very gratifying results obtained with this method, it was decided to discontinue work on the other methods. From the fall of 1942 to
the end of hostilities in 1945 the Berkeley project was concerned primarily with the design and testing of prototype units for the plant, in addition to the necessary training of personnel. For a good portion of the time Radiation Laboratory personnel was stationed at Oak Ridge to assist directly in putting the plant into operation. It was on May 1, 1943, that the Berkeley project came directly under the jurisdiction of the Manhattan District. This move, however, did not affect the organizational setup of the Radiation Laboratory in any way, and the development work proceeded without any break.

Perhaps the outstanding factor with regard to the entire electromagnetic separation project lies in the general smoothness with which the work proceeded. It was necessary to build a large development laboratory from a relatively small university research laboratory in a matter of months. This involved greatly multiplying the personnel and increasing the physical facilities and necessary experimental equipment appropriately. In spite of the rapid expansion, personnel and organizational difficulties were inconsequential. The entire laboratory organization was characterized by a minimum of formal procedure consistent with the nature of the work. It is indeed remarkable that the scientific and technical personnel of the Radiation Laboratory, many of whom had been accustomed to the academic freedom of educational institutions, could adjust themselves so readily to the necessary security, governmental regulations, and group action of the project. It must also be kept in mind that the work was predominately of a developmental rather than research nature. The form of laboratory organization was such as to allow a maximum of individual expression with regard to the various problems encountered, which undoubtedly contributed considerably to a maximum of cooperation. The fact that the first unit of the Oak Ridge plant was built and put into operation successfully within a matter of two years from the time that the first mass spectrographic unit was built attests to the close cooperation maintained among all people concerned—the Office of Scientific Research and Development, Manhattan District officials, Radiation Laboratory personnel, and the manufacturing and operating companies. It would not be fair to say that the organization used would have been adopted if the project had been built up on a long-range basis. However, in view of the haste with which the project had to be carried through, it worked extremely well.

In preparing the report on the work done at the Radiation Laboratory, the major emphasis has been placed on those subjects of most interest to people working in related fields. The engineering aspects have been minimized in view of the fact that this phase of the work will be covered in other project literature. A number of papers deal-
ing with the chemical problems of the project will similarly be made available in separate reports.

It is impossible to pay proper tribute to the many individuals — scientific, technical, and nontechnical — who participated in the Berkeley project. A cross section of scientific and technical personnel is contained in this report, as authors of the various chapters and in the lists of references at the ends of the chapters. Others are referred to in the text. However, the names of many persons who contributed substantially to the progress of the project do not appear in this report.

The Office of the Director takes pleasure in expressing its deep appreciation to the project personnel for their unfailing loyalty and confidence; to the university as a whole for its support and cooperation; to the Area Engineer’s Office for its very effective expediting of all matters pertaining to the rapid development of the Project; to the plant construction contractor, Stone and Webster Engineering Corporation; to the operating company, Tennessee Eastman Corporation; and to the major manufacturing contractors, Allis-Chalmers Company, Westinghouse Electric and Manufacturing Company, and many others, for their close cooperation and effective handling of the engineering and operations problems.

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Chapter 1

GENERAL THEORY OF REGULATOR SYSTEMS

By Burton F. Miller

1. INTRODUCTION

The degree of regulation required for several of the principal power-supply units essential to operation of the calutron will be described in the succeeding chapters of this volume. In referring to these chapters, it will be noted that the required constancy of load voltage or current is of an order which cannot be directly attained with conventional power sources. Accordingly, several types of regulator systems have been developed which, when employed as auxiliaries to the basic power-supply units, permit the attainment of a sufficiently high degree of voltage and current stabilization to meet the normal calutron operating requirements. Although the elements of the various regulator systems employed differ rather considerably in form, each of the systems is fundamentally of the type known to communication engineers as a "feedback" or "degenerative" system, and a study of the inherent characteristics and properties of these regulators is, therefore, closely allied to a corresponding study of feedback amplifiers.

As might be supposed, a considerable number of difficulties attend the design of precision regulators for calutron load circuits. Because of the comparatively great voltage amplification required in such regulators, and the inevitable presence of undesired circuit coupling, considerable care must be exercised to ensure over-all system stability over the complete range of operating currents and voltages employed. The design problem is complicated by the fact that several of the regulator systems must normally operate at a high potential with respect to ground, and on circuits subject to abrupt and frequent breakdown. Unusual precautions must be taken in the shielding of certain of the regulator circuits in order to minimize the chances of erratic operation or failure due to high-voltage surges. Not the least of the diffi-
culties encountered has been that of providing voltage standard sources of the requisite constancy and reliability. These and other problems incident to the provision of adequate power-circuit regulation have been solved through the application of well-known principles of electric-circuit theory, and it is the purpose of this chapter to outline certain of the fundamental considerations upon which various regulator circuit designs have been based.

2. DESIGN OF PRECISION REGULATORS FOR CALUTRON LOAD CIRCUITS

Fundamentally, a degenerative voltage or current-regulator system functions by suitably amplifying the difference between a comparison voltage (derived from the actual load voltage or current) and a known and extremely constant standard voltage and by injecting the amplified voltage difference into the load circuit so as to oppose any deviation in the value of the load voltage or current from normal. In some instances where the regulation requirements are not too severe, the voltage standard may be replaced by a nonlinear bridge circuit, which employs one or two elements whose resistance is a function of the applied bridge voltage. When properly designed, the degenerative type of regulator may be made to provide a higher degree of load voltage or current stabilization than can be achieved by any other form of regulator system.¹-³

The form of treatment adopted in the analysis of regulator systems is, to a certain degree, a matter of choice. When a specific circuit is under investigation, its general form and degree of complexity may indicate the desirability of employing a treatment particularly well adapted to the emphasis of certain of its primary features. Keeping in mind, however, that questions of stability invariably arise in all but the simplest systems and that all degenerative regulator systems may be analyzed for stability by means of criteria which have been developed for feedback-amplifier systems, it appears most desirable to frame the elementary treatment of regulator systems so as to emphasize their equivalence to feedback-amplifier systems.

2.1 Basic d-c Voltage-regulator Circuit. The circuit chosen as basic for the general elementary discussion which follows is shown schematically in Fig. 1.1 and represents one of a number of possible circuit equivalents.

A d-c power source of internal resistance $R_0$ and open-circuit voltage $E_0$ supplies the principal component of the regulated load voltage $E_L$. A four-terminal regulator device $D$, whose internal resistance between its 3-4, or output, terminals is essentially constant and equal to $R_d$, provides a regulating component of load voltage. Ideally, $D$
would be a unilateral device, transmitting only from its 1-2, or input, terminals to its output terminals, and capable of producing an open-circuit voltage rise $E_{3-4}$, defined by the linear relationship

$$E_{3-4} = \mu(\omega)E_{1-2}$$  \hspace{1cm} (1)

where $\mu(\omega)$ is a function of the radian frequency $\omega$ of the input voltage $E_{1-2}$.

A fraction of the load voltage equal to $\beta E_L$, and derived from the voltage drop across resistor $R_1$ of the voltage divider formed by resistors $R_1$ and $R_2$, is compared with a constant standard voltage $E_s$. In some cases the quantity $\beta$ may be a function of $\omega$, but for the circuit shown it will be a real positive constant. The difference voltage $(E_s - \beta E_L)$ is applied to the input terminals of the regulator device $D$, which results in an output voltage

$$E_{3-4} = \mu(\omega) (E_s - \beta E_L)$$  \hspace{1cm} (2)

The total load voltage is given by

$$E_L = \frac{(E_0 + E_{3-4})R'_L}{R_0 + R_d + R'_L}$$  \hspace{1cm} (3)

where $R'_L$ is the parallel resistance formed by $R_L$ and $(R_1 + R_2)$.

Substituting Eq. 2 in Eq. 3, and solving for $E_L$, it is found

$$E_L = \frac{\mu(\omega)R'_L E_s + R'_L E_0}{R_0 + R_d + R'_L + \mu(\omega)\beta R'_L}$$  \hspace{1cm} (4)

Upon dividing the numerator and denominator of Eq. 4 by $(R_0 + R_d + R'_L)$, and setting

$$\frac{\mu(\omega)R'_L}{R_0 + R_d + R'_L} = \mu_0(\omega)$$  \hspace{1cm} (5)

$E_L$ may be expressed as

$$E_L = \frac{\mu_0(\omega)E_s + \frac{R'_L E_0}{1 + \mu_0(\omega)\beta}}{R_0 + R_d + R'_L}$$  \hspace{1cm} (6)

The quantity $\mu_0(\omega)$ represents the vector voltage amplification between the input terminals of $D$ and the effective load resistance $R'_L$ at
the radian frequency $\omega$. The quantity $\mu_0(\omega)\beta$ defines the vector voltage amplification which would be obtained between the terminals a-b and the terminals c-d of Fig. 1.1 if the indicated junctions at those points were broken and a termination equal to the input impedance of D were connected across terminals c-d. In regulator systems providing a high degree of load voltage or current stability, the quantity $\mu_0(\omega)\beta$ may attain maximum values of the order of several hundred to several thousand.

The improvement in load-voltage stability due to action of the regulator system may be determined by differentiation of Eq. 4. In particular, if $R_o$, $R_d$, $E_s$, and $\mu(\omega)$ are assumed to be constant, and if $R_1 + R_2 \gg R_L$, the variation in load voltage resulting from variations in supply voltage and load resistance may be expressed as

$$\delta E_L = \frac{R_L \delta E_0 + (R_0 + R_d)I_L \delta R_L}{(R_0 + R_d + R_L) [1 + \mu_0(\omega)\beta]} \tag{7}$$

where $I_L$ denotes the value of the normal load current. The corresponding variation in $E_L$ in the absence of regulator action will be denoted by $\delta E'_L$ and may be determined by setting $\beta = 0$ in Eq. 7.

Accordingly,

$$\delta E'_L = \frac{R_L \delta E_0 + (R_0 + R_d)I_L \delta R_L}{(R_0 + R_d + R_L)} \tag{8}$$

It is quite true, of course, that $R_d$ could be removed from the circuit if regulator action were not employed, and that as a result the value of $\delta E'_L$ would be somewhat influenced. It is convenient, however, to compare the load-voltage regulation of circuits which are identical in all respects save that one employs an active regulator and the other an inactive regulator. On this basis the stabilization factor $S$ of the regulator system is defined as the ratio of $\delta E'_L$ to $\delta E_L$. From Eqs. 7 and 8

$$S = 1 + \mu_0(\omega)\beta \tag{9}$$

It is evident from the preceding analysis that the regulator system inherently tends to reduce variations in load voltage arising from variations in either supply voltage or load-circuit resistance by the factor $S$. It is also apparent that load-voltage variations can be reduced to any desired minimum if $S$ can be given sufficiently large values. The fundamental aim of regulator circuit design is that of providing physical systems whose performance approaches that of the ideal regu-
lator without violating such limitations as may result from considerations of circuit stability or other system performance requirements.

2.2 Simplest Form of Electronic Voltage Regulator. The simplest possible form of electronic voltage regulator is the degenerative circuit of Fig. 1.2. The regulator device in this instance is the triode V-1 whose plate-cathode circuit appears in series with the power source and the load resistance. The grid battery $E_c$ serves both as a source of biasing potential for the tube and as a voltage standard. The value of $\beta$ for this circuit is unity. Any tendency toward a variation in load voltage produces an increment in grid voltage of such polarity that the accompanying variation in plate-circuit voltage drop tends to maintain $E_L$ at its initial value.

If the triode is assumed to exhibit a linear relationship between plate current and grid and plate voltages, and if the amplification factor and plate resistance of the tube are denoted by $\mu$ and $R_p$, respectively, application of Eq. 6 gives the load voltage as

$$E_L = \mu E_c + \frac{R_p E_o}{R_p + R_o + R_L}$$

where

$$\mu_0 = \frac{\mu R_L}{R_p + R_o + R_L}$$

and is independent of frequency.

The principal limitations of this circuit arise from the necessity of supplying a standard voltage source of essentially equal voltage to that of the load and from the difficulty of providing a high value of $\mu_0$ without introducing an objectionably large plate-voltage drop into the system.

2.3 Improved Form of Electronic Voltage Regulator. The deficiencies of this circuit are largely avoided by the circuit of Fig. 1.3, which has been taken from a paper by Hunt and Hickman. Basically this circuit indicates the addition of a single stage of amplification to the circuit of Fig. 1.2 and utilizes a glow discharge tube V-3 as a voltage standard. A fraction of the standard voltage is introduced in the cathode circuit of V-1 by means of the voltage divider consisting of resistors $R_5$ and $R_6$. Resistors $R_1$ and $R_3$ form the $\beta$ dividing network. Resistor $R_2$ provides the coupling between the plate circuit of V-1 and the grid circuit of the regulator tube V-2. If identical regulator tubes are employed in the circuits of Figs. 1.2 and 1.3, the improvement in
regulation provided by the latter circuit over the former is essentially
determined by the net voltage amplification between the load-circuit
terminals and the grid circuit of regulator tube V-2.

2.4 High-stabilization Voltage Regulator. When greater voltage
stabilization is required than can readily be obtained from the circuit
of Fig. 1.3 and particularly when rather high voltages are involved in
the load circuit, the somewhat more elaborate system of Fig. 1.4 is
frequently employed. In this system the voltage standard may itself
be a precision-regulated power source of moderate voltage, while the
regulator amplifier may employ two or more stages of amplification
to provide the relatively high gain required. The regulator tube V-1
is selected with due regard to the class of service in which it is em-
ployed and frequently takes the form of one of the larger water-cooled
power triodes. Variations in value of the regulated load voltage may be
secured through control of the value of either \( \beta \) or \( E_s \).

If the voltage amplification between the input circuit of the regulator
amplifier and the grid circuit of V-1 is constant within the frequency
band of interest and is denoted by \( \mu_a \), the corresponding amplification
between the grid circuit of the regulator tube and the load resistance
is denoted by \( \mu_r \), and the regulator tube is presumed to exhibit a linear
relationship between plate current and grid and plate voltages, it may
readily be demonstrated that the load voltage will be given by the
expression

\[
E = \frac{\mu_o E_s + \mu_r E_c + \frac{R'_L E_0}{R_o + R_p + R_L}}{1 + \mu_o \beta}
\]  

(12)

where

- \( E_c \) = biasing potential in the grid circuit of V-1
- \( R_p \) = plate resistance of the regulator tube
- \( R'_L \) = parallel resistance of the voltage divider and the load
  resistance
- \( R_o \) = internal resistance of the d-c power source
- \( \mu_o = \) product \( \mu_a \mu_r \)

When the regulator amplifier input resistance is very high compared
to the value of \( R_1 \) plus the internal resistance of the voltage standard,
the value of \( \beta \) is approximately equal to \( R_1/(R_1 + R_2) \).

If the regulator provides a high stabilization factor, the denom-
inator of Eq. 12 is practically equal to \( \mu_o \beta \), and the load voltage is
practically equal to

\[
E = \frac{E_s}{\beta} + \frac{E_c}{\mu_a \beta} + \frac{R'_L E_0}{\mu_o \beta (R_o + R_p + R_L)}
\]  

(13)
The relative importance of the three terms in this expression may be illustrated by introducing a set of typical values for the constants which might be employed in such a regulator system. Thus, if

\[
\begin{align*}
E_s &= 300 \text{ volts} \\
E_c &= -60 \text{ volts} \\
E_0 &= 36,000 \text{ volts} \\
\mu_a &= 1200 \\
\mu_o &= 36,000 \\
\beta &= 0.0100 \\
R_o &= 2000 \text{ ohms} \\
R_p &= 5000 \text{ ohms} \\
R_L &= 60,000 \text{ ohms}
\end{align*}
\]

these terms take on the approximate values \(E_s/\beta = 30,000 \text{ volts};\)
\(E_c/\mu_a \beta = -5.0 \text{ volts; }\) and \(R_L E_0/\mu_0 \beta (R_o + R_p + R_L') = 90 \text{ volts.}\) The importance of the term \(E_s/\beta\) is immediately apparent and points to the necessity for providing extreme constancy in the values of \(E_s\) and \(\beta\) if a high order of constancy in \(E_L\) is required. It is also evident that the source providing \(E_c\) need only be moderately well regulated.

An examination of Eq. 3 indicates that the regulation process may be considered the result of supplying a component of the total load voltage, equal to \(E_s - 4 R_L/(R_o + R_d + R_L'),\) through the medium of the regulating device. Obviously, cases may arise in which this component might be secured through appropriate control of the primary power source itself. For instance, if d-c power is supplied to the load by a motor-generator set, an adequate regulating component of voltage might be secured from the generator by permitting the regulator device to superimpose corrective components of current in the main field circuit of the generator. Similarly, if load power is obtained from a grid-controlled rectifier system, variations in rectifier output voltage, corresponding to corrective components of load voltage, may be secured by permitting the regulator-device output signal to control the rectifier firing angle. Other and more elaborate systems may be devised by an extension of the basic regulator-function concept.

A particularly simple equivalent circuit for the regulator system of Fig. 1.1 may be derived with the aid of Eq. 4. Since the regulated current for this circuit is given by \(E_L/R_L',\) the value of the regulated current becomes

\[
I'_L = \frac{\mu(\omega) E_s + E_0}{R_o + R_d + [1 + \mu(\omega)\beta] R_L'} \tag{14}
\]
Upon dividing the numerator and denominator of this equation by 
$1 + \mu(\omega)\beta$,

$$I_L' = \frac{\frac{\mu(\omega)E_s + E_0}{1 + \mu(\omega)\beta}}{\frac{R_o + R_d}{1 + \mu(\omega)\beta}}$$  \hspace{1cm} (15)$$

It will be noted that the current given by this equation is precisely that 
which would exist in the equivalent circuit of Fig. 1.5 if the power 
source provided an open-circuit voltage

$$E_o' = \frac{\mu(\omega)E_s + E_0}{1 + \mu(\omega)\beta}$$  \hspace{1cm} (16)$$

and exhibited an internal impedance

$$R_o' = \frac{R_o + R_d}{1 + \mu(\omega)\beta}$$  \hspace{1cm} (17)$$

It is evident from Eq. 17 that the action of the regulator circuit re-
results in an apparent reduction of the power source and regulator-device 
resistances, as viewed from the load circuit, by the factor $[1 + \mu(\omega)\beta]$.

2.5 Basic d-c Regulator Circuit. Regulator systems designed for 
the purpose of maintaining load current at a constant value are basi-
cally similar in form to those intended for load-voltage regulation. 
The circuit shown schematically in Fig. 1.6 represents a degenerative 
current-regulator system, and it will be noted that it differs from the 
basic voltage-regulator circuit only in the manner employed to obtain 
a comparison voltage for the regulator-device input circuit. In this 
instance the comparison voltage is developed across a resistor $R_1$, 
which is in series with the load resistance and which therefore pro-
vides a voltage proportional to load current.

Analysis of the circuit of Fig. 1.6 follows the same procedure as 
that employed for the case of the voltage regulator. The expression 
for load current may readily be derived as

$$I_L = \frac{\mu_0(\omega)E_s}{R_L [1 + \mu_0(\omega)\beta]} + \frac{E_0}{(R_o + R_d + R_1 + R_L) [1 + \mu_0(\omega)\beta]}$$  \hspace{1cm} (18)$$

where $\mu_0(\omega) = \mu(\omega)R_1/(R_o + R_d + R_1 + R_L)$ and $\beta = R_1/R_L$. If $R_o$, $R_d$, $R_1$, 
$E_s$, and $\mu(\omega)$ are assumed constant, the total variation in load current 
is given by
\[ \delta I_L = \frac{\delta E_m - I_L \delta R_L}{(R_o + R_d + R_1 + R_L) [1 + \mu_o(\omega)\beta]} \]  

and it is apparent that the regulator tends to reduce load-current variations relative to those existing in an unregulated system by the factor \( S = [1 + \mu_o(\omega)\beta] \). The essential similarity between voltage and current-regulator systems is thus demonstrated.

2.6 A-c Load-current Regulator. Voltage- and current-regulator systems for alternating current load circuits differ from corresponding d-c regulators more in detail than in principle. In one commonly employed a-c current regulator, indicated schematically in Fig. 1.7, the comparison signal derived from the voltage drop across \( R_1 \) is converted to d-c voltage by means of a conventional resistance-capacitance filtered diode rectifier unit \( U \). The rectifier output voltage is compared with a standard voltage, and the resultant difference voltage is amplified to provide a signal of sufficient magnitude to control a regulating device which operates in the a-c power circuit proper. This device may take the form of a saturable reactor, a thyatron or ignitron variable conductance unit, or one of a variety of variable element a-c networks.

The basic features common to the voltage- and current-regulator systems described may be extended to systems intended for the regulation of such other variables as temperature, pressure, position, velocity, power, light intensity, and frequency. In all cases, provision must be made for sampling the magnitude of the variable to be regulated, comparing the sample against a suitable standard, amplifying the difference between the sample and standard values, and employing this amplified difference so as to minimize the tendency for deviation of the regulated variable from normal. The required amplification may be secured by electrical, electronic, optical, mechanical or other means, or by any combination of such means.

3. OVER-ALL REGULATOR-SYSTEM STABILITY

While each of the regulator systems associated with calutron power units is required to meet a number of basic performance specifications, none of these is of greater importance than the requirement of over-all regulator-system stability. As here employed, the term "stability" is used to denote freedom from any tendency of the regulator system to generate oscillations of sustained or continuously increasing amplitude in the regulated variable.

Concern with respect to the stability of regulator systems arises primarily from a consideration of the quantities involved in the stabilization factor \( S \). It will be recalled that the degenerative regulator
functions by virtue of the phase opposition which exists between changes in input voltage to the regulator device and the resultant changes produced in load voltage. This phase relationship can be approximately maintained over a relatively restricted band of frequencies only, since a sufficiently high time rate of change of input voltage to the regulator inevitably brings circuit reactances into play, which tend to destroy the optimum phase relationship required for degenerative action. At a sufficiently high-frequency variation of regulator input voltage, the resultant load-voltage variation may be essentially in phase with the regulator input voltage and, if the feedback voltage is equal to or greater than the regulator input voltage, a condition of instability will result. It may be assumed, therefore, that certain restrictions must be placed on the characteristics of the function \( \mu_\omega(\omega) \), and accordingly on \( S \), if stable operation is to be secured.

The range of frequencies over which the product \( \mu_\omega(\omega)\beta \) is required to exhibit a large positive real component is determined by the nature of the deviations of the regulated variable from normal. The deviation waveform is generally dependent upon a number of unrelated factors, such as the nature of random variations in load-circuit parameters, intentional variations in circuit loading, temperature effects in various circuit components, and random variations in voltage of the primary power source. Accordingly, it may be anticipated that the deviation waveform will contain components within the frequency range from zero to some upper limiting value determined largely by the electrical characteristics of the power source and the load. Ideally, the regulator system would be capable of reducing all deviation waveform frequency components by a like amount, but unfortunately practical design limitations do not permit the attainment of such performance.

The earliest studies relating to system stability were those concerned with the mechanical systems of classical dynamics. The work of Routh is of particular interest in this connection because of the close parallelism between one of the criteria for stability which he formulated and the criterion for feedback-amplifier stability later formulated by Nyquist. Unfortunately, Routh's conclusions were not phrased so as to readily permit their association with concepts derived from the steady-state analysis of electrical circuits, and they have not been directly applied to the development of electrical-circuit theory.

The Nyquist criterion of stability provides a tool of basic importance to the design of all regenerative and degenerative systems and hence warrants rather detailed consideration. It was derived from an investigation of the conditions for stability in a system consisting of a linear amplifier in tandem with a feedback network. Nyquist's original treat-
ment was based upon a Fourier integral formulation of the system response to a small impressed transient disturbance. An equally rigorous derivation of the criterion has been obtained by a study of the transient responses of such a system in terms of the differential equations which apply. The latter method will be utilized here because of its similarity to methods commonly employed in the analysis of a variety of communication circuits.

3.1 Application to Simple Resonant Circuit. The general procedure in this method of analysis may be reviewed briefly by considering its application to a simple resonant circuit such as that of Fig. 1.8. The equation for voltage equilibrium in the system after closure of switch \( K \) is

\[
L \frac{di}{dt} + Ri + \frac{1}{C} \int i \, dt = E
\]  

(20)

Upon differentiating this equation with respect to time, replacing the operator \( d/dt \) by \( p \), and multiplying by \( C \), there results the equation

\[
(LCp^2 + RCp + 1)i = 0
\]  

(21)

which may be factored and written as

\[
(p - p_1)(p - p_2)i = 0
\]  

(22)

The quantities \( p_1 \) and \( p_2 \) are the roots in \( p \) of Eq. 21 and are given by

\[
p_1 = -\frac{R}{2L} + \sqrt{\left(\frac{R}{2L}\right)^2 - \frac{1}{LC}}
\]  

(23)

\[
p_2 = -\frac{R}{2L} - \sqrt{\left(\frac{R}{2L}\right)^2 - \frac{1}{LC}}
\]

The solution of Eq. 22 is known to be of the form

\[
i = A_1e^{p_1t} + A_2e^{p_2t}
\]  

(24)

where \( A_1 \) and \( A_2 \) are constants determined by boundary conditions applicable to the circuit under consideration. If it is assumed that no current or charge existed in the circuit prior to closure of the switch, an evaluation of the integration constants provides the solution

\[
i = \frac{E}{(p_1 - p_2)L} \left( e^{p_1t} - e^{p_2t} \right)
\]  

(25)
In the case where \((R/2L)^2 > 1/LC\), \(p_1\) and \(p_2\) are negative real terms, and the current given by Eq. 25 represents the difference between two simple exponentially decreasing functions of time.

When \((R/2L)^2 < 1/LC\), the roots \(p_1\) and \(p_2\) may be written as

\[
p_1 = \alpha + j\omega \\
p_2 = \alpha - j\omega
\]

where \(j\) represents the imaginary operator \(\sqrt{-1}\). Substituting Eq. 26 in Eq. 25, the current for this case is given by

\[
i = \frac{E e^{\alpha t}}{\omega L} \left( e^{j\omega t} - e^{-j\omega t} \right) = \frac{E e^{\alpha t} \sin \omega t}{\omega L}
\]

The quantity \(\alpha\) is evidently a measure of the circuit damping, while \(\omega\) defines the radian frequency of current oscillation. If \(\alpha\) is a large negative quantity, the current given by Eq. 27 represents a rapidly decaying oscillatory transient. As \(\alpha\) approaches zero, the rate of decay diminishes; when \(\alpha\) is equal to zero the resulting current represents a continuous undamped sinusoidal oscillation. If \(\alpha\) takes on positive values, the current is oscillatory in nature and increases in amplitude as an exponential function of time. In accordance with the previously given definition of circuit stability, the resonant circuit would be considered unstable for values of \(\alpha\) equal to or greater than zero.

It is important to note that the stability of this circuit might have been directly determined from the nature of the roots of the differential equation (Eq. 22), for it is known that the complementary solution of any linear differential equation may be expressed as a sum of exponential terms, and that if the auxiliary equation contains complex roots with positive real parts, one or more pairs of the exponential terms must be equivalent to exponentially increasing real oscillatory functions of the independent variable. Thus the stability of even highly complex circuits may be determined solely from a knowledge of the roots of the differential equations defining the transient response of those circuits.

When the differential equations for a network are of higher order than the third or fourth, it is generally extremely difficult to secure a solution for the roots. Fortunately, the stability of a system is not dependent upon the precise value of the roots, but rather upon the absence of roots having positive real parts.
3.2 The p Plane. If the component of the root \( p_r \) of an \( n \)th order differential equation is denoted by \( \alpha_r \), and the imaginary component by \( j\omega_r \), the root \( p_r \) may be represented as a point in the complex plane shown in Fig. 1.9. The axis of reals is chosen to correspond with the damping factor \( \alpha \), while the axis of imaginaries is chosen to correspond with the real frequency component \( \omega \). The entire plane may be referred to as the "p plane."

It will be evident that all roots of stable systems must fall in the half-plane to the left of the frequency axis, while one or more of the roots of an unstable system will lie in the half-plane to the right of the frequency axis. Thus a system having the roots A, B and B', or C and C' would be stable. A system with any or all of these roots in addition to one or more roots such as D or E and E' would be unstable. A desirable criterion for degenerative-system stability would provide a ready determination by analytical or experimental means of the existence of roots for that system in the right half of the p plane without necessarily indicating their actual values. Such a criterion was established by Nyquist.

3.3 Current and Voltage Relationship. The theory underlying the Nyquist criterion may now be formulated in general terms. To this end the system of Fig. 1.10 may be considered, where T represents a multimesh linear network which may consist of both passive and active elements. Assume that the network consists of \( n \) meshes with voltages \( E_1 \) and \( E_n \) applied to the first and \( n \)th meshes, respectively. The instantaneous voltage drop in mesh \( k \) caused by the current \( i_k \) with all other meshes on open circuit will be written

\[
(L_{kk}p + R_{kk} + \frac{1}{C_{kk}p}) \quad i_k = W_{kk}i_k
\]  

(28)

where \( L_{kk}, R_{kk}, \) and \( C_{kk} \) represent the total inductance, resistance, and capacitance of the \( k \)th mesh, respectively. Similarly, the voltage induced in mesh \( k \) caused by the current \( i_j \) in mesh \( j \) will be written

\[
(L_{kj}p + R_{kj} + \frac{1}{C_{kj}p}) \quad i_j = W_{kj}i_j
\]  

(29)

where \( L_{kj}, R_{kj}, \) and \( C_{kj} \) represent the mutual inductance, resistance, and capacitance, respectively, which couple the \( j \)th mesh to the \( k \)th mesh.

Employing this nomenclature, the system of differential equations defining the network response to the voltages \( E_1 \) and \( E_n \) may be written
\[ W_{1i_1} + W_{12}i_2 + W_{13}i_3 + \ldots + W_{1n}i_n = E_1 \]
\[ W_{2i_1} + W_{22}i_2 + W_{23}i_3 + \ldots + W_{2n}i_n = 0 \]
\[ W_{3i_1} + W_{32}i_2 + W_{33}i_3 + \ldots + W_{3n}i_n = 0 \]
\[ W_{ni_1} + W_{n2}i_2 + W_{n3}i_3 + \ldots + W_{nn}i_n = E_n \]

If all of the driving voltages on the right-hand side of Eq. 30 are set equal to zero, the resulting system of equations defines the free transient response of the network. Since there are then \( n \) unknown currents and \( n \) equations, the existence of transient currents will require that the determinant \( \Delta \) of the coefficients in the left-hand side of Eq. 30 shall vanish. The values of \( p \) which represent the free modes of the system are, therefore, those for which

\[ \Delta = 0 \]  

and if the system is stable none of the roots in \( p \), derived from a solution of Eq. 31, may have positive real parts.

The solution of Eq. 30 for the currents \( i_1 \) and \( i_n \) due to the impressed voltages \( E_1 \) and \( E_n \) may be expressed as

\[ i_1 = \frac{\Delta_{11}E_1}{\Delta} - \frac{\Delta_{1n}E_n}{\Delta} \]  

\[ i_n = \frac{\Delta_{n1}E_1}{\Delta} - \frac{\Delta_{nn}E_n}{\Delta} \]

where the coefficients of the form \( \Delta_{jk} \) represent the cofactors formed by omitting the \( j \)th row and \( k \)th column of \( \Delta \). For convenience, the cofactors will be presumed to carry their own signs. It should be noted that, if \( \Delta \) does not contain roots in \( p \) having real positive parts, neither will the cofactors.

A limitation is now placed on the character of the network \( T \) by assuming that it contains one or more unilateral elements which permit transmission from the 1-2 terminals of the network to the 3-4 terminals, but prevent transmission in the reverse direction. This condition requires that \( \Delta_{1n} = 0 \), so Eqs. 32 and 33 become

\[ i_1 = \frac{\Delta_{11}E_1}{\Delta} \]

\[ i_n = \frac{\Delta_{n1}E_1}{\Delta} - \frac{\Delta_{nn}E_n}{\Delta} \]
Now let the network be iteratively terminated, and assume that the voltage rise $E_n$ is solely due to the flow of $i_n$ through the network termination. The ratio of voltage to current in this termination is identical with the ratio $E_1/i_1$ and is equal to $\Delta/\Delta_{11}$. Under this condition the current $i_n$ resulting from the application of a network input voltage $E_1$ is given by

$$i_n = \frac{\Delta_{n1}E_1}{\Delta} - \frac{\Delta_{nn}}{\Delta} \left( \frac{\Delta}{\Delta_{11}} \right) i_n$$

or

$$i_n = \frac{\Delta_{11}\Delta_{n1}E_1}{\Delta(\Delta_{11} + \Delta_{nn})}$$

Employing the relation $E_n = i_n\Delta/\Delta_{11}$

$$E_n = \frac{\Delta_{n1}E_1}{\Delta_{11} + \Delta_{nn}}$$

The cofactors $\Delta_{11}$, $\Delta_{n1}$, and $\Delta_{nn}$ are functions of $p$, and $E_n$ may therefore be expressed as

$$E_n = E_1 A(p)$$

It is next assumed that the input and output terminals of the network are connected, which again corresponds to an iterative termination. Under this condition $i_i = i_n$; consequently Eqs. 34 and 35 give

$$(\Delta_{11} - \Delta_{n1})E_1 + \Delta_{nn}E_n = 0$$

From Eqs. 38 and 39

$$\Delta_{nn}E_n = \Delta_{n1}E_1 - \Delta_{11}E_1 A(p)$$

Combining Eqs. 40 and 41

$$\Delta_{11} [1 - A(p)] E_1 = 0$$

Since the network was assumed stable before the input and output terminals were connected, $\Delta_{11}$ cannot have any roots in $p$ with positive real parts. Therefore if the closed feedback system is unstable, it must arise from the fact that the quantity $[1 - A(p)]$ has such roots.
Each of the cofactors $\Delta_{11}$, $\Delta_{n1}$, and $\Delta_{nn}$ is a rational polynomial in $p$, and accordingly $A(p)$ may be expressed as

$$A(p) = \frac{F_m p^m + F_{m-1} p^{m-1} + F_{m-2} p^{m-2} + \ldots + F_0}{G_n p^n + G_{n-1} p^{n-1} + G_{n-2} p^{n-2} + \ldots + G_0} \quad (43)$$

where the $F$'s and $G$'s are constants. Because of the fact that the transmission of any system must approach zero as frequency is permitted to increase without limit, the denominator of Eq. 43 must be of higher order than the numerator. Employing Eq. 43

$$[1 - A(p)] = \frac{H_n p^n + H_{n-1} p^{n-1} + H_{n-2} p^{n-2} + \ldots + H_0}{G_n p^n + G_{n-1} p^{n-1} + G_{n-2} p^{n-2} + \ldots + G_0} \quad (44)$$

which may be factored to give

$$[1 - A(p)] = \frac{(p - p_1) (p - p_2) (p - p_3) (p - p_4) \ldots}{(p - p_a) (p - p_b) (p - p_c) (p - p_d) \ldots} \quad (45)$$

The normal modes of the feedback circuit are determined by the roots $p_1$, $p_2$, $p_3$, etc., while the roots $p_a$, $p_b$, $p_c$, etc., are those of the quantity $(\Delta_{11} + \Delta_{nn})$. The assumed stability of the normal network requires that $p_a$, $p_b$, $p_c$, etc., contain no real positive components. Instability of the feedback circuit, if it exists, must arise through the nature of the roots $p_1$, $p_2$, $p_3$, etc.

It will be recalled from the previous discussion of stability that roots of $[1 - A(p)]$ which lead to system instability must lie in the right half of the $p$ plane. In normal practice, the experimental determination of system transmission characteristics is based upon measurements made with sinusoidal applied voltages for which $p$ is pure imaginary. It is desirable, therefore, to evolve a method of determining the existence of roots in the right half of the $p$ plane in terms of some property displayed by $[1 - A(p)]$ when $p$ is given a succession of pure imaginary values. To this end, an important theorem relating to line integrals in the complex plane may be employed. This theorem states that the integral

$$J = \frac{1}{2\pi j} \int \frac{f'(z) \, dz}{f(z)} \quad (46)$$

where $f(z)$ is a function of the complex variable $z$, and $f'(z)$ is its derivative when taken in the positive sense along the complete boundary of a domain in which $f(z)$ is everywhere regular except at poles.
and is equal to the number of zeros of $f(z)$ in this domain diminished by the number of poles, with every zero and every pole counted as often as its order of multiplicity indicates.

3.4 Path of Integration in $p$ Plane. To apply this theorem to the problem of stability, set the complex variable $z = p$, and

$$f(p) = [1 - A(p)]$$
$$f'(p) = -A'(p)$$

Since any roots of $f(p)$ in the right half of the $p$ plane must lie within a finite distance of the origin, all such roots will be included if the contour along which the integral $J$ is evaluated is taken of infinite extent. It is customary to choose a path of integration such as that shown in Fig. 1.11, which consists of a large semicircle lying in the right half of the $p$ plane, centered on the origin and closed by a diameter lying on the imaginary axis. As the radius of the semicircle is permitted to increase without limit, all roots of $f(p)$ eventually must be enclosed. However, it is apparent from an inspection of Eq. 45 that as $p$ is assigned larger and larger values along the semicircle, the value of $f(p)$ must approach unity and be independent of the position of $p$ on the semicircle. The value of $f(p) \, dp$ therefore approaches zero, and consequently the value of the integral $J$, taken along the semicircle, approaches zero as $p$ approaches infinity. Any finite value of $J$ must then result from integration along the path defined by the imaginary axis for which $p = j\omega$ and which corresponds to integration through the range of real frequencies from $-\infty$ to $+\infty$. Hence the conclusion is that the system will be stable if the integral

$$J = \frac{1}{2\pi j} \int \frac{-A'(p) \, dp}{1 - A(p)}$$

taken along the imaginary axis of the $p$ plane is equal to zero and that instability results when this integral differs from zero.

A point $p_r$ lying in the $p$ plane was previously expressed in terms of its real and imaginary components as $p_r = \alpha_r + j\omega_r$. Similarly, the rational function $f(p)$ may be expressed as $f(p) = u + jv$, where the quantities $u$ and $v$ are real variables and each is a function of the quantities $\alpha$ and $\omega$. Any specific value of $f(p)$ may be represented as a point on the $f$ plane, where values of $u$ are measured along an axis of reals and values of $v$ are measured along an axis of imaginaries. For any given function $f(p)$, each point on the $p$ plane is identified with a related point on the $f$ plane, and for each curve in the $p$ plane there exists a related curve in the $f$ plane.
Any point $p_r$ of the $p$ plane which corresponds to a root of the equation $f(p) = 0$ will evidently lie at the origin of the $f$ plane. It may readily be demonstrated that for each curve $C$ which completely encircles the root $p_r$ in the $p$ plane there exists a related curve $C'$ which completely encircles the origin of the $f$ plane. Accordingly, if the integral $J$, taken around the previously prescribed contour of the $p$ plane, encloses any roots of $f(p) = 0$, the corresponding contour of the function $f(p)$ in the $f$ plane must enclose the origin. But it has been noted that contributions to the value of the integral $J$, taken over the prescribed contour, can arise only from values of $p$ corresponding to points along the axis of imaginaries in the $p$ plane. Hence, if roots of $f(p) = 0$ lie in the right half of the $p$ plane, the curve $f(p)$, obtained by assigning $p$ a sequence of pure imaginary values between the limits $-j\infty$ and $+j\infty$, must enclose the origin of the $f$ plane.

The possibility of determining the existence of roots of $f(p)$ leading to system instability through a study of that function when $p$ is given pure imaginary values indicates that experimental or analytical procedures widely employed in the steady-state analysis of a-c circuits may be utilized in the determination of system stability. Thus, if it is determined that the curve representing $f(j\omega)$ encloses the origin of the $f$ plane when $\omega$ is permitted to vary from $-\infty$ to $+\infty$, the system is unstable. Correspondingly, the system is unstable if the curve representing $A(j\omega)$ in the $f$ plane, for the same range of values of $\omega$, encloses the point $(1,j0)$. Derivation of the Nyquist criterion is completed by noting the equivalence between the function $A(j\omega)$ and the product $-\mu(j\omega)\beta$ involved in the stabilization factor. If the function $A(j\omega)$ or $-\mu(j\omega)\beta$ is designated as the system 'transfer factor,' the Nyquist criterion for stability may be expressed as follows: Plot the imaginary part of the transfer factor against the real part for all frequencies from $-\infty$ to $+\infty$. If the point $(1,j0)$ lies completely outside the resultant curve the system is stable; if not it is unstable.

Reid has suggested a simple graphical interpretation of the stability criterion based upon considerations relating to Eq. 45. In general, $p$ may be assigned any arbitrary value in the complex $p$ plane. The roots $p_1, p_2, p_3, \ldots, p_a, p_{b}, p_{c}, \ldots$ are fixed points in this plane. More particularly, $p_1, p_2,$ etc., may lie in any portion of the $p$ plane, but $p_a, p_b, p_{c},$ etc., must lie in the left half-plane because of stability requirements when the feedback loop is opened.

Each quantity of the form $(p - p_r)$ appearing in both the numerator and denominator of Eq. 45 may be regarded as a vector, as may the function $[1 - A(p)]$. Now a sequence of pure imaginary values, $j\omega$, may be assigned to $p$ so that the left-hand member of Eq. 45 becomes $[1 - A(j\omega)]$. As $p$ varies along the imaginary axis from $-j\infty$ to $+j\infty$,
the vector \((p - p_a)\), indicated by \(r_1\) of Fig. 1.12, turns counterclockwise through an angle of 180 deg since \(p_a\) is to the left of the axis of imaginaries. The reciprocal of this vector, \(1/(p - p_a)\), therefore turns through an angle of 180 deg in a clockwise direction. Consequently the reciprocal of each vector in the denominator rotates 180 deg in the clockwise direction as \(p\) varies from \(-j\omega\) to \(+j\omega\).

By similar reasoning, every vector in the numerator whose root \(p_r\) lies in the left half-plane rotates 180 deg in the counterclockwise direction, while every vector whose root lies in the right half-plane rotates 180 deg in the clockwise direction as \(p\) varies from \(-j\omega\) to \(+j\omega\). If all roots of both the numerator and denominator lie in the left half-plane, the rotation of the vectors of the numerator will be exactly balanced by the counterrotation of the vectors of the denominator, and the resultant vector \([1 - A(j\omega)]\) experiences a net angular rotation of zero. This, of course, corresponds to a condition of system stability. On the other hand, every root of the numerator lying to the right of the axis of imaginaries will give, with its associated term from the denominator, a net rotation of 360 deg to the vector \([1 - A(j\omega)]\). Hence the vector \([1 - A(j\omega)]\) traces out a closed curve about the origin if roots leading to system instability are present.

Because of the wide field of application and the value of so general and precise a criterion of stability as that formulated by Nyquist, it was subjected to experimental test by Peterson, Kreer, and Ware, and fully confirmed within the limits of experimental error.

The preceding derivation of the criterion of stability has implied the existence of only a single feedback loop within the system. Recent years have witnessed the constantly increasing application of multiple feedback loop systems in regulator and amplifier service. Such systems may be analyzed by repeatedly applying the stability criterion after closure of each of the minor or internal feedback loops of the system until all loops have been taken into account. When the order of loop closure determines the stability of portions or the whole of the system, a state of conditional stability exists because a change or failure in one of the system elements may result in oscillation of the system.

### 3.5 Degenerative d-c Voltage Regulator

By way of illustrating the application of the stability criterion, the simplified degenerative d-c voltage regulator shown schematically in Fig. 1.13 will be considered. All sources of power except that for the load circuit have been omitted to reduce the circuit to its basic components. The source \(E_0\) is presumed to exhibit zero impedance at all frequencies. The load circuit consists of the resistor \(R_L\), paralleled by the capacitor \(C_L\). The voltage divider consists of resistors \(R_3\) and \(R_5\), whose combined resistance is assumed very much higher than that of the load.
When the switch K is thrown to position 1, a test signal $\delta E_1$ may be applied to the grid circuit of the first amplifier stage, and the resultant feedback signal $\delta (E_s - \beta E_L)$ may be measured between the points c-d. The regulator is operative when K is thrown to position 2.

Let it be assumed that K is in position 1. Let $\mu_1$, $\mu_2$, and $\mu_3$ represent the amplification factors of the first, second, and third tubes, respectively; let $r_{p1}$, $r_{p2}$, and $r_{p3}$ represent the corresponding tube plate resistances, and $Z_1$, $Z_2$, and $Z_3$ the corresponding tube load impedances. A linear relationship between plate current and grid and plate voltages of the form

$$i_b = \frac{\mu e_c + e_b}{r_p} \quad (49)$$

is assumed for each of the three tubes. Then by the definitions previously given

$$\frac{\delta (E_s - \beta E_L)}{\delta E_1} = \mu_0(\omega) \beta \quad (50)$$

where $\beta = R_a/(R_a + R_b)$ and where

$$\mu_0(\omega) = \left( \frac{\mu_1 Z_1}{r_{p1} + Z_1} \right) \left( \frac{\mu_2 Z_2}{r_{p2} + Z_2} \right) \left( \frac{\mu_3 Z_3}{r_{p3} + Z_3} \right) \quad (51)$$

when determined in accordance with the indicated positive senses of $E_1$ and $E_L$.

For the circuit shown

$$Z_1 = \frac{-j \frac{R_1}{\omega C_1}}{R_1 \left( -j \frac{1}{\omega C_1} \right)} \quad (52)$$

and, upon making the substitution

$$T_1 = \frac{r_{p1} R_1 C_1}{r_{p1} + R_1} \quad (53)$$

where $T_1$ represents the time constant of the circuit consisting of $C_1$, $r_{p1}$, and $R_1$ in parallel,

$$\frac{\mu_1 Z_1}{r_{p1} + Z_1} = \frac{\mu_1 R_1}{r_{p1} + R_1} \left( \frac{1}{1 + j\omega T_1} \right) \quad (54)$$
The coefficient of the term in parentheses in Eq. 54 is equal to the d-c voltage amplification of the first stage, and if this quantity is denoted by \( \mu_{o1}(0) \),

\[
\frac{\mu_1 Z_1}{r_{p1} + Z_1} = \frac{\mu_{o1}(0)}{1 + j\omega T_1}
\]  
(55)

In a similar manner

\[
\frac{\mu_2 Z_2}{r_{p2} + Z_2} = \frac{\mu_{o2}(0)}{1 + j\omega T_2}
\]  
(56)

and

\[
\frac{\mu_3 Z_3}{r_{p3} + Z_3} = \frac{\mu_{o3}(0)}{1 + j\omega T_3}
\]  
(57)

where

\[
T = \frac{r_{p2} R_2 C_2}{r_{p2} + R_2} \quad T_3 = \frac{r_{p3} R_L C_L}{r_{p3} + R_L}
\]  
(58)

Upon setting the product \( \mu_{o1}(0)\mu_{o2}(0)\mu_{o3}(0) = \mu_o(0) \), where the quantity \( \mu_o(0) \) denotes the over-all d-c voltage amplification of the three stages, the substitution of Eqs. 55, 56, and 57 in Eq. 51, followed by multiplication of the resultant expression by \( \beta \), gives

\[
\mu_o(\omega)\beta = \frac{\mu_o(0)\beta}{(1 + j\omega T_1)(1 + j\omega T_2)(1 + j\omega T_3)}
\]  
(59)

In accordance with the rule given for determining circuit stability, the function

\[
-\mu_o(\omega)\beta = \frac{-\mu_o(0)\beta}{(1 + j\omega T_1)(1 + j\omega T_2)(1 + j\omega T_3)}
\]  
(60)

must not encircle the point \( 1 + j0 \) if the system is to be stable.

The simplicity of the circuit in this instance permits a direct determination of the requirements for stability if Eq. 60 is expressed as

\[
-\mu_o(\omega)\beta = \frac{-\mu_o(0)\beta}{[1 - \omega^2(T_1 T_2 + T_1 T_3 + T_2 T_3)] + j\omega (T_1 + T_2 + T_3 - \omega^2 T_1 T_2 T_3)}
\]  
(61)
The transfer factor experiences a 180-deg phase shift when \( \omega \) takes on such a value that the imaginary term in the denominator vanishes, or when

\[
\omega = \sqrt{\frac{T_1 + T_2 + T_3}{T_1T_2T_3}}
\]  

(62)

At this frequency

\[
-\mu_0(\omega)\beta = \frac{-\mu_0(0)\beta}{1 - (T_1 + T_2 + T_3) \left( \frac{1}{T_1} + \frac{1}{T_2} + \frac{1}{T_3} \right)}
\]  

(63)

The system will be stable if

\[
(T_1 + T_2 + T_3) \left( \frac{1}{T_1} + \frac{1}{T_2} + \frac{1}{T_3} \right) > 1 + \mu_0(0)\beta
\]  

(64)

If, for example, \( T_1 = 0.8 \times 10^{-6} \), \( T_2 = 5.0 \times 10^{-6} \), \( T_3 = 20 \times 10^{-6} \), \( \mu_0(0) = 20,000 \), and \( \beta = 0.001 \), the left-hand member of Eq. 64 is equal to 38.7, while the right-hand member is equal to 21.0 and the system is stable. Doubling the value of \( \beta \), however, would lead to a condition of instability.

The alternative method of determining system stability from the transfer-factor diagram is illustrated by Fig. 1.14, which has been obtained by substitution of the indicated circuit constants in Eq. 60 and calculation of the transfer factor for a series of arbitrarily chosen frequencies. Several points on the curve are identified by their corresponding frequencies. The dashed portion of the curve corresponds to negative frequency values. It should be noted that the curve does not enclose the point \((1,j0)\) and, as previously indicated, the system is stable.

The preceding discussion has served to present the Nyquist criterion of stability and to indicate briefly its method of application. The engineer is, however, as often concerned with the design of stable circuits or the modification of unstable circuits to secure stability, as he is with the determination of their stability. Difficulties in securing system stability, when they exist, are generally due to undesirable phase shifts at frequencies outside the normal operating range for which an amplifier or regulator system was designed. As Bode has pointed out, design changes made in degenerative systems to overcome oscillation in one portion of the frequency spectrum all too often merely result in a translation of the difficulty to a different portion of the
spectrum. It is desirable, therefore, that some consideration be given to relationships which are of assistance in the design of stable degenerative systems, and the remainder of this chapter is devoted to that end. The discussion is based largely upon several network theorems of Bode.

4. ATTENUATION AND PHASE CHARACTERISTICS

The basic problem involved in the design of absolutely stable regulator systems might be stated as that of securing the desired stabilization factor in the normal operating frequency band of the regulator and of providing sufficient attenuation of the feedback signal to reduce the transfer factor to a value less than unity before it experiences a phase shift of 180 deg due to reactive effects in the system. One method of attaining this objective is based upon certain relationships between attenuation and phase shift which apply to a general class of four-terminal passive network structures. The connection between such passive structures and degenerative systems arises from the fact that the phase and attenuation characteristics of the transfer factor are determined solely by the passive circuit parameters of the system.

It is unfortunate that no general relationships exist between the attenuation and phase characteristics of all networks. It will be apparent that such could not be the case, however, when it is noted that a section of ideal transmission line or an all-pass network structure could be added to any given network structure to form a new structure having identical attenuation characteristics with the original network, but exhibiting modified phase characteristics. There exists, however, a class of four-terminal passive structures, known as "minimum phase-shift networks," which are characterized by a unique relationship between network attenuation and phase shift when the network is connected between resistive terminations. For such networks, the phase characteristic can be calculated when the attenuation characteristic is given, and vice versa.

Any four-terminal passive network which does not contain the equivalent of an all-pass section will be a minimum phase-shift network; in particular, all ladder structures are of this type.

A detailed discussion of the more important formulas relating the attenuation and phase characteristics of minimum phase-shift networks would require far more space than is available here. Accordingly, only a few of the relationships given by Bode will be considered, following which their significance with respect to degenerative-system design will be outlined.
4.1 Phase-area Theorem. One of the simplest of the phase-attenuation theorems is expressed by

\[ \int_{-\infty}^{\infty} B \, du = \frac{\pi}{2} (A_\infty - A_0) \] (65)

where \( u = \ln \left( \frac{f}{f_0} \right) \), the quantity \( f \) representing the actual frequency and \( f_0 \) any convenient reference frequency

- \( B \) = phase shift in radians
- \( A_\infty \) = attenuation in nepers at infinite frequency
- \( A_0 \) = attenuation in nepers at zero frequency

This equation states that the total area under the network phase characteristic curve, plotted on a logarithmic frequency scale, is dependent only upon the difference between the network attenuation at infinite and zero frequencies and not upon the nature of the attenuation characteristics between these limits. The relationship of Eq. 65 is known as the "phase-area theorem," and may also be written as

\[ \int_{0}^{\infty} B \frac{d\omega}{\omega} = \frac{\pi}{2} (A_\infty - A_0) \] (66)

where \( \omega \) is the actual radian frequency involved. This theorem indicates in a general way that, if the maximum phase shift permissible in a network employed to provide a specified attenuation between zero frequency and a very high frequency is fixed at a low value, the frequency band over which the attenuation is changing must be sufficiently wide to provide the required area under the phase characteristic and that the lower the maximum permissible phase shift is set the slower must be the rate at which the network attenuation is varying. Applied to regulator networks, the theorem indicates that the phase characteristics of systems providing a high stabilization factor must be carefully controlled over a frequency band many times greater in width than that in which a large fraction of the maximum stabilization is obtained.

4.2 Weighting Function. Another relationship between network attenuation and phase shift, and one which provides more detailed information, is given by

\[ B_c = \frac{1}{\pi} \int_{-\infty}^{\infty} \frac{dA}{du} \ln \coth \frac{|u|}{2} \, du \] (67)

where \( B_c \) = phase shift in radians at the frequency \( f_c \)

\( dA/du \) = slope of the attenuation curve on a logarithmic frequency scale when the attenuation is expressed in nepers
25

\[ u = \ln \left( \frac{f}{f_c} \right), \text{ where } f \text{ represents frequency} \]

\[ \ln \coth \frac{|u|}{2} = \text{real part of the natural log of the hyperbolic cotangent of } u \]

This equation expresses the network phase shift at the frequency \( f_c \) in terms of the slope of the attenuation characteristic \( dA/du \), and a logarithmic weighting function that expresses the relative importance of attenuation slopes in various portions of the spectrum to the phase shift at a specific frequency. The form of the weighting function is shown in the curve of Fig. 1.15. Upon expressing the hyperbolic cotangent in exponential form, the weighting function may be written

\[ \ln \coth \frac{|u|}{2} = \ln \left| \frac{\omega + \omega_c}{\omega - \omega_c} \right| \quad (68) \]

A few examples may serve to illustrate the type of information which may be obtained through the application of Eq. 67. Thus it may be assumed that \( A = ku \), which is equivalent to an attenuation characteristic having a slope of \( 6k \) db/octave. Then \( dA/du = k \) and

\[ B_c = \frac{k}{\pi} \int_{-\infty}^{\infty} \ln \coth \frac{|u|}{2} \, du \quad (69) \]

which is known to be equal to \( kn/2 \). The phase shift of the network is constant, therefore, and its value is dependent only upon the constant slope of the attenuation characteristic.

Again let the network attenuation be constant at all frequencies except one, at which the attenuation characteristic experiences an abrupt change from one value to another. Such a characteristic is illustrated by the solid curve of Fig. 1.16a, the frequency of discontinuity being designated as \( f_d \). This discontinuity may be regarded as the limit approached by a characteristic such as that shown by the dashed curve of Fig. 1.16a, when the frequency interval during which the change in attenuation takes place approaches zero. As the frequency interval corresponding to the attenuation change is made very small, the value of the weighting function over this frequency range may be assumed constant. Furthermore, \( dA/du \) is zero everywhere outside this frequency interval, so that the value of \( B_c \) is essentially equal to the product of the integral of \( dA/du \) over the interval and the value of the weighting function at the frequency under consideration and is therefore given by

\[ B_c = \left( \frac{A_d}{\pi} \ln \frac{\omega + \omega_c}{\omega - \omega_c} \right) \quad (70) \]
where $A_d$ is equal to the change in attenuation at the discontinuity. This characteristic is shown in Fig. 1.16b and is identical in form with that of the weighting function itself. It indicates that a finite discontinuity in the attenuation characteristic introduces a very large phase shift at frequencies in the neighborhood of the discontinuity and smaller phase shifts at both lower and higher frequencies.

A third form of simple attenuation characteristic is that shown in Fig. 1.17a, where the attenuation is zero or constant at all frequencies below a specified frequency $f_h$ and has a constant slope $k$ for all higher frequencies. The corresponding phase characteristic is shown in Fig. 1.17b. This case is of particular interest because it roughly approximates the situation existing in several of the simpler regulator systems. At low frequencies the phase shift is practically proportional to frequency, being approximately given by

$$B_c = \frac{2kf_c}{\pi f_h} \tag{71}$$

while at high frequencies it approaches the limiting constant value

$$B_c = \frac{k\pi}{2} \tag{72}$$

The phase shift resulting from an attenuation characteristic which consists of the sum of two or more simple characteristics will be equal to the sum of the corresponding phase characteristics. To the extent then that an actual attenuation characteristic may be approximated by a sequence of linear attenuation characteristics, the phase characteristic of moderately complex networks may be approximately determined from the simulated attenuation characteristic without recourse to actual integration of Eq. 67.

4.3 Stabilization and Time Constant of Load Circuit. Mention has been made previously of the fact that regulator-system instability is generally due to undesirable phase shifts at frequencies outside the band in which high stabilization is desired. At frequencies well beyond the upper limit of the regulating band, all regulators tend to exhibit the high-frequency attenuation characteristics common to multistage resistance-coupled amplifiers, regardless of whether or not electronic amplifiers form a part of the regulator system. This is due to the fact that the high-frequency attenuation characteristics are ultimately determined solely by circuit inductances and capacitances. As an illus-
tration of this effect, consider again the regulator system of Fig. 1.13. The high-frequency attenuation characteristic of this system is determined solely by the shunt capacitances $C_1$, $C_2$, and $C_L$, which may represent elements intentionally included in the circuit or such values as correspond to parasitic circuit and tube capacitances. In either case, when the frequencies concerned are such that the gain of the system is determined by these capacitances, the value of $\mu_0(\omega)$ decreases at the rate of 6 db/octave/stage of amplification. For the 3-stage system shown, the slope of the asymptotic attenuation characteristic is equal to 18 db/octave, and by Eq. 72, an asymptotic phase shift of 270 deg results. The slope of the asymptotic attenuation characteristic of many regulator systems may be determined from an inspection of their simplified circuit schematic diagrams.

In accordance with the criterion for system stability, the transfer factor must experience an attenuation at least slightly greater than its maximum value before it exhibits a phase shift of 180 deg. Since a phase shift of this amount corresponds to a uniform rate of attenuation of 12 db/octave, the provision of some margin of stability to the regulator system indicates the desirability of attenuating the transfer factor at a rate of not more than perhaps 10 db/octave until its value is less than unity. If the slope of the asymptotic attenuation characteristic is greater than 12 db/octave, the full required attenuation of the transfer factor must be obtained before the asymptotic attenuation of the system governs the phase characteristic. It is evident then that the maximum permissible stabilization factor of a regulator system is determined by the width of the frequency band available for a prescribed rate of attenuation of the transfer factor. The more nearly constant the rate of attenuation of the transfer factor may be held, the greater will be the maximum stabilization factor possible in any given system.

Since the width of the frequency band available for attenuation of the transfer factor effectively determines the maximum permissible stabilization factor of a system, the degree of regulation that can be attained is intimately related to the time constant of the load circuit. Load circuits characterized by long time constants provide initial attenuation of the transfer factor at comparatively low frequencies and, all other factors being equal, permit utilization of higher stabilization factors than can be employed when load circuits of short time constant are involved. It is fortunate, therefore, that the most important calutron load circuits exhibit comparatively long time constants, either as an inherent characteristic of the load or by virtue of energy storage elements associated with the load circuit.
5. CONTROL OF ATTENUATION CHARACTERISTICS OF LABORATORY REGULATORS

Several methods of controlling the attenuation characteristics of laboratory regulators have been employed. One of the most effective of these depends upon the utilization of one or more auxiliary feedback loops within the regulator system proper. The circuit of Fig. 1.18 shows the application of this method to a 3-stage regulator system and represents a method first applied to calutron voltage-regulator design by W. H. Nelson. All sources of power have been omitted from the circuit schematic to reduce the system to its essential elements.

The control of the complete regulator attenuation characteristic afforded by the internal feedback loop will be evident from an examination of the expression for the gain of the first two stages of the system alone. If the d-c amplification of these two stages is denoted by $\mu_a (0)$, and if $T_1 = r_1 C_1$, $T_2 = r_2 C_2$, $T_f = (r_a + r_b) C_t$, $\beta_\infty = r_a / (r_a + r_b)$, and $(r_a + r_b) \gg r_a$, while $g_i$ denotes the transconductance of the tube V-1, the voltage amplification at any radian frequency $\omega$ is given

$$\mu_a (\omega) = \frac{\mu_a (0) \left( 1 - \frac{1}{j \omega T_f} \right)}{\left( 1 - \frac{1}{j \omega T_f} \right) (1 + r_a g_i) (1 + j \omega T_1) (1 + j \omega T_2) + \mu_a (0) \beta_\infty}$$

(73)

If the internal feedback loop were effectively opened by reducing $C_f$ to zero, the corresponding gain equation would be

$$\mu'_a (\omega) = \frac{\mu_a (0)}{(1 + r_a g_i) (1 + j \omega T_1) (1 + j \omega T_2)}$$

(74)

In the absence of internal feedback, Eq. 74 indicates a phase shift of 90 deg in the first two stages at a radian frequency equal to $1 / \sqrt{T_1 T_2}$. If the load circuit also exhibits a phase shift of approximately 90 deg at this frequency, a tendency toward system instability will exist unless the transfer factor of the complete regulator circuit has been reduced to a value less than unity.

The situation existing when Eq. 73 applies will now be considered. Let it be assumed that $C_f$ is so chosen as to reduce the internal loop feedback to a negligible value over the entire range of low frequencies of interest, but to admit full internal loop feedback at or below the frequency $1 / \sqrt{T_1 T_2}$. Then the regulator stabilization factor is unmodified at low frequencies by the presence of the internal feedback
loop. At the frequency $1/\sqrt{T_1 T_2}$, however, the phase shift due to the first two stages is reduced from 90 deg to the value

$$\theta = \tan^{-1} \left[ \frac{-(T_1 + T_2) (1 + r a g_1)}{\mu_a(0) \beta_\infty \sqrt{T_1 T_2}} \right]$$  \hspace{1cm} (75)$$

If, for example, $T_1 = T_2$, $\mu_a(0) \beta_\infty = 50$, and $(1 + r a g_1) = 2.0$, the value of $\theta$ would be only 4.6 deg instead of 90 deg, and $\mu_a(\omega)$ at this frequency would be reduced approximately 22 db below its value in the absence of internal feedback.

The improvement in both the attenuation and phase characteristics of a regulator such as that of Fig. 1.18, effected by a simple internal feedback loop, is well illustrated by the curves of Figs. 1.19a and b. Data for these curves have been calculated for the circuit of Fig. 1.18, employing the equation

$$-\mu_o(\omega) \beta = \left[ \left(1 + \frac{1}{j \omega T_f} \right) (1 + r a g_1) (1 + j \omega T_1) (1 + j \omega T_2) + \mu_a(0) \beta_\infty \right] \left(1 + j \omega T_L \right)$$  \hspace{1cm} (76)$$

for the regulator transfer factor, where $g_1$, $T_1$, $T_2$, $\beta_\infty$, and $\mu_a(0)$ retain their previously assigned significance, and where $T_L = R_L C_L$, $\beta = r_c/(r_c + r_d)$, and $\mu_o(0)$ denotes the d-c voltage amplification between the grid circuit of V-1 and the load resistance. The numerical values employed in Eq. 76 for calculation of the curves of Figs. 1.19a and b are given by

$$T_1 = T_2 = 10^{-6} \quad T_f = 3 \times 10^{-6} \quad T_L = 3 \times 10^{-4}$$

$$(1 + r a g_1) = 2.0 \quad \mu_a(0) \beta_\infty = 50 \quad \mu_o(0) \beta = 1500$$

The advantages resulting from utilization of the internal feedback loop are rendered more apparent by separately plotting the attenuation and phase characteristics of the regulator as a function of frequency rather than by reproducing the transfer-factor characteristic in conventional polar form. Figure 1.19a indicates the regulator attenuation characteristics with and without internal feedback, while Fig. 1.19b indicates the corresponding phase characteristics. This method of presentation also assists in emphasizing the comparatively great frequency bandwidth required for attenuation of the transfer factor when a regulator exhibits a high stabilization factor.
It will be noted that the regulator system would be unstable in the absence of internal feedback but would exhibit excellent stability with the internal loop operative. The phase-area theorem indicates that further improvement in the characteristics of this system is readily possible since the phase shift of the transfer factor is not maintained at a uniformly high value during attenuation of the transfer factor.

The preceding discussion of the general theory of regulator systems has indicated rather briefly some of the more important factors involved in the design of regulator systems. Several of the chapters which follow present a more detailed discussion of the design and operational problems encountered in laboratory applications of specific regulator systems.

Table 1.1 — Data for Regulator Transfer-factor Diagram (Fig. 1.14)

<table>
<thead>
<tr>
<th>Frequency, kc*</th>
<th>Value of transfer factor</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>-20.0 + j 0.00</td>
</tr>
<tr>
<td>1</td>
<td>-19.6 + j 3.19</td>
</tr>
<tr>
<td>2</td>
<td>-18.36 + j 6.07</td>
</tr>
<tr>
<td>3</td>
<td>-16.82 + j 8.54</td>
</tr>
<tr>
<td>5</td>
<td>-12.30 + j 11.32</td>
</tr>
<tr>
<td>7</td>
<td>-8.25 + j 12.10</td>
</tr>
<tr>
<td>10</td>
<td>-3.71 + j 11.27</td>
</tr>
<tr>
<td>20</td>
<td>+1.74 + j 5.98</td>
</tr>
<tr>
<td>30</td>
<td>+2.24 + j 2.95</td>
</tr>
<tr>
<td>40</td>
<td>+1.87 + j 1.48</td>
</tr>
<tr>
<td>50</td>
<td>+1.53 + j 0.875</td>
</tr>
<tr>
<td>60</td>
<td>+1.12 + j 0.380</td>
</tr>
<tr>
<td>75</td>
<td>+0.76 + j 0.114</td>
</tr>
<tr>
<td>90</td>
<td>+0.535 + j 0.000</td>
</tr>
<tr>
<td>125</td>
<td>+0.257 - j 0.065</td>
</tr>
</tbody>
</table>

*For negative frequencies, imagine the curve obtained from the data above in the axis of reals. The curve is symmetrical about this axis.
<table>
<thead>
<tr>
<th>Angular frequency ω, radians/sec</th>
<th>Transfer factor without internal feedback†</th>
<th>Transfer factor with internal feedback†</th>
</tr>
</thead>
<tbody>
<tr>
<td>10⁰</td>
<td>750/180</td>
<td>750/180</td>
</tr>
<tr>
<td>$3.33 \times 10⁰$</td>
<td>748/174.3</td>
<td>745/173</td>
</tr>
<tr>
<td>10⁵</td>
<td>717/163.3</td>
<td>714/159</td>
</tr>
<tr>
<td>$3.33 \times 10⁵$</td>
<td>550/135</td>
<td>513/121</td>
</tr>
<tr>
<td>10⁶</td>
<td>237/108.5</td>
<td>187/70.5</td>
</tr>
<tr>
<td>$3.33 \times 10⁶$</td>
<td>74.5/91.7</td>
<td>26.4/31.5</td>
</tr>
<tr>
<td>10⁷</td>
<td>24.5/78.7</td>
<td>3.26/23.7</td>
</tr>
<tr>
<td>$3.33 \times 10⁷$</td>
<td>6.72/53</td>
<td>0.40/53</td>
</tr>
<tr>
<td>10⁸</td>
<td>2.50/0.0</td>
<td>0.10/69</td>
</tr>
<tr>
<td>$3.33 \times 10⁸$</td>
<td>0.062/−36.6</td>
<td>0.045/64.3</td>
</tr>
</tbody>
</table>

*See Figs. 1.19a and b.
†In columns 2 and 3, values to the right of the slash bar are for the phase angle, expressed in degrees. The zero frequency value in both instances is, of course, equal to 750/180.
Fig. 1.1—Basic d-c voltage-regulator circuit schematic diagram.

Fig. 1.2—Simplest form of electronic voltage regulator.
Fig. 1.3—Improved form of electronic voltage regulator.

Fig. 1.4—Components of voltage regulator capable of providing high stabilization factor.
Fig. 1.5 — Equivalent circuit for regulated system.

\[ \frac{R_0 + R_d}{1 + \mu(\omega)\beta} \]

\[ \frac{\mu(\omega)E_s + E_0}{1 + \mu(\omega)\beta} \]

Fig. 1.6 — Basic d-c current-regulator circuit schematic diagram.
Fig. 1.7—Fundamental components of commonly employed a-c load-current regulator.

Fig. 1.8—Simple resonant circuit.
Fig. 1.9—Representation of the p plane. All roots of stable systems are located in the left half-plane.

Fig. 1.10—Current and voltage relationships in system employed for derivation of Nyquist criterion.
Fig. 1.11—Path of integration in the p plane.

Fig. 1.12—Vector construction for Reid analysis.
Fig. 1.13 — Simplified schematic diagram of degenerative d-c voltage regulator.

Fig. 1.14 — Regulator transfer-factor diagram (see Table 1.1 for data).
Fig. 1.15 — Form of the weighting function.
Fig. 1.16a — Attenuation characteristic with discontinuity.

Fig. 1.16b — Phase characteristic corresponding to attenuation characteristic with discontinuity.
Fig. 1.17a—Attenuation characteristic showing attenuation constant zero to certain frequency and with constant slope at frequencies beyond that point.

Fig. 1.17b—Phase characteristic corresponding to attenuation characteristic of Fig. 1.17a.
Fig. 1.18—Simplified schematic diagram of d-c voltage regulator with internal feedback.

Fig. 1.19a—Attenuation characteristics of voltage-regulator transfer factor (see Table 1.2 for data). ———, without internal feedback; ----, with internal feedback.
Fig. 1.19b—Phase characteristics of voltage-regulator transfer factor.
—–, without internal feedback; -----, with internal feedback.

REFERENCES

Chapter 2
HIGH-VOLTAGE REGULATORS

By K. MacKenzie

1. INTRODUCTION

The general theory of regulation, which forms the subject material of Chap. 1, applies to all regulators in the plant; in particular it applies to the high-voltage regulator and also the magnet regulator which will be described in a later chapter. This chapter will deal with the growth and development of the high-voltage equipment, of which the high-voltage regulator is a part. In the initial phases of the Project, the problems associated with building a regulator that would regulate the voltage with extreme accuracy were not appreciated. In fact several regulators were built before it was realized that there were theoretical limitations to the accuracy that could be attained.

The U 235 and U 238 ions differ in mass by 1 part in 80. Referring to the equation of motion of an ion in a magnetic field, \( V = \frac{e r^2 H^2}{2m} \), in which \( V \) is the accelerating potential, \( e \) is the charge on the particle, \( H \) is the magnetic field strength, and \( m \) is the mass of the particle, it is seen that a reduction of 1 part in 80 in the accelerating voltage will allow the U 238 ion to fall into the U 235 receiver slot. It was arbitrarily decided that the ions could be allowed to move one-tenth the distance between slots, which corresponds to a variation in voltage of about 1 part in 800. This was rounded off rather early to about 1 part in 1000. Referring to the equation of motion again, it is seen that the magnet current must be held constant to twice this accuracy or to 1 part in 2000. Later on in the Project the required accuracies were doubled. Regulation to 1 part in 1000 seemed like a stringent requirement at the time, but it turned out to be simpler than expected to satisfy this requirement during short intervals. Obviously, if there are no changes in line voltage and no changes in load, a regulator is not needed at all. The regulator, in short, just corrects for any changes
that appear, and if the changes are small enough a constancy of 1 part in 1000 can be realized rather easily for short times. Measuring such a voltage in order to determine the exact degree of regulation was not very easy because of spark troubles which are described later, so that initially no attempts were made to measure the voltage regulation directly. If the beam in the U 238 pocket remained steady, it was assumed that the regulator was doing its job, and on this basis it was decided at the time that 1 part in 1000 was being attained.

2. POWER SUPPLIES USED WITH CALUTRONS IN THE 37-IN. BERKELEY CYCLOTRON MAGNET

The first high-voltage supply was patterned after a Radio Corporation of America (RCA) electron microscope supply, described in the 1939 RCA Review. All the regulation was attained by means of a saturable reactor in the 110-volt power line. The output was sampled by means of a bleeder resistor across the output. This sample voltage was compared with a battery and the difference was amplified and used to saturate or unsaturate the reactor in the 110-volt line. The result was a supply which could keep the average voltage constant but which could do nothing about the ripple because it could act no faster than the 60-cycle supply frequency. Consequently the ripple voltage was reduced by brute-force filtering. This supply and regulator is shown in Fig. 2.1.

The high-voltage supply itself was built from laboratory-type transformers, chokes, and low-current rectifiers (made from Eimac transmitting tubes, type 250T). With this equipment it was found possible to regulate successfully 20,000 volts at about 10 ma on a dummy load. When applied to the calutron, however, it soon became apparent that not more than 6000 to 8000 volts could be held successfully without having excessive spark trouble. It was undesirable, of course, to have every spark trip an overload relay, and so the alternative was to build the supply in such a way that it could stand a short circuit. The laboratory transformers could not do such a thing and still have what is called “good regulation,” that is, have the voltage change very little with load. The solution was to add a diode limiter in the form of an Eimac 250T, emission-limited as much as possible, so that when the short-lived spark occurred, all the energy went into heating the plate of the 250T. This was the first use of emission-limiting to allow the high-voltage supply to ride through sparks. In Fig. 2.1 a resistor-capacitor network will be seen connected to the grid of the diode limiter. This network introduced a little local inverse feedback and reduced the ripple voltage.
The spark problem made it very evident that although a regulator might be required to supply only 20,000 volts at 10 ma, it must be extremely sturdy. Around the laboratory there were some old Western Electric 232B transmitting tubes with 20-kw plate dissipation. Although they seemed ridiculously large for the job, a regulator consisting of two stages of amplification and the water-cooled 232B tube as the final stage was built around one of these tubes. The diagram is shown in Fig. 2.2. The current was limited on short circuit by emission-limiting the water-cooled tube with of course no difficulty so far as heating the plate was concerned. This circuit is interesting because it is essentially the basic design for all subsequent regulator circuits. It is interesting to note that up to this time the problems mentioned in Chap. 1 concerning oscillation, which is due to excessive gain and phase shift around the feedback loop, had not arisen in earnest. Oscillation, which is the chief annoyance of all feedback-amplifier circuits, had arisen several times, but it was easily cured by placing a large capacitor somewhere in the circuit.

This high-voltage supply and regulator, as described above, evolved during the three months after the start of the Project. By that time the ion-beam current and the process in general had improved to such an extent that it was time to examine the behavior of the regulator in earnest. It was necessary after a spark that the voltage return to its regulated value with the maximum possible speed (especially during the time when the U 238 beam was crossing the U 235 slot) in order to reduce contamination to a minimum. It was also required that the voltage be maintained constant for long periods of time, not for just a few minutes as previously had been the case. It was also desirable that large capacitances not be placed across the output of the rectified high-voltage supply because the stored energy in these capacitors discharged on sparks, and considerable damage was done by these sparks to the electrodes inside the tank. A capacitor, of course, does not have to be placed across the output to prevent oscillation, as was discovered by trial and verified by theory. However, a capacitor has to be placed somewhere in the circuit, and the transient response of the system is much better if the capacitor is placed across the output. The capacitor, however, should be as small as possible. These considerations led to a detailed study of the nature of the inverse feedback circuit as applied to a high-voltage regulator (see Chap. 1).

The regulator just described was sufficient for the needs of the calutron which had been in operation up to this time. However, for the next step in the development, it was obvious that a much larger calutron should be built and the corresponding power supply would have to deliver something like 20 to 30 kw at about 1 amp. This meant that in
addition to building a much larger supply, a regulator would also have to be built to regulate this current at about 1 amp. The supply circuit itself is shown in Fig. 2.3, and for the first time 3-phase mains were used. Standard single-phase pole transformers, about 11,000 to 220 volts were used as the plate-rectifier transformers and connected either delta-Y, delta-delta, or Y-delta, depending on the voltage required at the particular time. Full-wave rectification was used employing type-872 tubes since higher inverse tubes could not be obtained on short notice. Since these rectifier tubes will stand only 10 kv inverse, two of these supplies were connected in series in order to satisfy the requirement of 20 kv. Some components were extremely hard to obtain on short notice, in particular the filament transformers.

These filament transformers are of some historical interest in that they were all manufactured at the Radiation Laboratory from small rolls of iron baling wire. The roll of baling wire was in the form of a ring with the inner diameter about 8 in. and the outer diameter 12 in. The ring of wire was wrapped with tape to hold the wires together and also to supply insulation for the primary which was wound over the tape. The primary winding was therefore in the form of a toroid in the same manner as a variac. The secondary coil was wound in the form of a ring, about 12 in. in diameter, and consisted of about 10 turns, taped together. This secondary ring was wound so that it went through the 8-in.-diameter opening in the roll of baling wire. The spacing between primary and secondary was therefore about 4 in. in air, which was sufficient to stand 35 kv. About 35 of these transformers were constructed before delivery of commercial transformers was obtained.

Two 10-kv supplies of this nature in series should produce 20-kv output. However, because of the numerous transients in the system due to discharge or breakdown across the load, there were frequent arc-backs in the mercury rectifiers, and consequently the voltage limit was closer to 18 kv than to 20. In order to raise the limit, two more supplies were added making a total of four 10-kv supplies in series. A maximum of 35 kv was reached on some occasions. The regulator was of the type shown in Fig. 2.2. As will be noticed from Figs. 2.1 to 2.3, the negative terminals of all the supplies described so far have been grounded with the positive ends at high voltage. All the calutrons were operated so that the ion source was insulated from ground and had to run at high positive voltage. So far as the regulator circuit is concerned, all voltages worked out in such a way that the regulator amplifier could be direct-coupled to the grid of the 232B tube, although the regulator itself has to float at high voltage (see Fig. 2.2). This is a slight inconvenience because insulating transformers
are required to supply the power for the regulator and because the control knobs have to be driven through insulating shafts.

At about this time it was discovered that a damaging discharge which occurred in the calutron was due to the fact that the ion source was positive with respect to ground. Accordingly drastic changes were made. The ion source was placed at ground potential and an inner liner at high negative potential was placed in the rest of the space. This also meant, of course, reversing the power supply. As can be seen from Fig. 2.4, the power supply then floats completely isolated from ground and, in effect, floats on top of the regulator tube. While this appears to be an advantage since the regulator is not at high potential, it turns out to be quite a disadvantage on surges or sparks which occur inside the calutron. As can be seen from an inspection of the diagram, when the high-voltage end is short circuited to ground, the emission limit is reached in the regulator tube, causing the full power-supply voltage to appear across the regulator tube. This means that for every spark inside the calutron, the entire power supply is subjected to a potential of 20 or 30 kv, or whatever the output voltage may be at the time. All parts of the power supply then must be capable of withstanding a sudden 30-kv transient. On closer examination, the actual surge in some parts of the power supply is greater than the actual output voltage; this is due to inductances of leads and leakage inductance in the transformer windings. The actual voltage excursion in some types of power supplied can be as high as twice the surge voltage across the power supply in the form of short circuit to ground.

Accordingly, work was seriously started on a new type of regulator which could be inserted in the negative high-voltage lead. It is at once obvious that a direct-coupled regulator cannot be used because there is no way in which the output voltage can be sampled and fed back through an amplifier to the grid of the water-cooled tube. The obvious solution is a regulator using an r-f link. The voltage is still sampled at ground potential and compared with the standard voltage. It is then amplified and used to modulate an r-f oscillator. The r-f energy is transferred to the 232B, which is now at high voltage, by means of a low-capacitance vacuum capacitor. The r-f signal is rectified by a diode and put on the grid of the regulating tube. Such a system is considerably more complicated than a direct-coupled regulator but in principle there should be nothing wrong with it. However, it turned out that the problems mentioned in Chap. 1 arose in earnest. The extra components introduced large phase shifts between the input and output signals.

The first system tried involved frequency modulation. The input signal, which is proportional to the voltage across the power supply,
was used to modulate the frequency of an oscillator. This change in frequency was fed into a discriminator in the regulator at high voltage and the resulting d-c signal from the discriminator was amplified and applied to the grid of the regulator tube. A block diagram is shown in Fig. 2.5. This system would oscillate almost all the time. When sufficient capacitance was placed across the output of the supply, it was possible to stop the oscillation, but the capacitance was excessive (several microfarads), and the discharge of this energy through the calutron did considerable damage. These oscillations can be explained very simply, as has been mentioned in Chap. 1, by the fact that when the input signal undergoes a phase change of 180 deg in its round trip through the amplifier, it will arrive back at the input terminals in phase with the original signal, and if the gain is still greater than 1, the system will oscillate. An f-m device is inherently a little more complicated than an a-m device since an extra stage or so is required to produce the same power level. It is therefore to be expected that more phase shift will be suffered as the signal goes from input to output. The f-m system was therefore abandoned and the simplest possible a-m system was installed.

The a-m system is shown in Fig. 2.6 and consisted of a high-μ input tube (1851) which modulates directly an 804 oscillator tube. This energy is fed through a vacuum capacitor to a 2X2 rectifier tube, which rectifies the r-f signal and applies the voltage to the grid of a 304TL tube (which has a 1-kw rating). The 304TL in turn applies the voltage to the grid of the 232B. In this system of regulation it was considered essential to use a large tube such as a 304TL to drive the 232B. The reason for this is that the 232B can supply 1 amp with about an 8-kv drop at zero bias, and therefore a lot of power and regulating range is wasted if it is operated only in the negative grid region. The 304TL, however, is a power tube and is capable of driving the 232B in the positive region. A few hundred volts positive will produce a current of 1 amp at a voltage drop of 2 kv across the tube. This feature was retained in subsequent regulators until delivery of low-μ transmitting tubes allowed operation in the negative grid region. This regulator would reduce any change which appeared in the rectifier voltage by a factor of about 200. This does not sound very high, but since most changes that occur are only a small fraction (say one-tenth) of the total voltage, this really represents a stability of about 1 part in 2000, which was more than adequate for the purpose. Nevertheless, there were shortcomings such as a rather large output capacitor and also long-period drifts which were partly due to the battery standard. Batteries, as it appeared later, are not quite constant enough for the purpose unless they are temperature-controlled and also operated under conditions of no current drain.
Another feature of a supply such as this is that if desired several regulators can be connected to one common power supply. Offhand it would seem quite desirable to have one central supply and a large number of regulators that would control the voltage on several individual calutron units. With this in mind, the laboratory looked into the problem of converting the 25,000-volt 100-amp constant-current supply (for the 184-in. cyclotron oscillator) into a standard 3-phase rectifier. The idea was abandoned when it became evident that smaller and higher voltage supplies could be obtained much sooner.

The r-f regulators used on the Project really owe their development to this desire to operate several tanks from a common power supply, each with its own regulator tube. The d-c regulator at ground potential, with its floating power supply and spark problems, was still a better regulator, and if it had not been for the desire to operate several tanks from one supply, it is doubtful whether or not the r-f regulator would have been developed.

3. HIGH-VOLTAGE SUPPLIES FOR THE CALUTRONS IN THE 184-IN. CYCLOTRON MAGNET

The first r-f regulator and the 40-kv supply described previously were used on the calutron in the 37-in. magnet at the University of California. The next step in the general development was a further increase in power when the 184-in. cyclotron magnet became available for use. A main power supply of something like 40 kv at 5 amp was contemplated, which might serve two or three calutron tanks at once. The ability of such equipment to withstand short circuits in the calutron was of prime importance, and so at this point it is well to consider the high-voltage surge problem that was ever-present throughout the development of the high-voltage rectifiers and regulators. At that time nothing specific had been done about the surge problem, except to fix things when they broke down or make them bigger and heavier. However, as the power supply increased in size the effects of these rapid short circuits were sometimes quite disastrous.

In the 37-in. installation the power-supply leads carrying the high voltage to the tank were placed overhead on insulators, much like outdoor transmission lines, and the ground return was placed on the floor. The result was that on a spark inside the calutron all circuits in between the overhead lines and the ground return picked up high induced voltages which sometimes caused trouble, such as an arc across relay contacts, which would hold the main contactor closed, or surge voltages in control circuits, which would burn out tubes. One serious problem was the protection of meters, and in particular the meter which read
the d-c current to the calutron. Meter protection was originally obtained by means of a resistor in series with the meter and a VR tube across the combination. In order to minimize the surge current through all circuits, the capacitor across the output of the regulated supply was placed directly at the calutron tank. In this way the short-circuit current from the capacitor discharged directly into the calutron and passed through no meter circuits. This capacitor is the capacitor mentioned earlier for stabilizing the regulator and preventing oscillation.

When the high-voltage supply for the large 184-in. unit was installed, special attention was paid to the surge problem. This supply was much larger than any other supply and was finally designed to produce 50 kv at 6 amp. Three 100-kva transformers were used with their secondaries connected in delta. The rectifier tubes were General Electric 214 tubes which were at that time marked obsolete, but General Electric still had a few in stock (see Fig. 2.7). Filament transformers were again homemade, this time from much larger rolls of iron baling wire, each roll weighing about 100 lb. The poor quality of iron in these transformers and the large secondary spacing made them very suitable for heating tungsten filaments since the short-circuit current was not much higher than the operating current. The efficiency, of course, was low. The regulator tube employed was a Western Electric 342A. This tube is very similar to the 232B since the grid must be driven in the positive region. It was chosen rather than a low-μ tube simply because it was more readily available. It also served the double purpose of regulating and emission-limiting on a short circuit. With the correct filament temperature the emission-limit current was about 30 per cent greater than the normal running current, but in order to maintain a constant temperature it was necessary to use a Sola transformer in the primary line. The Sola transformer is very necessary since a 6 per cent increase in line voltage will double the emission limit, and twice the emission limit on sparks would cause considerable trouble in the calutron in addition to excessive plate dissipation in the 342A.

The rest of the r-f regulator for the 184-in. calutron is similar to the one described previously. A schematic diagram is shown in Fig. 2.8.

The surge problem was considered as carefully as possible. The high-voltage lead to the tank was mounted on insulators in a rectangular screen duct with about 8 in. spacing between the center lead and the duct walls. Figure 2.9 shows the 184-in. magnet with one of these ducts installed overhead. Figure 2.10 shows a section of duct framework minus the screening. The high-voltage lead itself consisted of a conduit for the purpose of carrying the numerous leads from the control desk to the calutron. The leads to the receiver pock-
ets, in other words the beam meter leads (all at high voltage), have to be thoroughly shielded from the high-voltage surges and are therefore placed inside the high-voltage conduit. All meters, such as beam meters, which were at high voltage were placed inside shielded boxes upon insulators. A picture of these meter boxes for a later installation is given in Fig. 2.22. Many of the breakdowns in the rectifier were due to the fact that the transformers were initially connected directly to the line (by a contactor) without a step-start device. There were also other breakdowns due to gas kicks in the 342A tube. Such a gas kick, of course, short circuited the whole rectifier without benefit of emission limit, resulting in a tripping of the overload relays.

The return time of the new regulator was also studied in much more detail. The voltage was caused to rise rapidly after a spark, overshoot the previous voltage, and then drop down slowly to the correct voltage. The U 238 beam, which, of course, passed very rapidly over the U 235 pocket when the high voltage discharged, will also cross the U 235 pocket when the voltage is rising again. However, when the voltage rises very steeply and the U 238 beam passes very rapidly over the U 235 pocket, the contamination is small. This was achieved by adding a small capacitor across the lower end of the high-voltage bleeder across which the sample voltage is taken. This output resistor had formerly been made aperiodic in its behavior by paralleling it with a capacitor divider as well as a voltage divider so that the sample voltage could follow quick changes in the output.

A further addition was a device known as a "beam scanner." In order to examine certain features of the beam, it was necessary to modulate the high voltage, that is, to add a certain a-c component in series with the high voltage. This swept the beam back and forth across the pocket, and with the sweeping voltage applied to the horizontal sweep of an oscilloscope, the shape of the beam could be studied. This scanning voltage was originally added in series with the high-voltage supply in the form of a transformer. Unfortunately the direct current of about 1 amp flowing through the transformer saturated the iron and required that a very large transformer be used. This difficulty was easily eliminated simply by adding a very small scanning voltage in series with the standard voltage in the regulator. The regulator will cause the high voltage to follow faithfully any change in the regulator standard. When viewed in this manner the regulator is seen to be nothing more than a very good feedback amplifier. The standard voltage is amplified by a factor known as $1/\beta$ (see Chap. 1), which is the ratio of the two sections a and b of the output voltage divider (see Fig. 2.6), or is equal to the ratio of output voltage to standard voltage. If a small a-c scanning voltage is connected in series with the stand-
ard, the regulator reproduces this voltage (multiplied by $1/\beta$) in the power-supply output.

In an attempt to solve some of the many problems which were associated with these regulators, a new regulator was designed (to work with the same power supply) using low-$\mu$ 891 triodes, which, of course, could operate entirely in the negative grid region. The chief feature of this regulator was that stability against oscillation was achieved by adding capacitance to the grid of the 891. This theoretically has a certain advantage since in a triode the effective capacitance between grid and filament is always increased many times by the Miller effect, and therefore considerable capacitance is always present between grid and filament, even if the stabilizing capacitance is on the output. Referring to Chap. 1, it is evident that capacitance in two places in a regulator circuit is very undesirable. This new circuit lumped the stabilizing capacitance from grid to filament and used as little capacitance as possible across the output.

This regulator, when tried on the calutron tank, had some very desirable features in that after a spark the voltage was returned with extreme rapidity. It is evident that the rapidity with which the voltage can return will depend on the output capacitance that must be charged by the emission-limit current. The energy which goes into the calutron tank on a discharge is very much reduced when the capacitance is small. However, the number of sparks per second was greatly increased. Such a regulator seemed very promising. Unfortunately, very shortly afterward the chance of using it appeared to be very small.

For plant use the high-voltage lead had to be x-ray cable. The bulky screen duct, about 15 in. square, used to house the high-voltage conduit would find very little room in a production plant. However, x-ray cable has a very high capacitance, so no matter how hard one tried, the capacitance across the output of the regulator would be quite high. Also, it was found necessary to run heater elements in the calutron at high voltage for heating slits, and these heaters had to be supplied by heater transformers also at high voltage. They too added capacitance to the output of the rectifier and the net result was that the output capacitance became so large that any capacitance added to the grid of the 891 would render the regulator unstable. However, this regulator served to emphasize the need for keeping the output capacitance as low as possible. A photograph of this regulator at a later date, with the 891 replaced by an 893, is shown in Fig. 2.11.

The next step in the development of the high-voltage equipment was the operation of two tanks on one supply. The 50-kv 6-amp supply was connected to two identical regulator tubes and circuits of the type shown in Fig. 2.7. Two automatic high-voltage disconnect switches,
interlocked with the safety doors, were installed (see Fig. 2.12). These switches were rather large in construction, consisting of a long arm which would break with approximately a 3-ft gap (they had to be reset with an insulated pole). It was demonstrated that they would open under a dead short circuit on the rectifier, but inasmuch as the 214 rectifier tubes themselves would emission-limit at 10 amp this was not a very good test of such an installation on a large scale. Actually these switches were not required to open under load. When one tank was shut down, the 342A regulator tube was first biased to cutoff, and then the disconnect was opened.

The power supply for one of the tanks is shown in Fig. 2.13. Above the supply and in the background is the cage enclosing the high-voltage switch. It is likely that these high-voltage switches would have been used for quite some time if it had not been for the fact that the hot-source calutron was given another try in one of the tanks. This meant that the polarity of the high-voltage supply would be positive with respect to ground. At this time there were two calutrons in operation, one known as C1 and the other known as D1. The C1 calutron operated with a high negative voltage and the D1 was to operate with high positive voltage. This meant, of course, that the same supply could not be used, and another supply was built to accommodate the hot source in D1. This put an end to the high-voltage switch program. The high-voltage switch, however, had been in use long enough to convince most people that such an installation would probably be impractical if extended to more than a few units at a time. The danger of something going wrong with the switch, such as failure to extinguish the arc on a heavy short circuit, was always present. This time there was no necessity for building a regulator to operate from a supply with the negative lead grounded since the r-f coupled regulator obviously will work from either type of supply with equal ease.

After a very short time it was decided to switch back to a cold-source operation and develop a prototype unit for plant use. This program had hardly started before the situation was complicated by the addition of another high-voltage supply known as the "accel supply." This was another supply in series with the regular accelerating voltage, so that the total supply voltage, instead of being 35 kv to ground, was nearer 50 kv to ground. The ions were accelerated to 50 kv and then they were decelerated, all in a space of an inch or two down to 35 kv for the rest of the trip. Consequently any variations in the 15-kv accel voltage canceled out completely, and the only supply that required the careful regulation was the original supply that now became known as the "decel supply." This supply was first installed using a small 15-kv transformer and 8020 rectifier tubes, supplying about
amp. Since the ions were accelerated and decelerated in going through the accel slits, the net current drain on the accel supply was small, usually less than one-fourth of the decel current.

A block diagram of this installation is shown in Fig. 2.14. An idea of the space required for the accel and decel supplies can be gained from Fig. 2.15, which shows the installation for D1. The rectangular house in the foreground is the control room, and the enclosure to the left surrounds the high-voltage equipment. The calutron unit used with this high-voltage equipment was a model (as near as possible) of the expected plant unit, except that it operated on its side. It did not seem feasible at this time to make any prototype of electrical equipment since the plans of the electrical supplier (General Electric) were unknown.

4. ELECTRICAL EQUIPMENT FOR THE VERTICAL CALUTRON MAGNET XA

Shortly after the design of this production calutron was started, it was decided to go ahead with an actual prototype of the plant equipment. This meant the construction of another magnet in which the calutron could operate on a vertical position. Along with this program it was also decided to attempt a plant-type construction for the electrical equipment. All thoughts regarding such equipment lay in the hands of the Radiation Laboratory since G.E. had not formulated any definite plans at the time.

By this time there were many studies of various building layouts for the Alpha I plant made by Stone and Webster, and it became apparent that a good fraction of the building space would be required for the electrical equipment. In order to reduce this space to a minimum it became necessary to study the layout of the high-voltage equipment and the heater- and arc-supply equipment in some detail. The basic idea was to package as much of the equipment as possible in a cubicle, the cubicle being long and narrow so that many cubicles could be stacked side by side.

One of the largest pieces of equipment in the high-voltage supply is the high-voltage transformer, and it was not desirable for the laboratory to keep this transformer in the cubicle. Consequently the layout made by the laboratory involved having a long row of transformers outside the building. This could be changed and the transformers put inside provided the transformers were filled with pyranol. The transformers had to be placed outside, of course, because of the building code (regarding fire hazard) which requires that all transformers filled with oil be placed in a concrete vault or outside the building.
The rest of the high-voltage equipment, consisting of the rectifier, the accel supply, the regulator, and the filter, were all placed in a cubicle measuring approximately 3 by 10 by 8 ft high. Several views of the cubicle in various stages of construction are shown in Figs. 2.16 to 2.19. The heater power supply and arc power supply were placed in a separate rack, and since these supplies were at ground potential they were placed in fairly conventional relay racks. Two such cubicles were built, as it was planned to operate two calutrons in the new XA vertical magnet. A third cubicle was planned as a standby unit and a switching arrangement was designed so that in case of breakdown there would be a serviceable third unit ready to take its place. As quite often happens with such plans, the third unit was never built.

This high-voltage supply was not quite so successful as some of the previous supplies, chiefly because of the desire to save space. The rectifier tubes, which previously had been G. E. type-214 water-cooled tubes with 20-kw plate dissipation, were quite bulky and were replaced with Machlett-type tubes with 50-kv inverse rating and approximately 1/2 kw plate dissipation. The anode on these tubes is external and requires forced-air cooling. These tubes were used in a full-wave 3-phase supply, which meant that the inverse on the tubes was limited to 50 kv. Both d-c and a-c regulators were built to control the 893 regulator tube. The 893 tube was chosen because G.E. had indicated that it was the tube that would be used in its design. Both these regulators operated quite successfully, but both had the feature of being at ground potential and the power supply essentially floated above ground. The Machlett tubes unfortunately did not stand up very well against the resulting surges to ground which are an inherent characteristic of this sort of floating power supply when the output is shorted. It is not known whether the Machlett tubes would have functioned successfully even in the absence of the surges since these tubes were rejects from the armed forces. The Machletts were replaced with KC4 tubes with 150-kv inverse and 1-kv dissipation. Because of the surge trouble, G.E. brought out some surge-measuring equipment, consisting of surge-crest ammeters of the magnetization-type and surge-crest voltmeters of the Lichtenberg figure type. Surge voltages were recorded, but they were not much in excess of the normal surge voltage caused by a short to ground. For this reason it is probable that the Machlett tubes failed simply because they were rejects.

In these installations a great effort was made to keep the output capacitance as low as possible. By careful design of the regulator the output capacitance could be made as low as 0.01 mmf without oscillation, but it was usually operated with about 0.025 mmf to give some
margin. The smaller the output capacitor, of course, the greater the speed of return after a spark. However, it was noticed that a large number of sparks would occur with a small capacitor, whereas only a few sparks would occur with a larger capacitor. Once a discharge started it continued until that particular area in the calutron cleaned up, and it appeared that a fixed amount of energy was required to accomplish this, i.e., the number of sparks times the energy per spark was roughly a constant. It was really never settled conclusively that it was better to have a small capacitor than a large capacitor so far as damage was concerned. However, it was felt by all that it was reasonable to keep the capacitance as low as possible.

The regulator-amplifier unit was designed to be replaceable, that is, regulators could be exchanged between the two cubicles and replacement regulators could be installed merely by pulling out plugs. The r-f regulator used more stages than the d-c regulator, these stages being of the type used in video amplifiers for television, with a small phase shift per stage. The cathode of the first tube was connected to the bleeder, with results to be explained later. The d-c regulator used three stages including the 893 and was patterned after some of the development studies which G.E. was conducting. It was built by the laboratory chiefly to aid in solving problems in the proposed G.E. regulator. When first installed, the r-f regulator had a much faster return time because of the fact that the cathode of the first tube was connected to the bleeder (see Fig. 2.20). On a spark the cathode would become positive, and since no grid current was drawn, the grid being negative with respect to the cathode, the cathode would change in potential by the full voltage picked out at the bleeder. This was around 400 volts; consequently the cathode would go 400 volts positive. When the spark cleared up the cathode immediately jumped back with nothing to slow it up until it reached its operating potential again and the regulator started to regulate.

The d-c regulator on the other hand returned very slowly. Had it not been for the faster return of the r-f regulator, which was very evident, it would probably have taken some time to locate the trouble in the d-c regulator, which was caused by the fact that the grid was connected to the bleeder. On a spark the grid went positive and would have gone 400 volts positive except for the fact that grid current would be drawn, which limited the potential change of the grid to a few volts. When the spark cleared up the grid would start to go negative and immediately the regulator would find itself in the regulating range and slow up the return time. This was cured by a network, shown in Fig. 2.21, which was later incorporated in the G.E. production regulator.
It consists of the network shown in the dotted circle in Fig. 2.21. The constants are \( R_1 = R_2 \) and \( R_3/R_2 = C_2/2C_3 \). On a spark the grid starts to go positive and gets no farther than a few volts because of the grid current. The point A takes on a value halfway between the grid and ground since \( R_1 = R_2 \). When the spark is extinguished and the voltage returns to normal, the point A rises rapidly in voltage (in the negative direction) by an amount equal to one-half the standard voltage since \( R_3/R_2 = C_2/2C_3 \). The time constant is therefore determined by \( R_1C_1 \), and since \( R_1 \) is around 250,000 ohms and \( C_1 \) is about 50 to 100 mmf, the time constant is about 10 to 20 \( \mu \)sec.

Once the XA installation was operating successfully, it was necessary to study the surge situation from the operating standpoint as well as the electrical standpoint. Frequently the calutron would break down continuously, causing what was called a "flurry." Quite often the flurry would continue for some time, and in such cases the strain on the electrical equipment was considered to be excessive, so a time-delay overload relay was installed. This was set to trip after approximately 1 sec on a 50 per cent overload. In normal operation this would correspond to the emission-limit current, which usually exceeded the running current by about 50 per cent. After the relay tripped, it was necessary for the operator to turn the supply back on. Eventually automatic recycling was installed and worked so well that it was considered desirable to put such a device in the plant. However, the G.E. cubicle construction had progressed to such an extent that installation of a recycling device would have delayed delivery dates a little, and so it was not installed.

These surges also hampered operation in another way. The 35-kv supply was short-circuited to ground, often a 1000 times per second, resulting occasionally in major damage, but usually creating minor annoyances such as incorrect meter readings and poor temperature regulation. There were quite a number of delicate circuits involved, such as the beam meter circuits, which measure currents in milliamperes, and thermocouple circuits, which give readings in millivolts. All meters, wherever possible, were employed as millivoltmeters, measuring the potential drop across a noninductive shunt. An r-f choke was placed in series with the meter, and the meter was then heavily by-passed by a capacitor. More complex schemes using protective glow tubes were tried, but since the simple circuit described above could give adequate protection, all other circuits were eventually discarded. In the G.E. cubicle a piece of thyrite was connected across the network. All these meters were placed in shielded boxes (when at high voltage) with wires in front of the meter faces as can be seen from Fig. 2.22. The same meter box is shown in Fig. 2.18, but the wire cover has been removed.
The maximum amount of care was given to the thermocouple leads. If the temperature of the charge should increase, because of pickup from high-voltage surges, the spark problem would get worse and result in a runaway condition. The thermocouple leads were therefore continuously shielded all the way into the calutron tank since for some time high-voltage surge pickup was suspected of causing burned-out heater elements. The heaters burned out very probably because of a combination of various factors. The heaters ran very hot or very near to the point where they would fail anyway, so that any small surge would cause them to break down. All heater leads were concentrically shielded up to the heater itself as much as possible. In some cases it was found impossible to make the concentric shielding extend the whole distance. In such cases the heater lead and the ground return were kept as close together as possible. All heater terminals were bypassed by pyranol capacitors.

One very common source of trouble was due to the use of terminal strips in control wiring. One of the tie points is usually reserved for the ground wire and the ground wire is relied upon to carry return currents rather than the cable shield. However, to eliminate surges it is imperative that the surge current be carried by the external shields. The shielding had to be complete, that is, all joints were complete joints with no cracks across the direction of current flow. Cracks in the direction of the current flow could be tolerated, of course. Eventually the surge voltage across the heater leads was reduced to a small fraction of the voltage applied to the heaters themselves.

In the XA installation all high-voltage connections were made by means of x-ray cable in an attempt to duplicate the conditions in the plant. Considerable effort was spent in obtaining correct cable terminations. If the x-ray cable is imperfectly terminated, local stresses develop, causing a failure of the cable. These cable terminations are made by opening the outer ground conductor of the x-ray cable, putting in extra insulation to reduce the stress where the outer conductor finally stops, and then taping everything up solidly. The essential point is to exclude all air in which corona discharge can take place. Figure 2.23 shows typical cable terminations. Everything possible was done to reduce the voltage stress at the end of the cable by terminating the x-ray cable at the cubicle in something approaching its characteristic impedance.

If the termination is correct, the end of the cable at the cubicle suffers the same surge as the end at the calutron. If no termination is provided, the surge at the cubicle may be as high as twice the surge at the tank.
When terminated in its characteristic impedance, the only voltage seen by the cable is the surge voltage. This, of course, is almost impossible to do since it is very difficult to get a terminating impedance close enough to the cable to act as the terminating impedance. In other words, the cable leads must be separated to stand the voltage, resulting in much higher impedance for a short length, so that the reflections are to some extent inevitable. The termination in its characteristic impedance cannot be done directly but has to be done through a capacitor. As shown in Fig. 2.24 the cable is terminated with a 25-ohm resistor in series with the capacitor which stabilizes the regulator against oscillation. This 25-ohm resistor, even though it carries an average current of approximately only 1 amp, must nevertheless dissipate considerable power during flurries. This is due to the fact that the 0.01-mmf capacitor discharges through this 25-ohm resistor about one to three thousand times a second, depending on the emission-limit setting. The actual power dissipated by the resistor under these conditions could be as high as 15 kw.

This termination was also useful in another way. The voltage which the regulator tube has to stand while emission-limiting on a short circuit is the surge voltage at the cubicle. It is very important to keep this voltage as small as possible because if the tube voltage exceeds 50 or 60 kv the danger of the tube breaking down in the form of a gas kick is very great. By terminating the cable with a resistor and capacitor, the danger of the voltage doubling is reduced tremendously. With the 25-ohm resistor and 0.01-mmf capacitor, the surge voltage measured at the tube exceeded the actual surge voltage at the calutron tank by a very small amount.

5. XA OPERATION WITH THE G.E. TEST CUBICLE

Sometime after the installation of this XA cubicle, the first G.E. cubicle was brought out to the laboratory for its initial tests and was used to operate the XA tanks. Eventually two G.E. cubicles were brought out; these cubicles may be seen in Fig. 2.25. Considerable time was spent switching back and forth between laboratory cubicles and G.E. cubicles since there were evidently some things in the G.E. cubicle that required considerable changing. The chief thing which became apparent was that there was some new and very violent surge trouble in the G.E. cubicle. This was occasioned by the fact that the rectifier circuit was very different from the circuit employed in the laboratory cubicle. The laboratory cubicle circuit was a 3-phase full-wave power supply, whereas the rectifier used in the G.E. cubicle was a 6-phase half-wave rectifier with interphase transformer, or, in
other words, the circuit known as the "double-Y" with interphase reactor. It is not at once obvious that one of these rectifiers should be much better than the other for this special application.

Closer examination of the two circuits (see Fig. 2.26), however, shows that in a 3-phase full-wave supply the surge can be applied to the transformer only through three conducting rectifier tubes. The rate at which the capacitance of the transformer windings to ground can be charged is therefore limited by the emission limit of the rectifier tubes, which are of the high-vacuum type. In series with this capacitance, leakage inductances are also found in the transformer windings. The inductance tends to keep the current flowing and eventually, as is well known, the voltage across the effective winding to ground capacitance would be double the surge voltage. This effect does not take place, however, because once the winding voltage has become equal to the impressed surge voltage the three other rectifier tubes start conducting and short-circuit most of the extra voltage that would normally appear.

In the double-Y circuit the surge voltage is transmitted to the windings through the tubes as before and also through the direct connection to the interphase reactor. This connection is really made through the output filter capacitor, but this is a dead short circuit so far as a surge is concerned. There is no emission limit to soften this sudden jolt. The center tap of the interphase reactor is normally near ground potential because the drop across the 893 is usually only a few kilovolts. After the surge it is suddenly about 40 kv positive with respect to ground. Current flowing through the interphase-reactor windings charges the capacitance of the transformer windings to ground. Again it is the situation where a capacitor is being charged through an inductance, so the maximum voltage on the capacitor will reach twice the surge voltage. This time there are no rectifier tubes to conduct and reduce the surge since it is observed (see Fig. 2.26) that all the cathodes are surging positive with respect to the plates. The total surge is, therefore, about 80 kv. For certain cathodes it is found that the normal 60-cycle transformer voltage of about 50 kv will be in phase with the surge, making a total of around 130 kv to ground. Voltages of this magnitude were actually measured, resulting sometimes in large sparks 6 to 10 in. long, which of course tripped relays and shut down the supply.

The solution was finally found by W. H. Nelson, who suggested installing a diode limiter in the output of the rectifier to emission-limit on sparks (see Fig. 2.27). The regulator tube then only regulates and does not emission-limit. This diode limiter, known as the "562 tube," is a new tube developed by G.E. for this purpose and has an 893 anode
and a 214 filament. This diode limiter thoroughly cured the surge problem except when gas kicks occurred in the diode limiter, in which case the 893 would emission-limit and the rectifier would go through the same excursion as before. This was cured by a spark gap across the 893 in order that the voltage would rise to a certain value, at which point the spark gap would fire and all relays would be tripped before the voltage could reach a dangerous value. These surges started G.E. on a surge-measuring program in which they eventually used the large G.E. lightning oscillograph usually kept at Pittsfield. The addition of the diode limiter cured most of the troubles on XA, which then became essentially a training unit for operators for the plant.

6. FOUR-ARC ALPHA II HOT-SOURCE PROGRAM

In the meantime, the Radiation Laboratory had again tried the hot-source calutron using the horizontal tanks R1 and D1. This was followed, in the same manner as in the Alpha I calutron development, by a prototype unit in the vertical XA magnet. With more knowledge of how to dispose of the bombarding electrons by means of the dumping system described in the National Nuclear Energy Series, Division I, Volume 6, this development had succeeded fairly well and allowed the source to be extended so that four arcs could be run in parallel. This meant doubling the requirements on the power supply. Actually, the arcs had become larger and the requirements on the power supply were a little more than doubled. For this purpose, then, it was necessary to develop another supply capable of giving about 2 amp with the voltage positive with respect to ground.

The accel supply, of course, was made simpler by this change since it was now at ground potential and was simply an ordinary supply with negative voltage high above ground. The regulators for the decel supply were much the same as the d-c regulators in XA. The tube used was the 893 with the 562 as the diode limiter. This was not always the case, however, as the 893 was also used as the limiter on many occasions. The 893 does not supply 2 amp very easily, and therefore for this supply two 893’s were placed in parallel. The accel supply had also grown in proportion and was supplied by G.E. type-214 water-cooled tubes in the same manner as the decel supply. Photographs of various rectifier arrangements used in this program can be seen in Figs. 2.28 to 2.30. The surge problems are intensified in the hot-source design since the arc supply and the heater supply are all at high voltage. These supplies actually take up a fair amount of space, for it is necessary to shield them from surges, which means the supplies have to be in shielded cages above ground. The receiver situation is simpli-
fied, though, in that the receivers are now at ground potential so that the receiver meters can be right on the panel of the control desk or control panel.

When it became apparent that the design for the cold source was pretty well frozen and that little could be achieved by trying to make any major changes, the effort of the laboratory was turned toward developing a hot-source calutron and the electrical circuits to go with it. By this time the design of the electrical components was pretty straightforward and actually the G.E. design for this Alpha II equipment proceeded fairly independently of anything the laboratory did. Consequently no attempt was made on the part of the laboratory to simulate a production cubicle; the emphasis was rather on building supplies which were as flexible as possible and which could be changed to meet any demand on a day's notice. As a result, the supplies were spread over quite an area. Figures 2.31 to 2.33 are photographs of various parts of the electrical equipment for the hot-source calutrons which were used in the XA magnet. This installation was known as "XAH" (H for hot source). The calutron itself was as exact a replica as possible of the unit that would be installed in the plant.

In order to handle $2\frac{1}{2}$ amp at 35 kv, it became obvious that it would be desirable to improve the regulator tube if possible and also the diode limiter tube. A joint study was made with G.E., and it was found that the 893 could emission-limit better than the 562 tube. This was due to the fact that the ratio of the diameters of the grid structure and anode came very close to the optimum ratio of 2.78 for minimum field strength at the grid. The 562 has a very small filament structure and consequently a much higher field gradient for the same applied voltage. Therefore in the Alpha II cubicle the regulator tube and emission-limiting tube are combined in the 893 tube; however, two tubes are used in parallel. This introduces some trouble in balancing the emission limit between the two tubes, which is a difficult operation because, as explained earlier, a 6 per cent change in the line voltage will double the emission from the filament.

At this point the laboratory essentially dropped the development of high-voltage equipment and depended on G.E. to furnish adequate supplies for the plant. Numerous miscellaneous circuits for monitoring and controlling the beam were developed, however, and these will be described in a later chapter. Before leaving the high-voltage rectifier some mention should be made of the copper-oxide rectifier built by Westinghouse. The long life of some moderately high-voltage copper-oxide rectifiers installed in broadcasting stations made it seem desirable to try such a supply on the calutron, particularly because such a supply should be able to withstand almost any kind of transients.
Accordingly a 45-kv 2-amp supply was built by Westinghouse and was shipped to the laboratory. However, it was too late in the program to be considered seriously for plant use and remained as a rather successful curiosity.

7. REGULATOR DETAILS

One of the problems in any regulator is to obtain a standard voltage with which to compare the voltage that is to be regulated. Batteries appeared to be the best standards although they were considered very undesirable because of servicing problems. Some of the designs involved a common battery supply to be used for all of the cubicles. The disadvantage here is obvious because of the possibility of trouble involving a large number of cubicles at once. The most appealing standard supply is the VR tube. Consequently much time was spent evolving a good VR standard. The regulation obtained from VR tubes was tested all the way from a few microamperes to 30 ma. It was found that good regulation could be attained at almost any value of current, provided the current was not allowed to vary. It was not found necessary to temperature-control these tubes, which, it must be admitted, was a tremendous advantage. As a result of these studies it was found that VR tubes varied in their characteristics tremendously, and this program helped in part to initiate a more standard procedure in VR tube manufacture. A d-c regulator using a 500-volt regulated standard with an internal VR standard is shown in Fig. 2.34.

To facilitate the testing of regulators in general, a special room was set aside in which a high-voltage supply was built. A filter and 893 tube were set up and the load was supplied by lamps. Two views of this room are shown in Figs. 2.35 and 2.36. It was found that the largest voltage the regulator has to remove is introduced by the magnetic field around the filament of the regulator tube itself. This magnetic field caused by the a-c current in the filament modulates the electron beam and can easily cause as much as a few hundred volts’ ripple. Consequently any rectifier filter which will reduce the ripple to the order of a few hundred volts is quite satisfactory. This can be achieved by a small choke and capacitor. Actually, in the production cubicles used by G.E. no choke was required, the leakage reactance of the transformer being quite sufficient. Transient effects in this test room were simulated by simply short-circuiting the supply to ground. The chief use of this test room, however, was in testing the stability of regulators against oscillation.

A brief discussion of regulator theory (see Chap. 1) as applied to the high-voltage regulator might be included here since the theory in Chap.
1 was of a very general nature and applies to any regulator. A few basic statements may be quoted as follows: A regulator can only remove the effects of an impulse when the correcting pulse travels through the feedback amplifier with negligible time delay and negligible change in shape. Obviously, if the pulse travels too slowly it cannot arrive back in time to annul the initiating pulse completely. Since all amplifiers have a certain minimum time delay, impulses shorter than the minimum cannot be allowed to appear. The 0.01- or 0.02-mmf capacitor in the output of the rectifier will ensure that the potential cannot change with rapidity unless there is an impulse source with enormous power behind it, such as a spark inside the calutron.

The conditions which the regulator must satisfy should now be examined. It must hold the average voltage constant to 1 part in 2000 and must reduce all transient effects such as the 360-cycle ripple to about 15 volts peak. The gain of the amplifier \((\mu \beta)\) must then remain fairly constant until well past 360 cycles per second. After that it must attenuate as a function of frequency as fast as possible but never exceeding, as pointed out in Chap. 1, a factor of 12 db/octave (or a factor of 4 in voltage) until \(\mu \beta\) is less than 1.

The laboratory regulators were initially all of the type in which attenuation is provided by the output capacitor only, which means that the system alternates by a factor of 2 per octave and a constant phase shift of 90 deg until the gain becomes less than one. Later on this situation was improved by attenuating a little faster. This was first done by attenuating in two places by a resistor-capacitor combination, which gave a factor of 4 per octave and a phase shift of 180 deg. Since this would be certain to cause oscillation, a phase advance network was added to reduce the phase shift to 150 deg and also to reduce the attenuation to a little over a factor of 3 per octave. As this system was rather difficult to adjust, it was replaced by internal feedback within the feedback loop, as described in Chap. 1. This internal feedback loop was frequency sensitive and was used to reduce the gain as a function of frequency in addition to the attenuation provided by the output capacitor. One of these circuits (developed by W. H. Nelson and used on XA) is shown in Fig. 2.37. The regulators installed in the plant differed from this circuit only in the manner in which the power supplies were arranged, which is a matter of preference and not of basic design.

8. BETA CUBICLE

A few words should be said about the Beta electrical equipment. The laboratory developed no electrical equipment for the Beta process and
followed the design only through visiting representatives at Pittsfield, Mass. The testing of the Beta equipment was done at Oak Ridge by both Tennessee Eastman and the Radiation Laboratory. Except for a few minor wiring changes to eliminate surge pickup voltages, the Beta cubicle performed as expected.

Fig. 2.1—Saturable-reactor type regulator.

Fig. 2.2—Simplified diagram of first d-c regulator.
Fig. 2.3 — Early high-voltage power supply for the calutron.

Fig. 2.4 — Power supply for calutron with grounded source.

Fig. 2.5 — R-f regulator using frequency modulation.
Fig. 2.6—Essential elements of r-f regulator using amplitude modulation.
Fig. 2.7—A rectifier using 214 tube. The original homemade filament transformers have been replaced with a commercial type.
Fig. 2.8—Simplified schematic of the high-voltage supply for the calutron in the 184-in. magnet. The large 2500-kva breaker was soon replaced by a step-start device.
Fig. 2.9 — A typical high-voltage duct with central conduit conductor is shown in upper left-hand corner. This particular duct is for the acceleration supply on D1. The connection to the tank is made with x-ray cable. The screen cage in the center surrounds the electrodes of the first Alpha I prototype unit. The large square duct leading from the cage through the floor houses the decel supply conduit.
Fig. 2.10—Metal-frame high-voltage duct. Sheet-metal sides were added when installed.
Fig. 2.11—R-f regulator with stabilizing capacitance on grid of 893. The unit in the upper left-hand corner is the r-f amplifier at ground potential. Energy is fed to the central unit at high potential by means of the coupling between the two pancake coils near the lower left-hand corner.
Fig. 2.12—High-voltage disconnect.
Fig. 2.13—Supply for D1. The rectifier transformer is seen on the left. The step-start panel is in the center.
Fig. 2.14—Simplified schematic showing the accel supply connected to the decel supply.
Fig. 2.15 — Caged-in area to the left shows space occupied by accel and decel supplies. The accel high-voltage duct is seen coming out of the top of the control room in the center foreground (see Fig. 2.10).
Fig. 2.16—XA cubicle during assembly. The side panels have been removed. Handling eyelets are on the top near edge. The lower half of the front panel (right-hand side of picture) is not in place. Brackets for Machlett tubes can be seen on the left. The accel transformer is in the center of the picture.
Fig. 2.17—Side-front view of XA cubicle. Lower half of front panel is not in place and upper half is swung open to allow access to inside.
Fig. 2.18—Front view of XA cubicle. The 893 and d-c regulator are in the lower foreground. Ion-gauge supply and hash-indicating oscilloscope are directly above. The arc meters are in the horizontal strip, while above, in insulated boxes, are the beam meters and decel ammeter and voltmeter on the right, and the accel ammeter and voltmeter in the small box to the left.
Fig. 2.19—XA cubicle in operation. All controls are on the strip at hand level. The window in the lower panel allows a view of the regulator tube and regulator amplifier meter (see Fig. 2.18). The window was later covered because of x-rays from the 893 while emission-limiting.
Fig. 2.20—XA r-f regulator.
Fig. 2.21—XA d-c regulator.

Fig. 2.22—Shielded meter boxes for use at high voltage.
Fig. 2.23—Cable terminations. Ground sheath of cable was always tied to the screen cage as directly as possible and not to a common ground lead.

Fig. 2.24—Cable terminating capacitance and inductance. All leads in dotted circle were made as short as possible.
Fig. 2.25—G.E. cubicles in the 184-in.-cycotron laboratory. The central panel is a test arc and heater supply.
Fig. 2.26—(a) Laboratory rectifier circuit. (b) Rectifier circuit in first G.E. cubicle.

Fig. 2.27—Use of diode limiter to limit emission on sparks.
Fig. 2.28—Type 214 3-phase half-wave rectifier and type 562 emission limiter for accel supply.
Fig. 2.29—The 3-phase full-wave supply and type 214 emission limiter.
Fig. 2.30—High-voltage supply for the D1 calutron.
Fig. 2.31—High-voltage room for XAH installation.
Fig. 2.32—XAH control desk under construction.
Fig. 2.33 — Control wiring, XAH.
Fig. 2.34 — The d-c regulator using a 500-volt regulated standard with an internal VR standard.
Fig. 2.35—Power-supply controls for high voltage behind screen.
Fig. 2.36—Power supply and lamp dummy load. Test regulator and 893 in foreground.

Fig. 2.37—Essential elements of internal feedback circuit.
Chapter 3

ARC REGULATION AND TEMPERATURE CONTROL

By R. deLiban

PART I. ARC REGULATION

1. INTRODUCTION

Early in the development of calutron ion sources, experiments indicated that an effective ion source was obtained by accelerating and projecting a stream of electrons from a thermonically emissive cathode into an arc chamber containing the vaporized charge material. In this manner an arc discharge was maintained in the vapor, and the positive ions formed could be withdrawn by a strong electric field to furnish the necessary ion beam for the calutron. In order to provide the required stability and focus, it was found necessary to provide some means of arc-power regulation. Although this could be accomplished by controlling the arc voltage or current, or both, the regulation of the current alone appeared to be the most practicable. Thus, a major part of the ensuing development was concerned with improved methods for regulating and controlling the arc in this basic arrangement. Many different approaches were tried and abandoned, only to be revived later in an improved form. The emphasis at any particular time depended, among other things, upon the state of development of other related calutron components, the availability of equipment and materials, and the status of the proposed over-all plant design. Influencing the entire development was the urgency of wartime with its attendant deadlines, policy changes, and personnel transfers. Hence, the general course of development was not so well defined as it might have been. However, the main development is traced in the following sections.
2. ORIGINAL CONTROL SYSTEM

The first usable system of arc control, as might be expected, embodied the most direct approach with no provision for automatic regulation. A tungsten filament, heated by storage batteries, furnished the electron emission, and a motor-generator or power supply furnished the arc voltage. The filament current, and hence the arc current, was controlled by manually adjusting a mercury resistor connected in series with the arc filament. The resistor consisted of a mercury column in a glass tube with adjustable electrodes cooled by means of a surrounding water jacket.

Although this system provided some control of the arc current, it was not very satisfactory. The storage batteries and mercury resistor were bulky and messy, but could not be replaced because more desirable equipment for handling the heavy filament currents was not yet available. The life of the filament was very limited because of the effect of ion bombardment by the arc. Also, the filament current varied erratically, not only because of the varying parameters such as filament emission, arc voltage, temperature, and vapor pressure, but also because of the heating effect of the arc upon the filament. After an arc discharge was obtained, the filament current had to be reduced rapidly to counteract this effect. Under some conditions this phenomenon would become cumulative, the filament emission increasing because of the effect of the arc, the increased emission causing greater arc current, which in turn caused greater heating, and so on, until a runaway arc current resulted.

3. BOMBARDMENT-HEATED CATHODE SYSTEMS

To remedy the above-mentioned difficulties and to provide automatic regulation, a new system was proposed in which a block of tungsten was heated indirectly to thermionically emissive temperatures by electron bombardment from a directly heated filament, the tungsten block acting as the arc cathode. Since the cathode could now be made larger than the original filament and could conceivably be rotated to offer a new surface to the arc after a given amount of wear, it was felt that much longer cathode life could be obtained, while the life period of the directly heated filament of this system should present no problem inasmuch as it was shielded from ion bombardment by the intervening cathode. Also, the larger possible size of the indirectly heated cathode should not only allow wider arcs, then
thought desirable for larger beams, but should reduce the heating
effect of the arc upon the emitting cathode. Finally, this system
should allow a more effective and compact method of arc-current
control and regulation since high-current apparatus would not be
required.

3.1 Diode-type Regulating System. Initial experiments with the
indirectly heated system showed that a simple manually controlled
arrangement, using only a filament and a bombardment-heated cathode
in a diode arrangement, could be operated satisfactorily. Shielding
of the cavity between the filament and the cathode to exclude vapor was
required to prevent an arc discharge from taking place in the filament-
cathode cavity and to reduce sparking between filament and cathode.
Of course adequate filament heating was required to give space-
charge-limited bombardment current to the cathode.

Regulating this arc arrangement resolved itself into two main prob-
lems: (1) the method to be used in obtaining a signal from the arc
current and (2) the method to be used in varying the bombardment
heating of the indirect cathode in response to this arc-current signal.
A number of ways of accomplishing both these objectives were devised
and combined in various complete regulating circuits. One of these
is shown in Fig. 3.1. In this arrangement a coil $L_1$, carrying a por-
tion of the arc current, is arranged to control the plate resistance of
a vacuum tube $V_1$, the tube being oriented in the coil with its axis
parallel to the magnetic field in a magnetron arrangement. A change
in the plate resistance of tube $V_1$ produces a voltage variation across
resistor $R_1$ which is amplified by tube $V_2$ and impressed upon the
grids of parallel tubes $V_3$, which act as a variable impedance in series
with the bombardment voltage supply and the indirectly heated cathode.
In operation, an increase in arc current intensifies the field of coil
$L_1$ and thereby reduces the plate-cathode current of tube $V_1$. The
resultant increased positive voltage appearing at the grid of tube $V_2$
is amplified and impressed upon the grids of tubes $V_3$ as an increased
negative signal, raising their plate-cathode resistance. This in turn
reduces the voltage applied between the filament and cathode of the
ion source and hence also reduces the bombardment heating of the
cathode by the electron stream from the filament. As the cathode
temperature is lowered the reduced thermionic emissivity there-
from causes a reduction in arc current, counteracting the original
increase and restoring the arc current to approximately its original
value. In a similar manner the arc current is restored to normal
should it tend to decrease.

This arc-regulating arrangement proved quite satisfactory in
remedying the disadvantages of the original directly heated filament
system, but it soon became apparent that the circuit was more complex than was desirable in view of the possibility of mass production for the plant.

In the arrangement of Fig. 3.2 an arc-current signal, obtained from a variable resistor connected in series with the arc circuit, controls the impedance of tubes $V_1$ and $V_2$, which are connected as a variable load across the secondary of transformer $T_1$. The primary of this transformer is connected in series with the power line feeding the grid bias supply of the variable impedance tubes $V_3$, which are connected as before, to control the bombardment heating of the cathode. An increase in arc current reduces the plate-cathode resistance of tubes $V_1$ and $V_2$, causing a reduction in the effective impedance appearing at the primary of transformer $T_1$. The increased negative grid bias from the bias supply increases the effective resistance of tubes $V_3$ and ultimately reduces the arc current in the manner described above. Another simplified circuit is shown in Fig. 3.3. In this arrangement a signal from a series arc resistor is compared with a standard voltage, and the difference is used to control directly the variable impedance tubes in the cathode circuit. However, the reduced loop gain precludes as close regulation as is obtainable with a circuit such as that of Fig. 3.1. Thus most of the attempts at circuit simplification involved a consequent reduction in regulation performance. Also, the total power requirements of even the simpler arrangements was felt to be excessive for plant operation. Some new approach seemed advisable. With this in mind, experiments were started on a double-cathode system.

3.2 Mutually Heated Double Cathode. In this system of arc regulation, illustrated in Fig. 3.4, a first cathode is bombardment heated by a stream of electrons accelerated from a directly heated filament. Thermionic emission from the first cathode, accelerated during each half-cycle of the a-c bombardment voltage, then heats the second cathode by bombardment. On alternate half-cycles the second cathode in turn furnishes emission for bombardment heating of the first cathode and in addition provides emission for the arc of the ion source. After the system is initially heated it becomes self-sustaining, and the starting filament and associated supplies may be turned off and used to start other units. It was expected that a considerable saving of electrical equipment could thus be effected. In theory, the bombardment current between cathodes was to be space-charge-limited to maintain stability, so that if the temperature of one cathode should increase, it would not greatly affect the bombardment heating of the other. If the two were emission-limited this would not be the case, as an increase in temperature of one cathode would increase the bom-
barriment of the other, which would in turn increase the bombardment heating of the first still further, causing an unstable condition. On the other hand, emission limiting of the arc emission from the second cathode was to be used to allow control of the arc current by the usual method of regulating the arc cathode temperature, which was to be accomplished in this case by adjusting the a-c bombardment voltage. However, the activation of the second cathode by the arc was greater than expected, necessitating a reduction in cathode temperature to keep the emission down to the value required for the normal low-current high-voltage type of arc. But when this was done (by reducing the bombardment voltage) the system became unstable. This was probably because the intercathode bombardment current was no longer emission-limited since, as the bombardment voltage was reduced, the available cathode emission dropped more rapidly than the bombardment current and eventually became less than the amount required to maintain a space charge. Attempts at reducing the required space-charge-limited current by increasing the intercathode spacing were not successful, and the system was finally abandoned in favor of another type of bombardment-heated arrangement.

3.3 The a-c Triode-type Regulating System. This type of indirectly heated cathode ion source makes use of a grid interposed between the filament and cathode of the diode-type ion source. Although this appreciably complicates the physical structure, it was felt that the increased simplification of the associated electrical circuits obtainable because of the amplification of the grid arrangement would result in an over-all improvement. In addition, the filament would be shielded from the heating effects of the cathode. This had caused some difficulty in the diode arrangement because after an arc was struck a reduction of filament power was required. As a further simplification, an a-c bombardment supply was used on the initial experiments. The grid structure consisted of a metal plate with a rectangular aperture slightly larger than the projection of the filament along the magnetic field lines to avoid interception of the electron stream. The location of the grid and the value of the negative bias applied thereto were adjusted just enough to cancel the positive electric field produced at the cathode by the filament on the half-cycle in which the filament was more positive than the cathode. In this manner the cathode was prevented from emitting to the filament on alternate half-cycles of the a-c bombardment voltage.

Although this arrangement functioned as expected, giving good control of the bombardment heating of the cathode with a relatively small signal on the grid, a number of other difficulties were encountered. Foremost of these was the fact that the power was delivered to the
cathode in half-cycle pulses every sixtieth of a second because of the a-c bombardment voltage. This meant that the peak power must be rather high to obtain adequate average power, and this in turn required either a higher than normal filament temperature or a higher bombardment voltage. If the former were used the filament life would be shortened; if the latter were used considerable difficulty would be encountered with sparking, which in turn would require increased spacing of the triode elements. Further difficulties were encountered because the emission furnished the arc by the cathode followed to some extent the pulsating bombardment power delivered to the cathode. This of course produced an undesirable 60-cycle modulation of the arc. In view of all these difficulties, further attempts to operate this grid-controlled arrangement on alternating current were abandoned in favor of a d-c system.

3.4 The d-c Triode-type Regulating System. The substitution of a d-c bombardment supply for the a-c supply of the above system, while retaining the same triode configuration in the ion source, eliminated most of the previous difficulties and provided a satisfactory arc-regulation system with relatively simple associated electrical circuits. Figure 3.5 illustrates a system of this type. Here again a magnetron arrangement was used to obtain a satisfactory voltage signal from the arc current without introducing an excessive voltage drop in the arc circuit. The tetrode tube $V_1$ and resistor $R_1$, connected in series across the bombardment-voltage supply, form an electronically variable voltage divider responsive to the magnetic field generated by a portion of the arc current flowing through coil $L_1$. The voltage appearing across resistor $R_1$ is applied as a positive bias between the grid and cathode of the triode ion-source structure. In operation, an increase in arc current flowing through coil $L_1$ strengthens the magnetic field thereof and thereby increases the plate-cathode impedance of tube $V_1$, which in turn decreases the voltage drop appearing across resistor $R_1$. The reduction in positive bias applied to the grid of the ion source causes a reduction in the bombardment heating of the cathode and a consequent reduction in the thermionic emission therefrom. The resultant drop in arc current counteracts the original increase and tends to stabilize the arc current about its original value. This operating value may be changed manually by adjustment of resistor $R_2$, which shunts a portion of the arc current around the magnetron coil.

Other schemes incorporating somewhat higher gain and giving correspondingly better regulation were also used successfully. In one of these, for example, the arc current controlled a saturable reactor,
which in turn controlled the input power to a bias supply for the ion-source grid. In another, a series resistor in the arc circuit furnished a grid signal to a pair of thyratrons in a full-wave grid-controlled rectifier circuit, the output of which biased the grid of the ion source. Still another arrangement combined a saturable reactor with a grid-controlled rectifier using a pair of triodes to provide a further increase of gain. In general, however, the increased gain of these circuits was obtained only at the cost of increased complexity of apparatus.

It is noted that in some of these grid-controlled arrangements the ion-source grid is run at a positive potential with respect to the filament, which ordinarily would cause difficulty from excessive grid current. However, the calutron magnetic field effectively collimates the electron stream from the filament in a narrow beam whose cross section is roughly equal to the projection of the filament area, and the size and shape of the grid aperture are chosen to allow this beam to pass through the grid plate unimpeded. The grid current was thus kept to a minimum. However, considerable difficulty was experienced initially in keeping the grid in alignment.

Further trouble was encountered with excessive sparking between the bombardment cathode and the grid and filament structures, particularly during the "bakeout" period of a run when the calutron tank pressure was higher than normal. Reducing this sparking by increasing the spacing between the ion-source elements was inherently an undesirable solution because of the desirability of keeping the space requirements to a minimum, and yet the sparking could not be tolerated because of the excessive current drain on the bombardment-voltage supply.

3.5 Arc Snuffers. Because of the above-mentioned difficulties with arc-overs between the triode elements, a number of auxiliary circuits were devised which would limit the bombardment current during arc-overs to a predetermined value. Although in operation this did protect the equipment from overload, it was ineffective in extinguishing an arc-over. In some cases an arc-over between the triode filament and bombardment cathode not only maintained itself but traveled on down the supporting elements to a glass insulating plate at the rear of the structure, which eventually shattered. Consequently the circuits were modified so that an increase in bombardment current above a predetermined value cut off the bombardment supply completely for a finite fraction of a second, allowing the undesirable arcing to extinguish. One of these circuits was incorporated in a grid-controlled bombardment supply as shown in Fig. 3.6. On alternate half-cycles of the supply-line voltage, the maximum positive voltage impressed upon
the anode of the gas tube $V_1$ consists of the peak secondary voltage of transformer $T_1$ plus the voltage drop produced by the total bombardment current flowing through resistor $R_1$. In normal operation this maximum voltage is insufficient to fire the gas tube, but should the bombardment current rise to a sufficiently high value, predetermined by the setting of resistor $R_p$, tube $V_1$ will fire, producing a voltage of the polarity shown across resistor $R_2$, which in turn causes capacitor $C_1$ to charge. In a short period of time the voltage across capacitor $C_1$, which is impressed upon the grids of the thyratron rectifier tubes $V_2$ and $V_3$, reaches a sufficiently high value to cut off these tubes completely, thus removing the bombardment voltage from the ion-source unit. After a period of time sufficient to extinguish the arc-over, determined by the time constant of resistors $R_2$ and $R_3$ and capacitor $C_1$, and by the degree of overload, the charge across capacitor $C_1$ leaks off sufficiently to allow conduction of tubes $V_2$ and $V_3$, and the bombardment voltage is reapplied to the ion source.

This system of arc snuffing proved quite satisfactory with the triode type of arc regulator and in addition could be used with the diode regulating system discussed previously. It allowed, for the first time, the long-desired completely automatic control of the arc, the operator's constant attention no longer being required during the bakeout period. The number of annoying time-wasting shutdowns caused by blown fuses and insulator breakages was greatly decreased. On the other hand, the system was inherently an indirect solution of the ion-source arcing problem inasmuch as it did not eliminate the arcing but merely allowed continued operation in spite of it. Thus, although both the diode type and the somewhat simpler triode type of arc-regulation systems, when used in conjunction with the arc snuffer, provided satisfactory control for laboratory use, the combined circuits were still more complex than was desirable for plant operation.

4. DIRECTLY HEATED CATHODE REGULATION

At this stage of the development a great deal had been learned about the use of an arc discharge in a magnetic field for an ion source, particularly about the stability requirements, the type of emitting cathode required, and the effects of heating and activation of the cathode by the arc. In the light of this new information the case for the directly heated cathode began to look a little more promising. It will be recalled that the original directly heated filament was abandoned in favor of the indirect cathode mainly because it could not furnish the large emitting area required by the large collimating slots used at that time, it had an unsatisfactory life period, adequate regulating
and power-supply equipment was not available at that time, and a larger cathode was thought to be less affected by heating and activation from the arc. However, the development of the long narrow collimating slot now eliminated the necessity for a large cathode area and also reduced the effect of ion bombardment of the cathode by the arc. This bombardment effect was further reduced by the lower arc voltage required with the narrow collimating slots. Also, the required power and regulation equipment for the directly heated filament was now available. Thus, with the development of satisfactory indirectly heated systems completed, attention was again focused on the directly heated filament. If it could be made to operate satisfactorily, it would offer numerous advantages over the indirectly heated types. The filament need not cover an area much greater than the narrow collimating slot; therefore less heating power would be required. Both the mechanical construction and the electrical circuits would be greatly simplified.

4.1 Saturable-reactor System. One of the first directly heated filament systems to be installed was that illustrated in Fig. 3.7. A simple filament, powered with direct current from a dry-plate rectifier, furnished emission to the ion-source arc directly with no intervening elements. The a-c windings of a 3-phase saturable reactor were connected in series with the power line feeding the filament rectifier, and the reactor d-c control winding, shunted by a variable resistor, was connected in series with the arc-current circuit. In addition, the reactor was supplied with another winding which was connected to a battery bias supply through a bias control resistor. The magnetomotive force (mmf) of this winding was used to saturate the reactor core, allowing the mmf of the control winding, which was in opposition to that of the bias winding, to desaturate the core. This gave the proper relationship between the arc current and the a-c impedance of the reactor for the required degenerative action around the arc-regulating loop. In another aspect, the bias-winding mmf can be considered as a magnetic standard, against which the arc-current mmf is compared, the difference being used to control the saturation of the reactor. In operation any tendency of the arc current to increase would lower the net d-c magnetization of the saturable reactor, thereby increasing the effective impedance of the a-c windings and reducing the power supplied to the filament. The resultant reduction in arc current would be in opposition to the original tendency to increase and would thus tend to stabilize the arc current about its original value. The degree of regulation and the value of arc current desired were adjustable by means of the two variable resistors.
After the initial trials demonstrated that this system was entirely practical, further development of indirectly heated cathode systems was all but dropped, and most of the effort was thereafter spent on improving the direct filament regulating circuits and lengthening the filament life, with the ultimate aim of developing a simple and reliable system for plant operation.

Although the system of Fig. 3.7 was used for quite some time it had several shortcomings. For instance, it was essential in operation to always maintain a net d-c magnetization in the reactor in the same direction as that supplied by the bias winding. If the magnetization reversed for any reason, the degenerative character of the regulating loop became regenerative since the a-c impedance of the reactor would then increase with an increase in arc current, and a runaway condition would obtain in which the maximum power possible was delivered to the filament. If this did not result in a burned-out filament, it was at least an operating inconvenience since it could be brought about by unknowingly dropping the bias-winding current too low or by increasing the control-winding current too far. Also, the degree of regulation obtainable and the flexibility of the circuit were no longer adequate in view of the more stringent requirements brought about by the general all-around improvements that had been made in calutron operation by this time. Thus, although this arrangement represented about the ultimate in the long-sought simplicity, it now appeared that the state of the calutron art called for some sacrifice in simplicity in order to obtain a higher degree of regulation, more flexibility, and greater operating convenience.

4.2 Thyratron-reactor Systems. Accordingly, a number of regulators were developed which utilized a d-c amplifier feeding a pair of thyratron tubes, which in turn controlled a saturable reactor. The a-c windings of the reactor were connected in the filament power line to control the filament power in the usual manner in response to an amplified signal derived from a series arc resistor. No bias winding was required on the reactor. Although some of these models were quite satisfactory, the general design trend seemed to be in the direction of overcomplication, some of the regulators having as many as 10 to 12 vacuum tubes. This degree of complexity did not seem to be justified, especially since system instability and even loop oscillations set in before a large fraction of the total gain provided could be utilized. This is to be expected, of course, as an examination of the loop characteristics will show. In a regulating loop of this type the longest or controlling time constant is the thermal delay of the filament. This is a relatively fixed quantity dependent upon other design considerations. The magnitude, number, and distribution of
the remaining loop time constants will then determine the maximum gain that can be used with stability. However, the reactor time constant is so large in comparison with the others that it becomes the principal factor. Increased gain can only be used if this time constant is reduced. This may be accomplished to some extent by careful design of the reactor and by the use of as high a series resistance as practicable in the control-winding circuit, but practical limitations prevent any large improvement. Replacement of the thy-ratrons with power pentodes of high a-c plate resistance would be effective, but this did not seem a desirable solution.

The regulation circuit illustrated in Fig. 3.8 was designed to provide the required gain, flexibility, and ease of control with the absolute minimum of added circuit complications. This would inherently provide the lowest probability of failure, since there are fewer circuit components, and would also make the servicing problem easier. It was desired to accommodate both hot and cold sources without making any changes in the circuit. Previously a hot-source regulator was floated at high potential, the controls being brought out to the operating position by means of insulating rods. This disadvantage was eliminated by obtaining a signal proportional to the arc current from the primary side of the arc supply, instead of from a series arc resistor. Since the arc-voltage supply was powered through insulating transformers, this allowed the signal circuit to be operated at or near ground potential. In addition, the saturable reactor was connected in series with the primary side of the filament insulating transformers, so that this circuit was also near ground potential. Thus the entire arc regulator could be operated at ground potential, offering numerous advantages in ease of installation, operation, and maintenance. Considering the operation of the circuit, an a-c signal, substantially proportional to the d-c arc current, is obtained from the three current transformers, $T_1$, $T_2$, and $T_3$, which are loaded by terminating resistors $R_1$, $R_2$, and $R_3$. This 3-phase signal is amplified by the voltage stepup of the delta-Y connected transformers $T_4$, $T_5$, and $T_6$, and is then rectified by the 3-phase full-wave rectifier consisting of the double triodes $V_1$, $V_2$, and $V_3$. The d-c output is filtered by a tuned filter designed to offer maximum suppression of the 360-cps ripple component while adding a minimum additional time lag to the loop. Resistor $R_4$ is chosen to terminate the filter in the proper value for optimum transient response. This resistor also serves as the arc-current control. It is noted that in the actual installation this resistor was a small 1- or 2-watt resistor located at the operating position, in contrast to the bulky power resistors used for control in some of the previous systems. Although this may appear to be of little advantage, any saving of
space at the crowded operating position was well worth while. The signal from the arc-current control resistor is compared with a standard voltage obtained from the battery $B_1$, and the difference signal is impressed between the grids and cathodes of the thyratrons $V_4$ and $V_5$. Also impressed on these grid circuits are prephased a-c signals obtained from transformers $T_7$ via the phasing network comprising resistors $R_5$ and $R_6$ and capacitors $C_1$ and $C_2$. These combined signals provide control of the firing points of the thyratrons in response to variations of the impressed d-c signal in the conventional manner. However, the thyratrons used were of the inert-gas type instead of the mercury-vapor type and therefore had much more stable characteristics. This, in conjunction with operation from a regulated line, allowed a small a-c grid voltage to be used, which in turn gave high gain since the thyratron firing points then change much more rapidly with a change in d-c grid voltage. In this manner adequate gain was obtained without resorting to additional d-c amplifiers.

Examine the circuit in more detail, it is noted that the primary of transformer $T_7$ is powered from the same phase as the thyratron plate transformer $T_8$, so that the secondary voltage of $T_7$ is in phase with the thyratron plate voltage. This voltage is reduced by the network of resistors $R_7$, $R_8$, and $R_9$, resistor $R_8$ providing an adjustment of the regulator gain. The voltage across this resistor is then shifted about 90 deg by the phasing networks $R_5$ and $C_1$ and $R_6$ and $C_2$ so that the a-c voltage applied to each grid lags the respective anode voltage by approximately 90 deg. The maximum d-c control current is adjusted to suit the reactor characteristics by adjusting resistor $R_{10}$, which is in series with the control winding. The higher the value of this resistance, the lower the time constant added to the regulating loop by the reactor, which of course is advantageous.

In the use of this regulating system, if the filament were energized when the arc supply is off, the regulator unit would tend to increase the current to the filamentary cathode to a maximum limited only by the capacitance of the filament supply and the saturation impedance of the reactor. Since this might damage or burn out the filament, an auxiliary circuit consisting of relay $RE_1$ and the variable contactor of resistor $R_{10}$ is used. Thus, if the arc supply is deenergized for any reason, the contacts of the relay open and an additional resistance is inserted in series with the d-c control winding of the saturable reactor. This in turn reduces the maximum current that can be supplied to the filament to a safe value. This value was usually made equal to the normal operating current, so that a step-start action was obtained when a cold filament was turned on, the filament first heat-
ing up to an intermediate temperature and then, when the arc voltage was turned on, rising to the higher value required for striking an arc.

Another auxiliary circuit, consisting of resistor $R_{12}$, tube $V_6$, and potentiometer $R_{11}$, is also included in the regulator unit for the purpose of limiting to a predetermined value the maximum filament current obtainable under normal operating conditions. It will be observed that since the filament current is a function of the d-c grid voltage applied to the thyratrons, it can be limited by limiting the grid voltage. This is done by connecting the anode of the limiting tube $V_6$ to the positive end of the standard battery $B_1$, which is at the same d-c potential as the thyratron grids, or by connecting it to an intermediate tap on the battery, which gives the same effect, inasmuch as it is displaced from the grid potential by a fixed amount. The cathode of the limiting tube could be connected to a fixed source of voltage in the conventional manner. However, a more definite limiting action is obtained if it is connected to a tap on resistor $R_{11}$. Then, as the grid voltage is made less negative or more positive by an adjustment of the arc-current control resistor, to obtain a higher arc current, for instance, the cathode of the limiting tube becomes more negative as the anode becomes less negative. Thus the tube passes from nonconduction to conduction more rapidly than if the anode voltage alone were changed, giving the sharper limiting action. The value of resistor $R_{12}$ is very high compared to the other limiting circuit components, so that any further decrease in the negative voltage from resistor $R_4$ appears as a voltage drop across resistor $R_{12}$, while the thyratron grid voltage remains substantially unchanged. The limiting value of filament current is adjusted by means of potentiometer $R_{11}$, which is also located at the operating position for convenience. It is noted that this potentiometer is of much greater resistance than resistor $R_1$, so that it does not appreciably affect the settings of this resistor. Because of the sharpness of this control, the limiting filament current can be set, once the arc has been initiated, to values only slightly higher than the required running current without interfering with the normal regulating action, thus providing maximum protection for the filament. Also, the variation in maximum allowable current with different filaments can be easily accommodated.

During the time that this regulator was in widespread use on the Project the development of improved filaments was proceeding at a rapid pace. Uniformity of emission and long life had been obtained by the proper mechanical configuration and by the use of tantalum. Furthermore, most of the problems associated with running filaments on alternating current were solved. It thus appeared quite likely that a-c filaments would be adopted as standard since this would eliminate the
bulky filament rectifiers required with the d-c type. Attendant with this
development was a large variation in the requirements imposed upon the
arc regulators. Most of the d-c filaments and their associated 3-phase
power supplies were accommodated quite satisfactorily by the above
regulator, but some difficulty was experienced with a few of the a-c
types. Provision had been made for a-c filaments by designing the
saturable reactor so that various series or parallel combinations of
the a-c windings could be used in series with the single-phase power
line feeding the transformer for the filaments, although it was not
thought likely at the time that a-c filaments would come into extensive
use. It was found, however, that the existing reactor could not always
be properly matched to the various filament loads imposed without sac-
rificing either control range or current-carrying capabilities. It was
decided, therefore, to retain this regulator for use with d-c filaments
and design a new model for a-c filaments that had complete flexibil-
ity with respect to the type and magnitude of filament loads to be
regulated.

4.3 Inverse-parallel Thyratron System. This new regulator for
a-c filaments, illustrated in Fig. 3.9, made use of a pair of thyratrons
connected into the filament line in an inverse-parallel arrangement,
replacing the saturable reactor as a variable a-c impedance device.
Certain fundamental features of the previous regulator, having proved
themselves in practice, were retained, such as the method of obtain-
ing a control signal from the arc-supply primary circuit, the step-
start interlock system, and the filament-current limiter, although
these latter two were incorporated in somewhat different fashion.
Again, the emphasis was put on obtaining the required operating fea-
tures with an absolute minimum of circuit complexity.

Considering the circuit in more detail, a portion of the output volt-
age of the signal rectifier, appearing across the arc-current control
resistor \( R_1 \), is compared with a standard voltage from the small
copper-oxide rectifier \( B_1 \), and the difference voltage is impressed upon
the grid-cathode circuit of tube \( V_1 \) in the phasing section. In response
to this d-c signal, the phasing section provides constant-amplitude
variable-phase a-c voltage to the thyratrons \( V_3 \) and \( V_4 \) to control
their firing points. The phasing section is an a-c bridge network
supplied with 60-cycle voltage at points A and C. Tubes \( V_1 \) and \( V_2 \)
form an electronically variable a-c resistance, which, together with
resistors \( R_2 \) and \( R_3 \), is connected in series with capacitor \( C_1 \) to form
the other leg of the bridge. The output voltage appearing across points
B and D is coupled to the thyatron grid circuits by means of trans-
formers \( T_1 \) and \( T_2 \), the primary impedance of which is as high as
possible, to reduce the loading effect on the phasing circuit. Tracing
the a-c path from point A to point B, it is observed that the current
flows from anode to cathode in one triode section through resistors
$R_3$ and $R_2$, and from anode to cathode in one diode section to point B.
During the reverse half-cycle it flows from B to A in a similar man-
ner, but through the other triode and diode sections. If it is assumed
that transformers $T_1$ and $T_2$ offer negligible loading on the bridge
circuit, then when tube $V_1$ is cut off with a high negative signal there
will be an open circuit between points A and B, so that the voltage
appearing between points D and B will be in phase with that between
points D and C. On the other hand, if the total resistance $R_{AB}$ be-
tween points A and B could be reduced to zero, the voltage between
D and B would be in phase with that between D and A, and hence 180
deg out of phase with its original condition. Of course, intermediate
values of $R_{AB}$ would give intermediate values of phase shift, so
that the output voltage could be shifted up to 180 deg by the d-c
control signal on the grid of tube $V_1$. However, this resistance cannot
be reduced beyond a minimum equal to the sum of the effective triode
resistances at zero bias, the effective diode resistance, and the resis-
tances of $R_2$ and $R_3$. Thus the phase of the voltage between D and B can
only approach that between D and A. The total phase swing is limited
somewhat further by the loading effect of transformers $T_1$ and $T_2$.
To compensate for these factors a prephasing network composed of
resistance $R_4$ and capacitor $C_2$ is inserted in the a-c line feeding the
bridge network. This advances the phase initially so that the output
voltage from the bridge at points B and D can be brought in phase
with the line voltage when the resistance between points A and B is
reduced to a minimum. In this manner the thyatron grid voltage can
be brought in phase with the anode voltage, which is also derived from
the same power line, so that the thyatrons can be fired over a full
half-cycle at the maximum current end of the regulating range. This
gives the minimum possible insertion loss in the filament line. In
addition to this prephasing function, resistor $R_4$ and capacitor $C_2$ also
serve to drop the line voltage fed to the bridge network, and hence the
thyatron grid voltage, to the minimum value required for stable
thyatron operation. This gives maximum regulator gain since a
minimum grid swing is then required on tube $V_1$ for a given phase
shift. This tube arrangement thus provides the required gain in ad-
dition to converting the d-c signal of varying amplitude to an a-c
signal of variable phase.

Resistors $R_3$ and relay $RE_1$ provide the step-start arrangement to
this circuit, the relay again being energized by the power line feeding
the arc-voltage supply. When the filament is turned on initially, tube
V₁ is in a condition of minimum resistance and relay RE₁ is open, so that the minimum total resistance between points A and B is a function of the setting of resistance Rₚ. The phase of the thyratron grid signal, and hence the maximum filament current, is thus limited to an intermediate value. When the arc-voltage supply is energized, relay RE₁ short-circuits resistance R₃, leaving resistance R₂ in the circuit, which then serves as the manual filament-current-limiting control by introducing a controllable minimum phase shift in the thyratron grid signal.

The grid resistors limit the grid current, and the thyratron grid capacitors by-pass to cathode the sharp pulses produced during each half-cycle when the grids suddenly draw current. They also eliminate any effect the plate-voltage wave might have on the firing point because of plate-grid capacitance coupling. Resistor R₅, in conjunction with the leakage reactance of the filament transformer, limits the peak current, in case of a short-circuited filament, until the fuse blows.

In addition to the high degree of flexibility obtainable with the inverse thyratron arrangement with respect to filament load, further adaptability was possible by interposing a load-matching transformer between the thyratrons and the filament primary line. Thus, with a 220-volt high-current filament load a 2 to 1 transformer could be used to double the normal regulator current rating without exceeding the voltage rating of the thyratrons, or the regulator could be modified, of course, to accommodate larger thyratron tubes.

In operation, this regulator proved quite satisfactory, provided there was high gain, quick response, flexibility, ease of control, and reliability of operation. It was adopted as a standard regulator for a-c filaments and was produced in quantity for laboratory use.

Some attempts were made to use the inverse-parallel thyratron arrangement for d-c filaments, but since these were powered by a 3-phase rectifier, even the simplest arrangement required a grid-controlled thyratron in each of the three lines, shunted by a gas diode connected in an inverse-parallel arrangement. Although this arrangement could be made to operate satisfactorily, it would require entirely too much equipment. Thus, the thyratron-reactor system of Fig. 3.9 was retained as the standard model for d-c filaments and was also produced in quantity for laboratory use.

5. MULTIPLE-ARC REGULATING SYSTEMS

During the development of the single-arc regulators described above, the source units often utilized two or even four separate arcs,
in which case the required number of single-arc regulators were used for arc control. However, a number of sources were built which used 12 or 18 arcs in a single unit. It became apparent that the amount of equipment and space required for the corresponding number of single-channel regulators was prohibitive. Furthermore, in the case of the directly heated filaments the problem of bringing through the face-plate 12 or 18 pairs of 300-amp conductors was not easily solved. Some attempts were made to control two or three arcs from a single standard regulator by connecting the filaments in series and regulating the common current. Individual manual trimmers were used to adjust the ratio of currents in the separate filaments. However, these arrangements were not very satisfactory; neither were similar arrangements using paralleled filaments. Thus there was a demand for special multiple-arc regulating systems.

5.1 Directly Heated Filaments with Thyratron Control. One of these multiple systems, designed for use with directly heated filaments, is illustrated in Fig. 3.10. The individual filaments are connected to the filament bus through a series resistance $R_1$ of a value comparable to that of the filament itself. This resistance is shunted by a variable-impedance device, responsive to the arc current, so that the total current flowing to the filament is controlled by the arc current. The regulation range was restricted to about 20 per cent of the normal filament current; consequently wires carrying the regulating current had only to carry a fraction of the filament current. Thus only two high-current filament conductors carrying most of the combined filament current were brought through the tank faceplate, the rest being small low-current wires. This greatly simplified the mechanical construction problem. In addition, low-powered regulation equipment could be used throughout. Separate anodes for the arcs were required, these being connected to a common arc supply through individual series resistors $R_2$. The signal from each of these resistors was compared with a separately adjustable standard voltage obtained from a potentiometer $R_3$ and the difference was used to actuate the respective variable-impedance device. The equipment requirements were simplified by the use of common arc, filament, and standard voltage supplies.

Considering the variable-impedance device of Fig. 3.10 in more detail, it is seen that a pair of thyratrons, $V_1$ and $V_2$, are coupled across the filament series resistor $R_1$ by the impedance-matching transformer $T_1$. The firing points of the thyratrons are controlled by the d-c signal from the arc current in combination with prephased a-c signals from the phasing network in the thyratron grid circuit.
The thyratrons are energized by the flow of a portion of the calutron filament current flowing through the primary winding of transformer $T_1$. When the thyratrons are fired, this primary current is at a maximum value, determined by the load resistance in the secondary of transformer $T_1$, which is adjustable by means of resistor $R_4$. When the thyratrons are unfired, the primary current is at a minimum value, determined principally by the primary inductance of transformer $T_1$. Thus control of the firing point of the thyratrons also controls the portion of each half-cycle that the tubes are conducting, and hence controls the average primary current of transformer $T_1$. Since this current flows through the filament, it is apparent that the filament emission is also controlled. Although the arrangement is essentially an on-off device, it may be considered as a continuously variable shunt impedance across the filament series resistor, controlled by a signal from the arc current.

Considering one-half of the symmetrical phasing network, the series circuit comprising resistor $R_5$ and capacitor $C_1$ is energized from transformer $T_1$ to provide an a-c signal between points A and B that lags the anode-cathode voltage of the thyratron by approximately 90 deg. This voltage is reduced to the proper magnitude by the dividing network comprising resistors $R_6$ and $R_7$ and is applied to the grid of the thyratron $V_1$. This signal actually appears between grid and cathode of the thyratron since point B is at the same a-c potential as the cathode. Similarly, an a-c voltage is applied to the grid-cathode circuit of tube $V_2$, lagging the anode voltage of this tube by 90 deg. The d-c signal derived from the arc current is also impressed upon the grid circuit of the thyratrons between the cathode circuit and point C, which is electrically neutral with respect to alternating current because of symmetry. These combined a-c and d-c signals act to control the firing points of the thyratrons in the usual manner.

Although this system operated satisfactorily, it was soon abandoned in favor of a multiple system using an improved saturable-reactor arrangement.

5.2 Directly Heated Filaments with Saturable-reactor Control. This system, illustrated in Fig. 3.11, employed the same filament arrangement as in the system just described, the main filament current from transformer $T_1$ being carried into the calutron tank by means of a pair of very high current conductors across which were connected the filaments in series with their individual filament resistors $R_1$. Relatively light conductors, from the junction points of the filaments and their series resistors and from individual arc anodes, were brought out through the faceplate. In order to simplify the circuit, by eliminating the necessity for reactor-bias windings and an
associated bias supply, the junction points between the filaments and their resistors were connected through individual saturable reactors to a second filament transformer $T_2$. This was powered from another supply line that was phased 120 deg lagging with respect to the line feeding the main filament transformer. An additional phase lag was caused by the inductive reactance of the saturable reactors, so that the current flow to the filaments by this path was approximately 180 deg out of phase with the main filament current. Thus a decrease in the a-c impedance of a saturable reactor, which is due to an increase in arc current, causes a net reduction in the total filament current. This phasing arrangement also increased the effectiveness of the reactor by providing a maximum change in filament current for a given impedance change of the reactor.

Each reactor was controlled by a single d-c winding through which a portion of the arc current flowed. Because of the elimination of the bias winding previously used, which acted as a magnetic standard against which the mmf of the control winding was compared, some other type was required. A standard voltage supply was used, against which the voltage drop produced by the arc current flowing through the arc series resistor was compared. This arrangement maintains a high rate of change of reactor control current vs. arc current without requiring the reactor to carry the entire arc current. This is utilized in the design of the reactor to provide a high effective amplification. The dry-plate rectifier $S_1$ prevents reversal of the reactor control current, thus eliminating both the possibility of the runaway condition discussed previously and the possibility that the reactor would be damaged by the high reversed current produced in the control winding by the standard voltage when there is no arc current.

It is apparent that no current flows in the control windings until the voltage drop across the arc series resistor $R_2$ is equal to the standard voltage. The value of arc current required to give this voltage is a function of the setting of resistors $R_2$ so that these provide manual control of the regulated arc current over the desired range.

Again in this arrangement, control range was sacrificed to allow the use of low-current conductors and equipment. It should be pointed out, however, that although the filament current can only be varied over a range of about 20 per cent above its normal operating value, this is adequate for arc-current control since an arc discharge is not usually obtainable at the lower filament currents.

5.3 Indirectly Heated Filaments with Grid Control. This arrangement, illustrated in Fig. 3.12, utilized the same saturable circuit as the system discussed previously. In this case, however, the reactor controlled a bias rectifier, which applied a negative potential to the
grid of a triode-type indirectly heated cathode arrangement. Thus the scheme is similar to one of the systems discussed under the d-c triode-type arrangement. A negative grid is used so that it may be controlled directly from the rectifier. This requires a higher bombardment voltage supply to obtain the required emission current but eliminates the additional grid-bias supply that would be required if the grid were run in the positive region. Again, common power supplies were used wherever possible, so that the multiplication of equipment required for a large number of arcs was held to a minimum.

The above circuits were used with varying degrees of success with the multiple-source units having a large number of arcs but were abandoned when this development program was terminated.

6. OTHER REGULATING SCHEMES

While the foregoing discussion traces the general history of arc-regulator development, there were in addition many other schemes proposed. Although some of these operated more or less satisfactorily, they did not influence to a great degree the general pattern of development. However, as a matter of interest a few of these will be discussed briefly.

Figure 3.13 illustrates a modification of the triode-type arrangement of indirectly heated cathode discussed previously in which the calutron accelerating-voltage supply is used as a source of bombardment voltage, thus eliminating a separate bombardment supply. The control grid of the source is connected directly to ground, and the filament is returned to ground through the auxiliary tube $V_1$, which acts as an electronically variable bias resistance in the ion-generator filament circuit and also provides the required amplification. Thus the ion-generator grid bias, and hence the bombardment heating of the arc cathode, is controlled by the grid signal on vacuum tube $V_1$. This signal is derived from the arc current by means of current transformers in the arc-supply primary circuit and a signal rectifier in an arrangement described previously. An additional voltage from a bias supply is inserted in the grid circuit of tube $V_1$ by means of the arc-control potentiometer $R_1$, which thus provides manual control of the regulated arc current. Resistor $R_2$, potentiometer $R_3$, and the diode tube $V_2$ comprise a limiting circuit that limits the maximum obtainable bombardment current by preventing the grid of tube $V_1$ from going less negative than the potential appearing at the variable contactor of potentiometer $R_3$. This potentiometer thus serves as the cathode-current limiting control.
It is noted that all of the circuit components, including the various controls, located below the dashed line of Fig. 3.13 are at or near ground potential, whereas those components above the line are at practically the full calutron accelerating potential. Thus the spacing of the triode elements must be much greater than in the triode type discussed previously since about 40 kv appears between the arc cathode and the control grid and filament. This high accelerating voltage imparts a high velocity to the electrons bombarding the cathode, so that only a small bombardment current is required to give adequate cathode heating. This is an advantage, of course, because of the decreased current-handling requirements imposed upon the other circuit components associated with the bombardment circuit. On the other hand, the high voltage between the triode elements is apt to cause increased difficulties with sparking.

In the operation of this system the bombardment current is varied automatically to maintain a constant arc current. Although this variable bombardment current is drawn from the source (M) voltage supply it does not appreciably affect the M voltage stability because of the high degree of voltage regulation provided by the M regulator. For the same reason, variations in the normal source (M) current drain, caused by the variable tank load, do not affect the constancy of the bombardment voltage furnished the ion generator. However, during M-sparking flurries and glow discharges there is much interaction between the source (M) system and the arc-regulating system because the average M voltage, and hence the bombardment voltage also, is considerably reduced. To counteract this, the arc regulator will increase the bombardment current by decreasing the negative bias supplied to the ion-generator control grid, but, if this does not provide sufficient bombardment power to the arc cathode, its emission will decrease, the regulating action will cease, and the arc current will drop to a lower unregulated value.

Another regulating system proposed the use of a thermal element to be heated by passing the arc current through it. The element could be a section of wire with a high coefficient of expansion. The small movement caused by variations in arc current was to be amplified mechanically and used to vary the grid-filament spacing of a triode-type ion source. This would in turn control the bombardment heating of the arc cathode and hence the arc current also. This system offered the advantage of extreme electrical simplicity, but this was outweighed by the required mechanical complications, and the scheme was dropped while still in the paper stage. A similar scheme, in which the control grid was to be operated by an electromagnet ener-
gized from the arc current, was also abandoned before it was actually tried.

7. ARC POWER SUPPLIES

In general, the various power supplies used with the calutron ion sources were of conventional design. The arc supply usually utilized a 3-phase full-wave or 6-phase half-wave rectifier circuit with mercury-vapor rectifier tubes. A single-section low-pass filter was usually used, often with a tuned series element. Overload protection was provided by overload cutout relays and fuses. The arc voltage was adjusted by means of a primary variac or other type of variable autotransformer, either manual or motor-driven. At one time motor generators were also used for supplying arc power, but they did not prove as practical for the laboratory installations. The size of the arc supplies ranged from small 10-amp units for single-arc circuits to 100-amp units for multiple-arc sources.

The a-c filament power was of course furnished by a single-phase filament transformer. The d-c power was provided by a 3-phase transformer and full-wave dry-plate rectifiers, usually with no filtering. A good deal of attention was given to the method of carrying current from the supplies to the filaments in the tank, particularly in the case of a-c filaments, where the reactive line drops, and the vibration due to the reaction in the strong d-c calutron magnetic field provided additional problems.

When the calutron ion source was operated at 30 or 40 kv above ground potential, the filament and arc-supply transformers also served as insulating transformers since the entire high voltage was impressed between primary and secondary. In this case the transformers were provided with extensive double-shielding between windings so that transients due to tank sparking could be confined to the shielding paths. Sometimes separate insulating transformers were used to provide greater flexibility.

8. ARC-REGULATOR STABILITY

Most of the stability problems of the arc regulators discussed above can be most easily attacked with the aid of feedback-amplifier theory, as indicated in Chap. 1. The same criterion may be used whether the regulating loop contains only d-c components or a combination of d-c and a-c components. The transfer factor must provide adequate stability in the operating range and undergo sufficient attenuation outside this range to reduce its value to less than unity before 180-deg phase shift is reached. Since the operating range
extends to zero frequency, this requires control of the high-frequency attenuation only. The thermal time constant of the ion-source filament, which is determined by other design considerations, is allowed to remain the longest time constant of the regulating loop since this arrangement provides the most rapid regulating action. The magnitude and distribution of the remaining time constants then determine the manner in which the transfer factor falls off at higher frequencies and also determine the phase-shift characteristic since it can be assumed that only minimum phase-shift networks are involved and that Bode's relationships between phase and attenuation are applicable.

Adequate phase and amplitude margins are particularly important in this application of negative feedback since a variety of filaments are likely to be used in the ion generator, each with a different thermal time constant. The regulator must provide the required stability when used with even the shortest filament time constant. Absence of loop oscillations does not mean sufficient stability since even an unstable condition below the oscillation point would give poor transient response, allowing a system disturbance to die out too slowly.

In the application of the above principles it was not usually practical to make a mathematical analysis of the complete system. However, mathematical investigations were made at various times of the transmission characteristics of the filaments, saturable reactors, signal-rectifier filters, etc. These calculations were supplemented with experimental data on portions of the loop. Quite often it was sufficient to make all the time constants within the designer's control as small as practicable, although of course this did not give the optimum attenuation characteristic.

In the early development work the required d-c gain could usually be obtained quite easily with a number of different regulating systems. However, later on when closer arc-current regulation was desired, some of the regulating systems that had a large inherent time constant, such as a saturable reactor, became less desirable because it was difficult to maintain the required stability with greatly increased gain. It was then that systems with smaller inherent time constants, such as those using inverse-parallel thyratrons, offered the greatest advantage since a higher degree of regulation could be obtained without instability. This was especially true when the additional time constants of the arc-voltage supply and signal rectifier were added to the regulating loop, as was done in the later systems, although, of course, these time constants were kept to a minimum by careful design.
Although the foregoing discussion assumes a simple single-path feedback loop, the operation is actually modified by the phenomenon of cathode bombardment and activation by the arc. Positive ions of the arc are accelerated by the arc voltage and impinge upon the negative cathode, giving up their energy as heat and raising the cathode temperature and hence its emission. In addition, a cathode activation process takes place that greatly increases the electron-emission characteristics of the cathode. Thus both of these tend to increase the cathode emission above that due to a direct change in cathode heating power. This has the effect of an auxiliary regenerative-feedback path since a signal is fed to the filament that is dependent upon the arc power. In the absence of an arc regulator this phenomenon is apt to cause instability if the feedback power is large compared to the normal cathode heating power. An increase in filament current would increase the arc current, which in turn would increase the filament temperature still further by positive ion bombardment. This, in conjunction with the increased cathode activation, would further raise the cathode emission and a runaway condition would obtain in which the cathode and arc currents are limited only by the power capabilities of the equipment. In the early development work this condition was prevented by making the total cathode power more than twice that transmitted to the cathode from the arc. By thus reducing the fraction of cathode heating supplied by the arc, this regenerative effect could be reduced enough to avoid the runaway condition. Also, a large arc series resistor was used between the arc and its power supply. This reduced the regenerative effect and, if the resistor and arc voltage were chosen properly, would even result in an over-all degenerative action. In other words, the arc power could be made to decrease as the arc current increased. This is shown by the idealized curves of Fig. 3.14. Adequate stability was obtained by these methods, but they were undesirable because of the large waste of power, both in the series resistor and to a smaller extent in the oversized cathodes used. Later in the development the use of the deep collimating slots greatly reduced the effect of the arc upon the cathode or filament. This, together with the improved regulator systems, reduced this arc-feedback effect and allowed the use of a smaller cathode and a small arc series resistor. This in turn allowed smaller filament and arc supplies, resulting in substantial equipment and power savings.

In addition to the cathode heating and activation phenomena discussed above, the arc-regulating loop was somewhat further complicated by the effects of the negative resistance characteristics of the arc. This, in conjunction with the inductance and capacitance elements of the
arc-supply filter, would sometimes form a separate oscillating circuit within the main regulating loop. This effect was reduced to a minimum by nullifying the negative resistance with a relatively small arc series resistance and by using a low L to C ratio in the arc-supply filter. Tuned filters, resonating at the principal ripple frequency, were also effective since they required only a small inductance.

PART II. TEMPERATURE CONTROL

The question of temperature regulation and control in the calutron ion sources was concerned chiefly with maintaining the optimum temperature at the charge reservoir and at the arc chamber. Most of the development work involved the thermodynamic problems encountered, such as the proper location of the temperature-measuring elements and the heater units, the elimination of unduly long thermal time constants, and the control of heat flow between various parts of the ion source. The electrical equipment and circuits used were, for the most part, quite conventional.

9. BRIDGE-CONTROLLED ON-OFF SYSTEM

A typical temperature-regulating system is illustrated in Fig. 3.15. It was usually sufficient to regulate automatically only the charge-reservoir temperature and use manual control of the arc-chamber temperature. Hence only one regulating circuit is shown. It incorporated a commercially built self-balancing bridge, such as a Foxboro or Micromax, actuated by the charge-reservoir thermocouple. A pair of control contacts, controlled by the bridge, short-circuited or unshort-circuited a dropping resistor in the heater circuit, depending upon whether the temperature was above or below the desired value. This on-off instrument provided a simpler arrangement than a continuously variable control and was usually adequate for this regulating job. The inherent cyclic fluctuations were filtered out by the thermal characteristics of the ion source, which provided a low-pass filter between the heater and the charge reservoir. Hot ion-source units were accommodated by isolating the circuit with insulating transformers and floating the equipment at source potential, as shown in Fig. 3.15. The primary variac in the heater circuit allowed manual adjustment of the heater range.

The arc-chamber heater was also isolated with an insulating transformer and manually controlled with a primary variac. The temperature was read directly on a calibrated meter energized from the arc-chamber thermocouple.
The effect of high-voltage transients, which were due to sparking in the calutron tank, was minimized by the use of adequate shielding, doubly shielded insulating transformers, and high-frequency filter networks.

10. CIRCUIT VARIATIONS

Although this regulating system was usually employed as described above, various adaptations were also used throughout the calutron development work. For instance, in some units both the charge reservoir and the arc temperatures were regulated, using separate self-balancing bridge circuits. On other units, both temperatures were controlled by the same regulator, with provision for manual adjustment of the difference temperature.

11. ELECTRONIC ON-OFF SYSTEM

Another regulating system of the on-off type is illustrated in Fig. 3.16. In this arrangement the thermocouple voltage was converted to alternating current by a synchronized interrupter and then amplified by a stabilized high-gain a-c amplifier. The output controlled a thyatron-relay unit, which in turn controlled the heater power by short-circuiting or unshort-circuiting a series resistor in the heater primary circuit. This arrangement could provide a closer degree of regulation than the self-balancing bridge system since a very small differential between on and off temperatures could be utilized. In addition, it provided greater flexibility because the a-c temperature signal could be transferred by means of an insulating transformer to the relay unit, thus allowing the latter to be operated at ground potential. This regulator, however, was superseded by a commercially built reactrol unit before it had been put into extensive operation.

12. THERMOHM-REACTROL SYSTEM

The reactrol unit outlined in Fig. 3.17 also provided a grounded control unit for operating convenience. It differed from the previous system in that, within limits, the power delivered to the heater was a function of the temperature deviation. The temperature to be controlled was measured by a thermohm unit which was essentially a resistor with a high temperature-resistance coefficient. This was coupled into one leg of an a-c self-balancing bridge circuit by a spe-
cial insulating transformer, which spanned the 30 or 40 kv between the ion source and ground. The bridge unbalance signal was amplified and impressed on a thyatron control circuit, which in turn actuated a saturable reactor in the heater primary circuit. Because of the high gain of the system, the preciseness of regulation was limited only by the stability requirements. Usually the thermal characteristics between the heater and thermohm necessitated running at reduced gain to eliminate overshoot and oscillations of the controlled temperature.

13. THERMOCOUPLE-REACTROL SYSTEM

An adaptation of the reactrol system was developed for experimental hot sources equipped with thermocouples instead of thermohms. Since the d-c thermocouple signal could not be coupled to ground with a transformer, it was used to actuate an auxiliary self-balancing bridge at the source potential, as shown in Fig. 3.18. The a-c unbalance signal from the bridge was then coupled to ground potential through an insulating transformer and used in place of the original bridge signal to control the amplifier of the conventional reactrol system. Of course with units employing grounded sources the thermocouple could be coupled directly into the original self-balancing bridge, and the auxiliary bridge could be dispensed with.

![Fig. 3.1 — Diode-type regulating system using magnetron arrangement.](image-url)
Fig. 3.2—Simplified diode-type regulating system.

Fig. 3.3—Diode-type regulating system using a standard voltage.
Fig. 3.4 — Mutually heated double cathode.

Fig. 3.5 — The d-c triode-type regulating system.
Fig. 3.6—Arc-snuffer circuit.

Fig. 3.7—Saturable-reactor system of regulation.
Fig. 3.8 — Thyratron-reactor system.
Fig. 3.9—Inverse-parallel thyratron system.
Fig. 3.10—Arc-regulator system for directly heated filaments using thyratron control.
Fig. 3.11—Arc regulator for directly heated filaments using saturable-reactor control.
Fig. 3.12—Arc regulator with indirectly heated filaments using grid control.
Fig. 3.13—Arc regulator using the calutron accelerating-voltage supply as a source of cathode bombardment voltage.

Fig. 3.14—Effect of degenerative action on power requirements.
Fig. 3.15—Typical charge-heater and arc-heater circuits with self-balancing bridge.
Fig. 3.16—Electronic on-off temperature-regulator circuit.

Fig. 3.17—Reactrol-thermohm circuit.

Fig. 3.18—Reactrol-thermocouple circuit.
Chapter 4

MAGNET REGULATORS

By K. MacLeish

1. INTRODUCTION

For operation of the calutron at maximum efficiency, it is essential that the correct relationship between accelerating voltage and magnetic field strength be maintained to a high degree of accuracy. A fluctuating magnetic field can be tolerated if each high-voltage supply is equipped with a device (such as the beam sniffer described in Chap. 5 of this volume) which will continuously vary the accelerating voltage in such a way as to hold the beam focused on the collector. In practice, however, it has been found desirable to provide regulating equipment designed to keep the variations in magnetic field as small as possible, preferably less than 1 part in 5000. Since the reluctance of the magnetic flux path does not vary appreciably with time, a steady magnetic field is obtained if the current supplied to the exciting coils of the magnet is held constant. The problem of magnetic field regulation thus reduces to that of providing a suitable constant-current supply.

From the point of view of current control, a calutron magnet is simply a large iron-core inductor possessing a high ratio of inductance to d-c resistance. Since most of the magnetic reluctance is due to the air gaps, the magnetic field varies almost linearly with the current in the coils. Various electrical properties of typical prototype and plant magnets are given in Table 4.1. The time-constant values, which are the times required for the magnet coil current to rise to 63 per cent of its final value after voltage is applied, are of importance in determining the stability of the current-regulating system.

Excitation power is supplied by a conventional motor-driven d-c generator, or in some cases by two or more generators in series or parallel. Some early consideration was given to the possibility of using polyphase mercury-arc rectifiers instead of rotating machines, but
limited time schedules and the availability of suitable motor-generator sets dictated the use of the latter in all installations.

If a fixed voltage were applied to the field circuit of the generator, the magnet current would be subject to fluctuations of several per cent arising from such factors as the change of magnet and generator winding resistances with temperature. The regulator system operates by continuously measuring the magnet current and automatically adjusting the generator field excitation as required to hold a constant load current.

2. BASIC PRINCIPLES

Although differing widely in circuit details, all of the commonly used systems of magnet-current regulation are represented in principle by the diagram of Fig. 4.1. The current to be regulated passes through a shunt connected in series with the magnet windings. This is simply a resistor designed to have a constant resistance under load; usually an ordinary ammeter shunt of appropriate current rating is used. The voltage drop in the shunt is compared by series opposition with a reference voltage which can be adjusted to an arbitrary constant value. The algebraic sum of the shunt voltage and the reference voltage appears across the input terminals, A and B, of a high-gain voltage amplifier. The output of the voltage amplifier controls a power amplifier of sufficient capacitance to provide excitation for the shunt field of the main generator. The polarity of the connections is such that an increase in magnet current (and hence in the voltage across the shunt) causes a reduction in the generator field excitation.

It is readily apparent that this type of circuit tends to hold the magnet current at a value such that the shunt voltage is, on the average, nearly equal to the reference voltage. Because of time lags in the response of the various circuit elements, however, a sudden disturbance occurring anywhere in the system will cause the magnet current to differ transiently from the equilibrium value. Following a disturbance such as a sudden change in line voltage, the current goes through a series of damped oscillations before returning to the regulated value. The speed with which the oscillations are damped out governs the precision with which the current can be held constant over short periods of time. The long-period stability of the average current, on the other hand, depends on the ability of the regulator system to counteract the effects of slow changes such as variations in the resistance of the magnet windings with temperature or drift in the characteristics of the regulator amplifiers themselves.
Analysis of the short- and long-period behavior of the system is most easily carried out by applying the theory of feedback amplifiers. The generator is regarded as an amplifying element having a certain voltage gain (change in armature voltage divided by change in field voltage), and the magnet windings and shunt in series are regarded as a voltage divider by which a fraction of the generator output is fed back to the input terminals of the sequence of amplifiers. The system is characterized by a loop gain which is the product of the voltage gains in the voltage amplifier, power amplifier, generator, and voltage divider. The gain in the voltage divider is of course less than unity, being the ratio of the resistance of the shunt to the total series impedance of the shunt and the magnet windings. The loop gain has its maximum value at zero frequency, that is, at frequencies so low that attenuation in the system is negligible. At higher frequencies the loop gain falls off with increasing frequency. Application of feedback-amplifier theory shows that the long-period stability of the current is related to the loop gain at zero frequency, whereas the response of the current to transient disturbances depends on the manner in which the loop gain varies with frequency. By analyzing the problem from this standpoint, it is possible to derive two simple rules on which the design of a regulating system can be based.

The long-period stability may be obtained by the following rule, which applies when the zero-frequency loop gain is large compared to unity: Changes of current that occur with feedback are equal to the changes that would occur without feedback divided by the loop gain at zero frequency. In applying this rule, the phrase "without feedback" means with the feedback loop open, that is, with the connection from point A (Fig. 4.1) to the shunt removed and a steady voltage of appropriate magnitude connected between points A and B. For example, suppose that the natural stability of the amplifiers, generator, and magnet are such that over a given period of time, with the feedback loop open, the current would drift by 5 per cent. To reduce the drift to 0.02 per cent will then require a zero-frequency loop gain of 250. Since one of the factors in the loop gain is the ratio of the resistance of the shunt to that of the magnet windings and since this ratio is usually of the order of $10^{-4}$, the voltage gain from the input of the voltage amplifier to the armature terminals of the generator must be about 2,500,000. Thus, with the feedback loop open, changing the voltage between A and B by 10 $\mu$V causes the generator armature voltage to change by 25 volts. The attainment of reasonable stability in a d-c amplifier having this much gain is one of the problems of magnet-current-regulator design.
To estimate the transient response and short-period stability of the system, it is necessary to consider the behavior of the loop gain at frequencies other than zero. In a magnet regulator system, it is convenient to express the variation of gain with frequency in terms of the time constants of the various elements of the loop. The longest time constants are usually those of the magnet, the generator, and the power amplifier. If all other time constants are negligible in comparison with these three, the following rule applies: For rapid damping of transient oscillations, the loop gain at zero frequency must be less than the ratio of the largest to the smallest time constant. The presence of additional time constants in the loop will further limit the loop gain that can be used. With a given combination of three or more simple time constants, it is always possible for the system to exhibit spontaneous oscillations of large amplitude if the loop gain is made sufficiently high.

In order to use the high loop gain required for long-period stability and at the same time to avoid oscillation or poor transient response, an attempt is made to select in accordance with the above rule. The time constants of the generator and of the magnet are generally not at the disposal of the regulator designer. Hence it is desirable that the voltage and power amplifiers be made as fast-acting as possible. If, however, the power amplifier is a rotating machine of several kilowatts capacity, as is often the case, its time constant may be so large that the loop gain permitted by the above rule is not sufficient to provide the desired long-period stability. In such cases it is possible to increase the permissible loop gain through the use of a damping circuit which improves the transient response by effectively modifying the frequency characteristic of the magnet-shunt combination. Various methods of damping will be described in connection with the regulator circuits with which they have been employed.

3. SURVEY OF REGULATOR TYPES

The various regulating circuits that have been used with calutron magnets all follow the basic scheme of Fig. 4.1 but differ in the details of the voltage amplifier, power amplifier, reference voltage source, and damping circuits. For purposes of comparison, a brief survey of the salient features of several regulator types is given here, arranged in historical sequence. The details of particular installations are described in Sec. 4.

3.1 Early Galvanometer-type Regulator. The first regulator used with the calutron process was an adaptation of a circuit that had served for several years for regulation of the current of a cyclotron magnet. This regulator was used with the 184-in. Berkeley magnet at the start of operations. The input stage of the voltage amplifier employed a
suspension-type galvanometer, deflections of which were translated into voltage changes by means of a photoelectric coupling system. Further amplification was provided by a direct-coupled d-c amplifier. The power amplifier was a 6-kw self-excited d-c generator controlled by vacuum tubes in series with the shunt field. The main generator was a 350-kw compound-wound machine with the series field disconnected. Reference voltage was furnished by a 1½-volt dry cell and a resistance network. Transient damping was derived principally from the voltage induced by change of magnetic flux in a loop of wire encircling one leg of the magnet yoke.

3.2 Converter-type Regulators. This type was developed at the Radiation Laboratory to overcome some of the difficulties of the earlier regulator by elimination of the galvanometer. It was used first with the 37-in. magnet and subsequently with the 184-in. XA and XC magnets. In the voltage amplifier, the signal was first converted to an alternating voltage by a mechanically vibrating converter. After amplification by an a-c amplifier, the signal was converted back to direct current through the action of a grid-controlled rectifier. The power amplifier was usually a self-excited generator controlled by vacuum tubes in series with the shunt field, although in some instances the exciter generator was omitted and the control tubes were placed directly in the field circuit of the main generator. In the XA and XC installations, transient damping was obtained by means of a current transformer which furnished a signal proportional to the rate of change of magnet current. No damping was needed with the 37- and 184-in. magnets since these magnets, being designed for economy of power, had very large time constants. Reference voltage was again derived from a 1½-volt dry cell.

3.3 General Electric Regulators. The regulators for the plant magnets at Oak Ridge were furnished by the General Electric Company. In these regulators the voltage amplifier utilized a galvanometer-photo-cell arrangement similar in principle to the one mentioned above, except that provision was made for neutralizing the restoring torque of the galvanometer suspension by means of a small permanent magnet mounted on the moving element. For power amplification the regulator used a phase-controlled thyratron rectifier feeding the shunt field winding of an exciter generator. Stability against oscillation was obtained by purposely making the time constant of the galvanometer very large compared to that of the magnet. Since the current-measuring circuit could not detect rapid changes, a direct feedback was provided from the main generator armature to the photocell amplifier for the purpose of holding the generator voltage constant over short periods.
The reference voltage was originally furnished by a second galvanometer-photocell circuit controlled by a standard cell; this arrangement was subsequently replaced by a dry-cell circuit similar to that of the preceding regulators.

3.4 Amplidyne-type Regulators. A regulator system using an amplidyne generator was developed by the Tennessee Eastman Corporation (TEC) as a possible replacement for the galvanometer-type regulators in the Oak Ridge plant. Regulators of this type were installed in the XAX and XBX pilot-plant units and on one Beta refining plant magnet. The voltage amplifier was similar to that used in the converter-type regulators. For power amplification, advantage was taken of the low excitation power and high speed of response inherent in the amplidyne-type generator. No damping circuit was needed in the XBX installation owing to the fast response of the 2-kw amplidyne exciter. The slower response of the 15-kw machine used with the Beta magnet necessitated the use of damping in that installation. The damping circuit took the form of a resistance-capacitance network for feeding rapid changes in magnet voltage directly to the voltage amplifier.

4. CIRCUIT DETAILS

As examples of the various regulator types outlined above, three typical installations will now be described in some detail. The G.E. regulators used in the Oak Ridge plant are covered in detail elsewhere and will not be described further here.

4.1 Galvanometer-type Regulator; 184-in. Magnet. The complete circuit diagram of this installation is given in Fig. 4.2. The reference voltage, derived from dry cell B1, appears between the sliding contacts of potentiometers R2 and R6. Motion of the coarse current control R2 varies the reference voltage over a range from zero to about 70 mv, and the fine current control R1 provides an additional variation of about 2.5 per cent. In series with the reference voltage and the galvanometer coil are potentiometers R1 and R16, which inject adjustable damping voltages into the input circuit. The damping voltage across R1 comes from a single turn of wire surrounding one leg of the magnet yoke and is proportional to the time derivative of the magnetic flux. The effect of this damping voltage is similar to an anticipation of changes in magnet current, for, when the current starts to change for any reason, the voltage induced in the pickup loop affects the galvanometer before the total change in shunt voltage becomes appreciable. The damping voltage fed in from T1 and R16 is proportional to the rate of change of exciter field current. The effect of this second feedback is difficult to analyze, and it was of doubtful value in actual operation.
The position of the galvanometer armature controls the grid voltage of tube V1 through an optical coupling arrangement. A diaphragm near the light source is imaged on a type-920 twin photocell by reflection from a concave mirror attached to the armature of the galvanometer. As the armature turns, the image moves off one cathode of the photocell and on to the other. The potential of the grid of V1 depends on the fraction of the total light falling on each photocell cathode and therefore varies as the galvanometer armature turns. Changes in the grid voltage of V1 are amplified in V1 and V2 and are fed to the grids of eight type-48 tubes, V3 to V10, connected in parallel. For simplicity, only one of the eight tubes is shown in Fig. 4.2.

The 6-kw exciter generator is operated self-excited, and the variable plate resistance of tubes V3 to V10 controls the exciter output in the same manner as a field rheostat. A resistor, R17, in series with the exciter field absorbs some of the voltage in the field circuit and reduces the required power dissipation at the plates of the control tubes. The purpose of capacitor C1, paralleling the control tubes, is to absorb any voltage surges arising in the highly inductive exciter field caused by sudden changes in the grid voltage of the control tubes.

When operating properly, this regulator held the magnet current constant to about 1 part in 1000 over periods of 6 hr. Most of the operational difficulties were associated with the galvanometer system; the principal ones were as follows:

1. Sticking of galvanometer armature. To prevent overtravel of the light spot past the turn-on photocell element and consequent loss of control, it was necessary to limit the possible angular displacement of the armature. An adjustable stop was provided for this purpose. In spite of careful design of the stop (a ground tungsten point bearing against a polished glass surface), the armature would occasionally stick against the stop while the magnet current was being adjusted. The current would then rise to the maximum value permitted by the circuit, and control could be regained only by physically jarring the galvanometer.

2. Difficulty of maintenance. Breakage of the galvanometer suspension and burn-out of the light source were frequent occurrences. Readjustment of the mechanical zero of the galvanometer, a delicate operation, was required periodically.

3. Susceptibility to vibration. In spite of shock mounting, building vibrations and even loud noises would cause sufficient vibration of the galvanometer armature to affect the magnet current appreciably.

4. Slow response. The inertia of the armature acted as a time constant in the voltage amplifier circuit, seriously limiting the loop gain that could be used without oscillation.
Because of these difficulties, the use of galvanometers in calutron magnet-current regulators was abandoned at the Radiation Laboratory in favor of the converter type of circuit.

4.2 Converter-type Regulator; XA Magnet. Figure 4.3 is an elementary circuit diagram of this installation. The 350-kw main generator and 6-kw exciter were the same machines that had formerly been used with the 184-in. magnet, which was then excited by a smaller machine.

A signal of about 30 mv is obtained from the shunt in series with the grounded negative current bus. The reference voltage is provided by dry cell B2 and resistors R2 to R6 in a circuit similar to that used with the galvanometer-type regulator. A damping transformer T1 feeds in a voltage proportional to the rate of change of magnet current. Although the damping action so obtained is equivalent to that of the pick-up loop of Fig. 4.2, the current transformer is preferable here because the signal from it is less susceptible to interference from stray alternating fields. Nevertheless, a filter composed of R1 and C1 is necessary to prevent 60-cycle voltages induced in T1 from affecting the voltage amplifier. Dry cell B1 furnishes a polarizing voltage for the electrolytic capacitor C1. Transformer T1 is somewhat similar in construction to a bushing-type current transformer; it consists of a laminated iron core surrounding the current bus and wound with a few turns of No. 12 wire. An air gap in the core prevents saturation of the iron. The length of the air gap and the number of secondary turns are selected so that the mutual inductance between the current bus and the secondary winding is about 5 μh, that is, a current change of 1 amp/sec produces a secondary voltage of about 5 μv. Potentiometer R2 controls the degree of damping.

The converter and step-up transformer T2 convert the difference signal appearing between points A and B to a 60-cycle square wave. The signal is applied between the primary centertap of T2 and the moving contact of the converter. The moving contact, vibrated at line frequency by a small a-c electromagnet, touches the fixed contacts alternately and thus connects the input signal alternately across the two halves of the primary of T2. The resulting voltage induced in the secondary of T2 is a 60-cycle square wave whose amplitude is proportional to the d-c input voltage. The step-up ratio of transformer T2 is about 8 to 1. The converter and transformer are the same type used in continuous-balance potentiometers manufactured by the Brown Instrument Company.

Tubes V1 and V2 and associated circuit elements form a two-stage a-c amplifier of conventional design. Negative feedback from the plate circuit of V2 to the cathode of V1 stabilizes the gain, which is about
560 for the two stages. The amplified signal is fed to tube V3 through a linear potentiometer R14, whose function is to control the gain of the amplifier and hence the over-all loop gain of the system. The first half of tube V3 provides a further voltage gain of about 45, and the second half operates as a phase inverter by amplifying a small fraction of the plate signal of the first half of the tube. The a-c voltages at the plates of V3 are equal in magnitude but 180 deg out of phase with each other.

Tube V4 and transformer T3 are connected in a grid-controlled full-wave rectifier circuit. The d-c voltage output of this circuit is controlled by the amplitude of the a-c signals from the preceding stage. If the voltage on each grid of V4 is in phase with the voltage supplied by T3 to the corresponding plate, the tube passes current and a rectified voltage appears across capacitor C16. If, on the other hand, the grid voltages of V4 are 180 deg out of phase with the corresponding plate voltages and are sufficiently large, V4 does not pass current on either half-cycle and the rectified voltage across C16 is zero. Either of these two conditions may prevail, depending on the polarity of the d-c input signal applied between points A and B, for a reversal in the polarity of this signal is equivalent to a 180-deg phase change in the a-c signal passing through the amplifier. The polarity of the voltage between A and B depends in turn on whether the voltage across the shunt is greater or less than the reference voltage. If the shunt voltage is within a few microvolts of the reference voltage, a partial throttling of the rectifier occurs. In this range the output of the rectifier is a smoothly varying function of the shunt voltage, and the entire circuit up to C16 acts as a high-gain d-c amplifier. The response characteristic at zero frequency with R14 at maximum gain setting is plotted in Fig. 4.4. The maximum slope of the curve corresponds to a voltage gain of $2.5 \times 10^6$ between the shunt and capacitor C16.

The rectified voltage from V4 is reduced by a factor of 2 by resistors R27 and R28 and applied to the grids of four paralleled 6Y6G beam tubes, V5 to V8, only one of which is shown in the diagram. The voltage reduction is used in order to make the maximum voltage from V4 just sufficient to cut off the 6Y6G tubes, thus causing the normal operating point of the voltage amplifier to lie on the steep part of the curve of Fig. 4.4. Filter capacitors C16 and C17 serve to remove some of the 120-cycle ripple from the voltage applied to the grids of the 6Y6G tubes. This filter determines the over-all frequency response of the amplifier, and the attenuation introduced by it has a small but nevertheless undesirable effect on the transient response of the regulator system. Small resistors in the grid, plate, and screen leads of each of the 6Y6G tubes are required for suppression of high-
frequency parasitic oscillations. The four tubes, capable of passing 0.5 amp, control the output of the 6-kw self-excited generator by acting as a variable resistance in series with the generator field. The over-all loop gain of the system, with R14 at maximum setting, is about 800.

By comparison with earlier regulators, the performance and reliability of the converter-type regulator were very gratifying. In the XA installation the current was constant to 1 part in 6000 over 20-min periods and 1 part in 3000 over 6-hr periods. The converter gave absolutely trouble-free operation and was highly insensitive to external vibration. The chief limitation on the amounts of gain and damping that could be used was set by distortion of the waveform in the amplifier caused by hum pickup in the low-level circuits. Noise arising from stray magnetic fluxes linking the input circuit had to be minimized by tightly twisting the wires and enclosing the long leads from the shunt and damping transformer in steel conduit. Care was required in the physical layout of the reference voltage circuit to avoid drift due to temperature inequalities. The stability of the dry cell used in the reference voltage circuit is discussed below in connection with the amplidyne-type regulator.

4.3 Amplidyne-type Regulator; Beta Magnet. Development of the amplidyne-type regulator by TEC was prompted by persistent difficulties with the galvanometer-type regulators in the plant. Previous experience with converter-type regulators at the Radiation Laboratory suggested that the same principles could be applied to the regulation of the large calutron magnets at the Oak Ridge plant. Two new factors entered into the design:

1. The magnet windings were operated ungrounded to minimize the destructive effects of accidental grounds within the coil structure. Hence the control shunt could not be kept at ground potential as before, and the input circuit of the regulator had to operate at a potential that not only differed from ground by as much as several hundred volts direct current, but also varied up and down with respect to ground at the 2000-cycle commutating frequency of the main generators.

2. Conventional d-c exciter generators large enough to supply field excitation for the main generators had a time constant that was too large in comparison with that of the magnet for satisfactory precision control.

The amplidyne-type generator, characterized by high speed of response and low control power, was therefore selected for power amplification.

Figure 4.5 shows the circuit of the Beta installation. The total input signal to the voltage amplifier is the sum of the shunt voltage, the
reference voltage, and a damping voltage appearing across R3. The damping action of the circuit composed of R1, R2, R3, C2, and C3 is similar in principle to that of the other damping circuits described previously, but in this case the signal that anticipated a change in magnet current is derived from a change in the voltage across the magnet coils, rather than from the rate of change of current or of magnetic flux. Any rapid change in generator voltage is transmitted through the large capacitor C2 and appears across the series combination of R1, R2, and R3. A small fraction of the total voltage change is fed from R3 to the amplifier input circuit, where it produces the same effect as a change in magnet current. Capacitor C3 is a by-pass or filter capacitor which prevents the high-frequency generator commutator ripple (usually about 20 volts peak to peak) from reaching R3 and interfering with the normal action of the voltage amplifier.

The reference voltage is produced by current from an Eveready No. 742 dry cell, B1, flowing through two resistors, R4 and R5, each of which can be varied in 10 steps. The two resistors have dials calibrated in hundreds and in thousands of amperes of magnet current, respectively. A fine current control, R8, providing a stepless variation of about 5 per cent, is used for small current adjustments and to compensate occasionally for slow drift in the voltage of B1. To determine the way in which the voltage of B1 varies as the cell ages, a series of tests was made with several cells. It was found that the cell voltage decreased with time at a rate depending on the current drain. The voltage dropped off most rapidly when the cell was new. The stability of a new cell could be improved by seasoning the cell before installation. For the seasoning operation, the cell was subjected to a relatively heavy current drain (5 ma) until its voltage had dropped to about 1.44 volts. This initial drop took place over a period of about six weeks. The current load was then reduced to the normal value of 0.5 ma required by the circuit of Fig. 4.5. After about two weeks at this load, the cell was placed in operation, its voltage then being about 1.46 volts. Over the succeeding twelve months, the voltage decreased at an average rate of about 0.3 per cent per month. A curve of voltage vs. time for a typical cell, covering a period of 463 days, is shown in Fig. 4.6. In addition to falling off with time, the cell voltage was subject to temporary fluctuations due to changes in ambient temperature. The temperature coefficient seemed to vary from cell to cell, but an average value was in the neighborhood of 0.02 per cent/°C.

Since the operation of the voltage amplifier circuit is similar to that already outlined for the converter-type regulator, only the main points of difference will be mentioned here.

The feedback over the first two amplifier stages is from the cathode of the second stage to the grid return of the first stage. This arrange-
ment permits by-passing the cathode resistor of V1 and eliminates some undesirable hum that was produced in the circuit of Fig. 4.3 by 60-cycle voltage pickup from the heater of V1.

Capacitor C1, connected between the control shunt and ground, causes most of the generator commutator ripple to appear on the negative current bus. Nevertheless, owing to distributed capacitance to ground in the magnet coils, an appreciable ripple voltage also exists between the control shunt and ground. It is necessary for the regulator to amplify a difference signal whose magnitude is of the order of microvolts in the presence of a 2000-cycle voltage that may be several volts. For this reason, a shielding system is provided which surrounds the low-level parts of the circuit. Both the shield and the ground bus of the amplifier are connected to earth ground through high resistances R39 and R40 to the positive current bus through a large capacitor C21. Hence the entire amplifier and shield are carried up and down with respect to ground at commutator-ripple frequency, and the low-level parts inside the shield are unaffected by the ripple voltage. If the shield were omitted or connected directly to earth ground, the interference in the amplifier from commutator-ripple would be several hundred times as strong as the control signal being amplified. The reference circuit, damping network, converter, and input transformer are at the same d-c potential as the positive current bus, but blocking capacitors C4, C5, and C21 allow the remainder of the electronic circuit to operate at ground d-c potential while at the same time carrying the commutator-ripple voltage.

Equalizing capacitors C12 and C13 are used in the phase-inverter stage for the purpose of flattening the tops of the square waves presented to the grids of V4. Their action is to emphasize the low-frequency components of the carrier signal in comparison with the higher frequencies and thus to compensate for the low-frequency attenuation introduced by coupling capacitors C4, C5, C9, C11, C14, and C15. Omission of C12 and C13 causes the carrier wave to have a sloping top.

The grid-controlled rectifier stage of Fig. 4.3 is replaced by a full-wave synchronous rectifier V4, which rectifies the carrier signal directly. The grids of V4 are driven by 60-cycle voltage from transformer T2 in synchronism with the vibration of the converter contacts. The relative polarities of the supply connections to the converter coil and to the primary of T2 are such that V4 rectifies the signal only when point A in the input circuit is positive with respect to point B. A small phase lag that exists between the motion of the converter contacts and the voltage from T2 is compensated by the networks R29-C16 and R31-C17 in the grid circuits of V4.
Because of the low control field power requirement of the 15-kw amplidyne generator, a single 6AG7 tube, V5, suffices for control. Discharge tube V6 protects V5 from inductive-voltage surges arising from abrupt changes in the amplidyne field current. The amplidyne is provided with two separate control fields. The current in the No. 1 field comes from the amplifier power supply and passes through the plate circuit of V5. The No. 2 field is connected across the amplidyne output terminals and produces a control flux that opposes the flux from the No. 1 field. The resulting negative feedback within the machine makes the control characteristic practically linear at the expense of an easily afforded reduction in the over-all power gain. The static control characteristics of the amplidyne with and without the feedback connection are plotted in Fig. 4.7. The high dynamic plate resistance of the pentode tube V5 and the negative feedback from the second control field both improve the speed of response of the amplidyne. Controlled in this manner, the machine has an effective time constant of 0.13 sec, compared with about 0.4 sec for a conventional generator of comparable power rating.

To ensure continuity of operation and facilitate maintenance, two duplicate regulating channels are provided, only one of which is shown in Fig. 4.5. A stand-by regulator and amplidyne are thus kept in readiness for operation in the event of circuit failure. While one regulator is controlling the magnet current, the stand-by regulator is left connected to the control shunt and serves to monitor the operation of the controlling regulator. This is effected by adjusting the current control knobs on the stand-by regulator to bring the reading of its output milliammeter, M1, to approximately midscale. The stand-by regulator then acts as a stable high-gain amplifier, and minute fluctuations in the magnet current are registered on meter M1. The sensitivity is such that each scale division of M1 corresponds to a change in magnet current of 1 part in 10,000. Use of the stand-by regulator for monitoring in this fashion also yields a continuous indication of the readiness of the stand-by regulator to take over control of the current. Control is switched from one regulator to the other by means of transfer relays in No. 1 field circuits of the amplidynes, the relays being so arranged that either regulator can be used with either amplidyne. Selection of the amplidyne to be used is accomplished with a switch in the main generator field circuit. The magnet must be deenergized while changing amplidynes, but the transfer from one regulator to the other can be made without appreciably disturbing the magnet current.

Figure 4.8 is a 2½-hr record of the Beta magnet current controlled by this type of regulator, operating with a loop gain of 500. The record
was made by a recording milliammeter connected in series with the output meter of the stand-by regulator. The indicated stability of the current is of the order of 1 part in 10,000. This record is typical of the short-period behavior of the regulator. Over longer periods a drift occurred, chiefly because of slow changes in the reference voltage. Slow drift, however, was of minor importance in calutron operation since other considerations necessitated periodic readjustment of the accelerating voltage of each calutron unit. Of much greater importance was the ability of this type of regulator to operate continuously for periods as long as nine months without maintenance attention.

Table 4.1—Various Electrical Properties of Typical Prototype and Plant Magnets

<table>
<thead>
<tr>
<th>Magnet</th>
<th>XA</th>
<th>Alpha I</th>
<th>Beta</th>
</tr>
</thead>
<tbody>
<tr>
<td>Operating current, amp</td>
<td>820</td>
<td>7500</td>
<td>4100</td>
</tr>
<tr>
<td>Operating voltage, volts</td>
<td>250</td>
<td>610</td>
<td>305</td>
</tr>
<tr>
<td>Power consumption, kw</td>
<td>205</td>
<td>4580</td>
<td>1250</td>
</tr>
<tr>
<td>Inductance, henries</td>
<td>2.4</td>
<td>1.13</td>
<td>1.35</td>
</tr>
<tr>
<td>D-c resistance, ohms</td>
<td>0.30</td>
<td>0.081</td>
<td>0.075</td>
</tr>
<tr>
<td>Time constant,</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>inductance/resistance, sec</td>
<td>8</td>
<td>14</td>
<td>18</td>
</tr>
</tbody>
</table>

Fig. 4.1—Basic magnet regulator system.
Fig. 4.2—Galvanometer-type regulator as used on 184-in. magnet.
Fig. 4.3—Converter-type regulator as used on XA magnet.
Fig. 4.4—Response characteristic for converter-type regulator at zero frequency with R14 at maximum gain setting.
Fig. 4.5—Amplidyne-type regulator as used on beta magnet.
Fig. 4.6—Change in voltage with time of dry cell used to supply a standard voltage for amplidyne-type regulator.
Fig. 4.7—Static control characteristic of amplidyne with and without feedback connection.
Fig. 4.8—Time record of beta magnet current controlled by amplidyne regulator. Loop gain = 500.
Chapter 5

MISCELLANEOUS ELECTRICAL CIRCUITS

By A. Guthrie

INTRODUCTION

In the course of the work on the electromagnetic separation process, a large number of circuits were devised for special purposes. In this chapter a number of these circuits are described. Because of the large variety of devices which were built and the lack of sufficient information regarding a number of them, the treatment is necessarily rather incomplete. The greatest proportion of effort was devoted to beam regulators and beam-monitoring equipment. Consequently this subject is treated in greater detail than the other subjects. Use is made in a number of sections of the terms "Q" and "R" beams, which refer to the beams of ions consisting preponderantly of (U 238)$^+$ and (U 235)$^+$ ions, respectively. Devices which were designed for use in high-vacuum measurements and magnetic measurements are not included here in view of the fact that they are considered in other volumes of this series. The treatment used is in general historical. The chapter is divided into two parts, Part I dealing with auxiliary operating equipment and Part II dealing with auxiliary measuring equipment. It is not possible to assign credit to all of the people involved in the development of these circuits. The names of some individuals appear in the text, and others are included in the list of references at the end of the chapter.

PART I. AUXILIARY OPERATING EQUIPMENT

1. BEAM REGULATORS

Early in the history of the Project, it became apparent that it would be desirable to have available some device that would vary the accel-
erating voltage over a restricted range in order to stabilize the R current for minor changes in the high-voltage magnetic field, etc. A number of such devices were constructed and used on the Project, being called "beam regulators," "beam followers," "beagles," "sniffers," etc.

The regulators which were designed made use of the R current to supply a control signal to the high-voltage regulator. The differences between the various designs lay in the methods of transmitting and applying the signal. In the early methods used, the variations in R current were converted to an r-f signal for application to the high-voltage regulator. Many of the later models also made use of this feature. This method was followed in some other units by the use of direct amplifier coupling. In the case of the so-called "differential sniffer" or "regulator," the input signal is obtained from the difference in current to the two electrodes. Examples of the various types of regulators designed in the Radiation Laboratory are discussed in the sections following.

Much of the work was done on the Project in connection with this subject, and consequently it would be difficult to list all of the personnel involved. However, most of the basic circuits were designed by the following: Robert deLiban, W. R. Baker, W. H. Nelson, K. M. MacKenzie, and L. J. Black. Others are listed in the references at the end of the chapter.

1.1 The r-f Coupled Regulators. As noted above, the earliest regulators built on the Project made use of r-f coupling. What are believed to be the two earliest such circuits to be used are described below.

The first scheme used is shown in Fig. 5.1. This circuit worked successfully from the start and gave a better understanding of the problems connected with such a system of regulation. As shown in the figure, a standard potential of 22.5 volts is connected in series with a variable resistor and the R current (PG1 and PG2). The flow of beam current through the resistor tends to produce a voltage drop in the opposite direction to the battery voltage itself. When this drop approximately equals the battery potential, the vacuum tube V1 is then in a position to amplify any changes in the current through the resistor P1. Next the output of V1 is transmitted to the voltage standard of the high-voltage regulator (accelerating supply), and the phase is so connected that an increase in R current raises the output of the high-voltage regulator and brings the R current back to its previous level. It should be noted that V1, the amplifier referred to, is in reality a grid-modulated oscillator. The purpose of transferring beam-current variations to r-f variations was to make it possible to transmit them from the high
potential beam level down to ground, where the high-voltage regulator standard was located; i.e., some 35 kv of d-c potential had to be spanned. The frequency of the oscillator was chosen at 7 megacycles primarily because the coupling loops between transmitter and receiver were of a very convenient size at that value. Variation in the R current operating level was obtained by simply varying the magnitude of the resistor P1.

After the above circuit had been in service for some time, several points came up that led to the design of a new regulator. The most important one was that the circuit of Fig. 5.1 would not maintain the differential voltage between the R pocket and the tank faceplate at a sufficiently low value to prevent serious defocusing or blowing up of the beam. Another weak point was that the improvement in beam stabilization was only between 5 and 10 and should actually be much higher. Consequently the circuit of Fig. 5.2 was tried.

Changes in the circuit of Fig. 5.1 were confined to the transmitter and amplifier and did not affect the receiver, the same one being used as previously. To reduce collector potential variation to a permissible value (assumed to be 1 volt at the time), an additional amplifier was incorporated. This amplifier is unique in its function in that it completely eliminates all nonlinear tube characteristics normally encountered in ordinary amplifier schemes. By means of this circuit there is 100 per cent unity transfer of the R beam current to the plate resistor R1 of the tube V1 (see Fig. 5.2). In this unity current stage the beam variations are amplified by the effective amplification of the tubes (V1 is a high-μ triode) and then used to modulate the same sort of r-f oscillator as was used in the scheme of Fig. 5.1. In fact, from this oscillator on, the remainder of the circuit is exactly the same as in the first case. It should also be noted that a control (P2) was added to allow adjustment of collector voltage to any desired value, either plus or minus or zero, the latter, of course, being the preferred point. In service this regulator performed very well. One estimate placed the equivalent stabilization that it reflected into the high-voltage acceleration supply at 100,000.

The detector unit of an r-f coupled high-voltage regulator, which was designed at a later date than the above schemes, will be described briefly before this section is concluded. This unit included some additional features although the basic principle is the same.

Referring to Fig. 5.3, it is seen that the input signal is received through the r-f cable and inductively coupled from inductance L2 to the tuned circuit comprising inductance L1 and capacitor C2. The r-f voltage of this tuned circuit is rectified by means of vacuum tube V2 in order to be applied between grid and filament of the main regulator tube.
Vacuum tube V1 is arranged to limit the grid of the main regulator tube to negative voltages. In particular, this is important during transients as well as when the high voltage is first turned on. Because of the capacitative coupling between plate and grid of the main regulator tube, the grid may go quite positive whenever plate voltage is first applied. Vacuum tube V1 acts as a limiter to prevent the grid from actually going positive under such conditions.

Vacuum tube V3 is a voltage-regulator tube arranged between grid and filament of the main regulator tube to limit the negative grid swing of the main regulator tube. Should the negative bias on the grid of the main regulator tube become excessive, the power dissipated in the main regulator tube could easily exceed the rating of the tube. To prevent this, the voltage-regulator tube V3 will fire before the negative grid bias on the main regulator tube becomes excessive. Unfortunately it is difficult to extinguish vacuum tube V3 once it is fired without making major adjustments in the circuit. For this reason this feature was not retained.

It will be noted in all the above circuits that the final output signal which is applied to the standard voltage unit of the high-voltage regulator is d-c voltage, obtained by rectifying the r-f signal. Also, the use of a constant-voltage input on the collector electrode is to be noted.

1.2 Direct-coupled Regulators. The first such regulator was designed, following the schemes of Figs. 5.1 and 5.2, with a view toward simplifying controls. The basic elements of the circuit are shown in Fig. 5.4. The design was guided by the following specifications for the regulator:

1. It should be adjustable for collection currents of 1 to 5 ma.
2. It should be capable of holding the R current to within a few tenths of 1 ma over its range.
3. It should inject less than 0.1-volt drop in the R circuit under collecting conditions.
4. It should have a single adjustment to accommodate various collecting currents.
5. It should have an indicating device to advise the operator as to where in its range it is operating.

The circuit decided upon uses a two-stage degenerative input loop, with a resulting input impedance of approximately 2 ohms (see Fig. 5.4). This feeds a direct-coupled amplifying stage, which in turn feeds an output stage that acts as a variable shunt resistor across a portion of the high-voltage bleeder, varying the sample voltage to the high-voltage regulator.
A degenerative input circuit is used (in preference to a low series resistor in the R lead followed by a high-gain direct-coupled amplifier) because:

1. The combination of higher input level and lower over-all voltage amplification renders the sniffer less susceptible to line-voltage changes and tube drift.
2. It eliminates the necessity of a low-voltage battery standard.
3. It offers the possibility of inserting testing devices in the R lead directly (recording milliammeter, etc.) without exceeding the voltage drop and without tapping the sniffer amplifier.

The particular form of degenerative circuit used obtains the required results with a minimum of tubes. An alternate type of circuit, in which the R current is passed through a tube whose plate resistance varies inversely with the R current, was discarded.

An output circuit that varies the high-voltage divider resistance is used instead of one that injects a variable voltage because it eliminates the necessity of high-voltage power supply in the sniffer to accommodate the probable required maximum output of 600 to 1000 volts. The unbalance introduced in the capacitance split across the high-voltage divider is not appreciable.

The beam-switching arrangement was included for flexibility. The off position gives a constant output voltage equivalent to that at mid-range. This enables the operator to turn on the sniffer after the proper collecting conditions are obtained, without readjusting the high voltage.

With the sniffer output connected in series with the high-voltage divider, there exists the possibility of oscillation occurring around the combined loop of sniffer, high-voltage regulator, and load. The gain of that portion of the loop from sniffer output to regulator output can be considered unity, so that the gain from regulator output to sniffer output represents the total gain. This will depend primarily on the sniffer amplification and the beam shape and may be anything from 0 to approximately 500 for a good beam. The regulator loop alone includes an attenuating capacitance (minimum of tank capacitance and leads) that starts cutting off the loop somewhere above 360 cps, to prevent oscillation and yet allow degeneration of rectifier ripple. If this attenuation is to remain the controlling one in preventing oscillation of the combined loop, the sniffer circuit must introduce negligible phase shift throughout the pass-band. The alternative is to put the primary attenuation in the sniffer circuit, making its frequency response low enough to preclude the possibility of the regulator introducing appreciable phase shift in the pass-band. This method was
decided upon. By increasing the sniffer time constant, its function approaches that of a manual adjustment of the high voltage. Even if a wide-band sniffer were thought desirable from the oscillation prevention standpoint, it appears undesirable because of the existence of hash on the beams.

1.3 Differential Regulator. It will be recalled that this type of regulator derives its name from the fact that it can obtain its input signal from a difference in two currents. Normally the design is such as to allow the use of one side of the beam or from the difference in current to two electrodes. The first regulator which operated in this manner was not desired primarily for this purpose. However, since the unit operated satisfactorily in the 37-in. magnet for about a year and a half, it will be described briefly.

The circuit is shown in Fig. 5.5. The most unusual feature of the regulator is the use of the Brown converter T1 in the upper left-hand portion of the figure. This converter takes a small d-c input signal, changes it to an a-c signal which is amplified by means of a balanced amplifier using vacuum tube V1, and further amplifies and detects it by means of tube V2. This method of amplification was not used in any of the other beam regulators. The differential feature appears in the use of the switches D1 and D2 in the left-hand portion of the drawing. This arrangement makes it possible to obtain the input signal from the difference between the currents to two electrodes.

The first unit designed specifically as a differential regulator is shown in Fig. 5.6. In the upper left-hand portion of this figure, selector switch D2 determines the type of operation. By turning this one switch, the beam sniffer obtains its input signal either from the lower side of the beam, from the upper side of the beam, or from the difference in current from the two electrodes. The current from F1 produces the input voltage to the grid circuit of vacuum tube V1 while the current from F2 produces the input voltage to the grid circuit of vacuum tube V2. Vacuum tubes V1 and V2 have a common cathode resistance consisting of resistors R7 and R8, so that any signal into the grid circuit of vacuum tube V2 is reflected through the cathode circuit to vacuum tube V1. The output of vacuum tube V1 is fed to vacuum tube V3, which provides the output.

1.4 Multiple Beam Regulators. Figure 5.7 shows the first Multi F Pilot, that is, the first beam regulator for testing two or more beams at once. Selector switches D1 to D4, inclusive, are arranged to select one or more of the various R currents. These currents produce voltages which are applied to the amplifier tubes V1 to V4, inclusive. The output of these four tubes is combined by means of vacuum tubes V5 and V6 in such a way that the highest R current is
the ruling factor in determining the output of the diode tubes V5 and V6 and consequently the input to vacuum tube V11. In other words, the highest R current produces the largest voltage drop across one of the resistors R1 to R4, inclusive, which causes the smallest plate current in one of the vacuum tubes V1 to V4, inclusive, which in turn causes the smallest voltage drop in one of the resistors R8 to R5, inclusive. This, of course, will cause one of the four anodes in the two duodiode tubes V5 and V6 to be most positive. Whichever one is most positive will cause plate current to flow through resistors R10, R11, R12, and R32. The resultant voltage drop in these resistors will be such that no current can flow in the plate circuits of the other three anodes of vacuum tubes V5 and V6. In this way a signal is obtained across resistor R12 and a portion of potentiometer R11, which is dependent upon the largest R current.

Beyond this point, the operation of this beam regulator is similar to that of single beam regulators. This circuit has the advantage that it compares the currents in the various R pockets and regulates according to the best one automatically.

In case of tank sparking, the Q beam sweeps across the R pocket so that the current in the R pocket increases markedly for a short period of time. This occurrence would introduce large transients into the beam-regulator circuit, which would cause instability and might tend to overload portions of the regulator. To overcome this difficulty, overcurrent circuits were incorporated in later versions of the multiple beam regulators.

1.5 Sniffer Alarm. The purpose of a sniffer alarm is to warn operators when the beam sniffer has been out of range for an appreciable period of time. A sniffer alarm circuit which was used with satisfactory results on one of the XA units is described here.4

The general principle of operation is the following: Two diodes are connected through high-resistance potentiometers to the output of the sniffer. One of the diodes is inverted with respect to the other, and both of them have suitable bias applied to prevent conduction until the desired voltage is reached. When the conduction point is reached, current flows, developing voltage across the diode load resistances which are used to bias a relay tube operating the alarm.

For a detailed explanation of the circuit operation, see Fig. 5.8. This shows, stripped to bare essentials, the part of the circuit which indicates on overvoltage. Figure 5.9 shows the essentials of the under-voltage alarm. The complete circuit combines both of these into a single unit.
Taking Fig. 5.8 first, a high-resistance potentiometer is placed across the sniffer output voltage and returned to a source of negative voltage of the order of 75 volts. This source is supplied by a VR-75 tube. Along this potentiometer are available voltages all the way from whatever positive value the sniffer output is at the moment to −75 volts. If this potentiometer is set so that the tap is at some potential below ground, the diode plate will be negative and will not conduct. Hence, no voltage will be developed across RD in the diode cathode, and the grid of the relay tube will have only the bias provided by RK. If the sniffer output voltage is now made more positive, the voltage at the tap on the potentiometer will tend to go positive. Eventually the place where the diode will begin to conduct is reached. The drop across RD will make the grid of the relay tube go positive, closing the relay and setting off the alarm. This control may be made quite sensitive by adjusting RK so that the relay tube draws almost enough current to close the relay, requiring only a small increment of positive voltage on the grid to set it off.

Figure 5.9 represents the low-voltage alarm. This is a similar arrangement, except with the polarities reversed and with RD in the plate circuit of the diode rather than in the cathode circuit as before. If the potentiometer is set so that the tap is some value more positive than ground, the diode will not conduct. If the sniffer output voltage falls, the tap will go below ground potential, and the diode plate will be positive with respect to its cathode and will conduct. This produces a drop across RD which biases the second section of the relay tube negative, causing a drop in its current. The cathode of this tube goes negative, taking the cathode of the other section negative also. Since the grid of the first section is effectively at a constant potential for this condition, a negative swing of its cathode is effectively a change of the grid voltage in a positive direction, raising its plate current and again closing the relay. This somewhat complicated arrangement results in using the second section of the 6SN7 as a means of shifting the phase of the input signal 180 deg. It is similar to that used in the differential sniffer.

Since in practice a potentiometer is not needed to cover the entire range of voltage from sniffer output to −75 volts, a lower resistance potentiometer built with fixed resistances is used. This gives a smoother control but limits the range over which the alarm can be set to function. Calculation of these resistances is relatively simple, and the ones in the unit were set to allow adjustment for the overvoltage alarm from 300 to 487 volts and for the undervoltage alarm from 325 to 125 volts. These resistances can be changed if other ranges are required.
2. BEAM-MONITORING AND SLUGGER CIRCUITS

In the preceding section devices were described which stabilize the \( R \) current for small changes in high voltages, arcs, etc., by using this current as a control signal to vary the accelerating voltage over a restricted range. This section treats devices designed primarily to give information on the ion-current distribution in a beam. Equipment of this type will be treated under the headings Beam Scanners, Three-sluggers, and Six-sluggers.

2.1 Beam Scanners. These devices provide an indication of the current-density distribution across the ion beam. The earliest beam-scanner circuit is shown in Fig. 5.10. Its principle of operation is the following: An a-c voltage is introduced into the high-voltage regulator system and the effect of this introduced voltage on the collector currents is observed by means of an oscilloscope. The vertical deflecting plates of the oscilloscope receive a voltage which is proportional to the collector currents, while the horizontal deflecting plates receive a voltage which is proportional to the a-c voltage introduced into the high-voltage regulator system. Such a procedure provides a visual indication on the oscilloscope screen of the effect upon collection of a variation in the high voltage. This is equivalent to an indication of the current-density distribution across the ion beam.

The connections to the various collectors are shown at the top of Fig. 5.10. By means of a circuit-opening switch marked "selector," it was possible to observe the current to any one of the various collectors. This switch also made it possible to observe the total current to all the collectors of the desired isotope or the total current to all the collectors of the undesired isotope. In any case the current selected would pass through the switch marked "Vert. Gain" to one of a series of resistors. By choosing a suitable resistor the current selected would produce a voltage suitable for an indication on the vertical plates of the oscilloscope.

By operating the necessary relays the variable transformer \( L1 \) is energized, producing the a-c voltage which is introduced into the high-voltage regulator system. Further, a-c voltage of the same phase relationship is introduced into the horizontal plates of the oscilloscope by means of the "Hor. Gain" potentiometer near the center of the figure. Thus, when the beam scanner is in operation a certain a-c voltage is introduced into the high-voltage regulator system and at the same time applied to the horizontal plates of the oscilloscope. The effect of this change in the regulated high voltage upon the various collector currents is observed by feeding a selected voltage to the vertical plates of the oscilloscope.
A beam-scanner circuit which is an improvement over the circuit of Fig. 5.10 is shown in Fig. 5.11. Referring to this figure, it is seen that the beam current to one or more of the electrodes in the tank is selected by a connection through terminal strip TS-1 to a suitable switching arrangement. This current passes through one of the resistors R47 to R50, inclusive, depending on the setting of D5, to ground (terminal 2 of TS-1). This produces a voltage across R51, a portion of which is introduced into the vertical amplifier through the grid circuit of vacuum tube V9. In order to prevent this voltage from becoming excessively high, a thyratron tube V7 is connected as a voltage limiter in the input circuit. The output of the amplifier and phase-inverting tubes V9 and V10 is connected through a d-c voltage correcting network to the vertical plates of the cathode-ray tube.

At the same time an a-c sweep voltage has been produced by thyratron tube V3 and has been introduced into the regulated high voltage and also applied to the horizontal plates of the cathode-ray tube. Any current which flows through the Beta divider network must also flow through the inductance L1 and the thyratron V3 to get to ground (see BP1). The action of this circuit, depending on the setting of D3, is to produce an oscillation of a suitable frequency to introduce a desired a-c voltage into the high-voltage regulator system. Selecting the different capacitors from C4 to C9, inclusive, allows the frequency of this voltage to be changed. By means of a resistance-capacitance network including capacitors C1 and C2, resistors R1 and R2, and potentiometer R9, a portion of the voltage between BP1 and ground is introduced into the grid circuit of vacuum tube V1, which together with V2 feeds a signal to the horizontal plates of the cathode-ray tube. Although switch D2 is normally kept in the open position, it is sometimes used to start the oscillation of thyratron tube V3.

Tube V4 provides a blank-out circuit which prevents the trace on the screen of the cathode-ray tube from showing on the reverse sweep. On the slow cross travel of the bright spot, V4 has no effect, but on the rapid return trip it causes the bright spot to be extinguished. Capacitor C10 is used to prevent defocusing of the cathode-ray tube on the blank-out. The cathode-ray tube used had a medium-persistence screen. This type of circuit was used in the XC equipment.

Other beam scanners which were designed included various changes over the above circuit, but the principle of operation was the same. For example, a scanner used on the R-1 unit used a cathode-ray tube with a dual phosphor screen which had both short and long persistence and which could be used with a color filter to obtain whichever persistence was desired. This made it possible to cover the range of sweep frequency which was used. None of the other scanner circuits
will be described here because the principle of operation is the same as for those described above.

Apart from the purely electronic type of beam scanner, as described above, some use was made of a mechanical scanner. Such a scanner was used during the summer of 1944 in the R1 tank in conjunction with electric shims. This type of scanner consisted essentially of a plate with a narrow slot which was vibrated before an insulated plate so that the current to this plate as a function of time gave a picture of the beam-current-density distribution. Referring to Fig. 5.12, it may be seen that the movement of the flexible strips, A and A', is produced by the reaction of the magnetic field on the current flowing through the circuit consisting of resistance R, inductance L, breaker C, and the strips themselves. The breaker C allows the strips to return to their original positions, at which time they receive another impulse upon the circuit being closed. The frequency of motion of the strips is determined by the resonant frequency of the system. The inductance L provides current phase shift so as to permit proper time distribution of the energy transmitted to the strips and to obtain an oscillation of sufficient amplitude. The signal from the detector electrode B is amplified by amplifier 2 and is applied to the vertical plates of the cathode-ray tube T. After the current through R is amplified by amplifier 1, it is then fed to the horizontal plates of T. Consequently the trace on the cathode-ray screen is a graph of beam current density across a transverse section of the beam. The wire W at the R slot position in front of the plate causes a dip in the current to indicate where the R peak should be set. Instead of the vibrating system described above, a negative resistance oscillator could be used to drive the plate. This was done successfully. Although this system of beam scanning operated satisfactorily, there were a number of difficulties associated with its use. Care had to be exercised in the choice of slot size and wire size. Also, the vibrating system did not always operate as well as might be desired.

2.2 Three-sluggers. In the operation of a calutron, a separation of the light isotope (U 235) and the heavy isotope (U 238) is produced at the collector. This separation amounts to 3 mass units, or "slugs," the inches equivalent being determined by the beam radius used. In the Alpha process the low concentration of U 235 in the feed material, together with a lack of complete separation of the two beams due to beam spreading and source width, results in an undiscernible R (U 235)^+ peak. Consequently monitoring of the beams has to be done by means of the Q (U 238)^+ beam. [In the remainder of this chapter the letter R is used to mean (U 235)^+ and Q is used to mean (U 238)^+.] If two pockets are disposed 3 mass units apart, then when the Q beam
is peaked in its pocket it would be assumed that the R beam will peak in its pocket. However, owing to slight differences in beam patterns and inaccuracies in the placement of the pockets, there may be some lowering of the enrichment of the material in the R pocket. Because of these factors the three-slugger method of monitoring was devised.

In this method a single pocket can be used in the collector. The Q beam is peaked in the pocket on a collector electrode, and then the accelerating voltage is increased so as to move the Q beam focus exactly 3 mass units away from the pocket (and away from the source). This then puts the R beam peak into the pocket. It is true that this procedure results in some contamination of the R pocket with Q material, but this can be kept low by keeping the Q peaking time and the number of monitoring cycles per unit time to a minimum. Probably the first application of this method was tried in the R1 tank. This arrangement consisted of a mechanical switching device, which, when closed, shunted out a part of the voltage-dividing resistor in the accelerating-voltage supply (see Chap. 2) so as to decrease the accelerating voltage by a predetermined fraction ($\%38$). This allowed maximizing of the Q beam in the pocket, after which the switch was opened in order to return the R beam to the pocket. In later installations the whole process was made automatic with the accelerating voltage lowered periodically in order that the operator could keep the Q beam maximized. Such an automatic three-slugger will now be described in detail.

The accelerating-voltage supply is of the type where the output voltage is regulated by the IR drop across a tapped portion of a voltage divider shunting the voltage-supply output. By changing the ratio of the regulating portion to the accelerating-voltage-dividing network, the high-voltage output may be varied, e.g., if the voltage-dividing resistor is reduced by the fraction ($\%38$) without changing the value of the regulating portion, the accelerating voltage will be reduced by the same fraction. By referring to the block diagram of Fig. 5.13, it may be seen that the operating control network periodically and for short intervals causes the 3-slug resistor to short circuit, causes a voltage ripple from the ripple-injector network to be injected into the accelerating voltage, and operatively associates the memory circuit with the phase detector. During this interval the accelerating voltage is so regulated by the correction-voltage network as to cause the peak-current portion of the beam (Q) to enter the collector and impinge on the current-collector electrode. After the beam-maximizing cycle is completed, the operating control network disconnects the memory circuit from the phase detector, removes the ripple from the accelerating voltage, and increases this voltage a given fraction by connecting the
3-slug resistor into the high-voltage divider network, thus directing the desired portion of the beam into the collector pocket. The memory circuit maintains the correction voltage constant until the following beam-maximizing cycle. During the monitoring cycle, ripple voltage is impressed on the accelerating voltage, which results in a ripple in the $Q$ current to the collector electrode. If the accelerating voltage is too high for maximum beam current to the collector, an increase in voltage gives a decrease in current and a decrease in voltage gives an increase in current. In this case the accelerating-voltage ripple and the collector-current ripple are in phase opposition. When the voltage is too low for a maximum current, the voltage and current ripples are in phase. During the beam-maximizing cycle, the phase detector charges a capacitor in the memory circuit in a direction responsive only to the phase relationship between the accelerating voltage and the collector-current ripples. The direction of the capacitor charging current determines the direction in which the accelerating voltage is changed in order to maximize the beam current.

The circuit will now be examined in greater detail with the aid of Fig. 5.14. V24 and the associated network constitute an RC oscillator of adjustable frequency. The output is coupled to a pulse-forming network, including V25 and V26, which is connected as a "flip-flop" circuit. The output is then a pulse of predetermined duration. The frequency of operation of the monitor is controlled by the frequency of the RC oscillator, and the duration of operation is determined by the duration of the pulse output of the flip-flop circuit. The ripple oscillator, consisting of V1 and associated network, is coupled through a capacitor and the diode-connected triode V1b to the triode V2a. The output swing of V2a is substantially equal to the cathode bias on V2a, which is determined by the charge on C4. A portion of the output signal of V2a is inverted and amplified by V2b, and the outputs of V2a and V2b are further amplified by the pentodes V3 and V4. The outputs of these tubes are square waves in phase opposition. The output from the current-collector electrode goes to the input of a two-stage current feedback amplifier (including V10 and V11). The signal from this amplifier is passed into the phase-splitting network (including V12), the outputs of which go to the control grids of the pentodes V13 and V14. The plate voltages of V13 and V14 are square waves, modulated in phase opposition by the outputs of V3 and V4. If the inputs to V13 and V14 are in phase, respectively, with the inputs to V3 and V4, then the outputs of V13 and V14 are high-amplitude square waves in opposition. If they are in opposite phase, then square waves of relatively low amplitude are obtained. The outputs of V13 and V14 are coupled, respectively, to the control grids of the double triode V15, which is connected
as a cathode follower. Therefore the cathodes have a common cathode resistor, and the signal output of the cathode follower is substantially equal to the more positive input to the two grids of V15. Thus the output of the cathode follower is a voltage of value determined only by the phase relationship between the ripple voltage injected in the accelerating voltage and the ripple in the collector current. If the voltage is too high, its ripple is 180 deg out of phase with the collector-current ripple, and if the voltage is too low the ripples are in phase.

The output of the cathode follower is amplified by the pentode V16, the signal from which is applied through two cathode-follower stages (including V18a and V19), LP6, LP7, and the tetrode V20 to the memory capacitor C33. This capacitor is charged in a direction determined by the phase relationship of the collector current and accelerating-voltage ripples, which in turn is determined by the relative position of the maximum current path of the beam. Thus C33 charges in such a direction as to urge the maximum-current part of the beam to traverse a path directed toward the current collector. The charge on C33 passes through the impedance transforming network, which includes V18b, V21, and V22. V22 is a constant-current regulator and V18b maintains a constant plate voltage on V21. The output of this stage follows the charge on C33 and is applied in series with the voltage drop across R40 to the control grid of V8. Also, the voltage drop across R40 is a fraction of the potential of the lower end of the voltage-dividing resistance of the accelerating-voltage supply with respect to ground, being determined by R39 and R40. Moreover, ripple voltage is impressed on the control grid of V8 through R130. V8 is normally at cutoff because of the high bias on the suppressor grid. However, during the monitoring cycle the multivibrator, V25-V26, gives a positive pulse to the suppressor grid so as to drive V8 to an operating point. The output of V8 is coupled to the grid of V6 (regulating tube), and V6 is connected between the lower end of the 3-slug resistor, A, and +350 volts. The lower end of the resistor is connected through the current-regulated tetrode V5 to -1000 volts. With this arrangement, during a monitoring cycle the lower end of the high-voltage (accelerating-voltage) dividing resistor is regulated to a voltage determined by the charge on C33 (which acts as a standard voltage). Also, since an a-c voltage is applied to the grid through R130, ripple voltage is impressed on the high-voltage supply. At the end of the monitoring cycle, V8 is driven to cutoff by the negative bias on the suppressor grid and the pentode V7 takes over control. As the voltage across C33 also acts as a voltage standard and since the grid of this tube is connected to the lower end of A, the latter point is now brought and regulated to the same voltage as was the lower end of the voltage-divider network.
It is so regulated during the interval between monitoring cycles. Therefore the high voltage for maximum collector current is increased by a percentage determined by the relative importance of \( A \) with respect to the high-voltage resistance, in this case 3 to 238.

During the monitoring cycle, the triodes \( V_{17} \) are driven to cutoff by the negative pulse output of the multivibrator \( V_{25} \) and \( V_{26} \). However, at the end of the cycle these tubes conduct, the input to \( V_{17b} \) being the output for the cathode-follower stage, including \( V_{21} \), which follows the potential across \( C_{33} \). \( V_{17b} \) and the associated network constitute a cathode-follower stage feeding tubes \( V_{19} \) and driving their cathodes to the same potential as the upper part of \( C_{33} \), thus extinguishing LP6. Since \( V_{20} \) also is driven to cutoff by the negative bias on the control grid following completion of the cycle, LP7 is also extinguished, thus completely isolating \( C_{33} \). If the current to the current-collector electrode is exceedingly high for a long period, means are provided for initiating the cycle. The output of the current collector is applied to the amplifier stage \( V_{10}-V_{11} \), and to the input of the network including \( V_{23} \), which network, responsive to a given minimum current for a predetermined time, injects a pulse into \( V_{25}-V_{26} \), thus initiating the cycle. The switch \( D_{1} \), in the nonoperating position, discharges \( C_{33} \) and connects the lower end of \( A \) to ground, so that the beam can be maximized manually.

### 2.3 Six-sluggers\(^{5,6}\)

This type of device was designed to give a visual indication of the distribution of ion currents in the calutron beam. Originally the six-sluggers were designed to work automatically at set intervals of time, but this was very soon changed so that the unit operated once when a button was pushed. A unit which was the first such device used on the Project and which was designed to operate with hot sources is described in detail below.

Figure 5.15 is an elementary sketch of the sweep and interpolation circuit as used with high positive operation. Considering first the sweep portion only, the operation is the following: Diode \( V_{6} \) is normally conducting and point \( A \) is essentially at ground potential; so \( C_{2} \) is charged to 1100 volts in the polarity indicated. Tube \( V_{5} \) is normally cut off by the high negative potential on its grid.

When \( PB \) is closed to actuate the sweep, the grid of \( V_{5} \) suddenly rises to a potential equal to

\[
E_I - \frac{(E_I + E_2)R_{11}}{R_{11} + R_{10}} = 350 - \frac{750}{3} = +100 \text{ volts}
\]

This high positive grid potential causes \( V_{5} \) to conduct heavily, thus effectively tying point \( E \) to ground, causing point \( A \) to assume a
voltage determined by the charge on C2, or about -1100 volts with respect to ground. Closing PB also charges C3 to E1 = +350 volts with respect to ground. Thus, when PB is released the grid of V5 still remains positive for a length of time determined by the time constant C3(R11 + R10), as the charge on C3 gradually goes from +350 to -400 volts because of voltage E2. The charge on C2 also goes from -1100 volts to 0 by virtue of the positive current through divider R1 + R0 + R6 + R5 + R4 caused by the source voltage (R12 is normally very high). C2 would charge positively were it not for diode V6, which conducts when point A tends to go positive with respect to ground.

The sweep cycle is now complete; C2 has discharged. The grid of V5 is, meanwhile, going negative since C3 is losing its positive charge, and V5 gradually cuts off again, causing C2 again to charge to its original potential, and the sweep is ready to be repeated. Now the action of the regulator is such that the change in output voltage due to the injected sweep voltage e_s at point A is, if the ratio of the 893 plate resistance to load resistance is low,

\[
\Delta E_0 = \frac{\pm e_s}{1 + \frac{\mu A\beta}{R}}
\]

where \(\mu\) is the 893 amplification factor (36) and A is the regulator gain.

\[
\beta = \frac{1}{R1} = \frac{R0 R4 + R5 + R6}{R1 R0 + R4 + R5 + R6}, \quad R12 \gg R4 + R5 + R6
\]

Since \(\mu A\beta\) is normally large, \(E_0 = \pm e_s = -e_s\) in this case.

Thus the change in output voltage is almost exactly equal to the change in voltage at point A in both magnitude and phase, differing only by a percentage \(1/\mu A\beta \times 100\) per cent (about \(3/100\) of 1 per cent for average conditions) from the true value of voltage at point A. Since the time constant C2(R1 + R0 + R4 + R5 + R6) is very large and the voltage to which C2 is charged is small compared to the applied source (M) voltage, the rate of rise of voltage \(e_s\) is almost exactly constant for the interval involved, even though strictly speaking it is an exponential function.

Figure 5.16 is an elementary sketch of the interpolator circuit. Before the sweep is actuated, point A is very nearly at ground potential; thus tube V1 is conducting and causes a negative potential to appear
on the intensity grid G of the cathode-ray tube, blanking the trace. At this time also, points B, C, and D are, respectively, 2.5, 3.5, and 6 slugs positive with respect to ground. Tubes V1, V2, and V3, therefore, are conducting but have only a few volts positive on their grids because of the grid-current flow through the 40-megohm grid resistors.

Upon actuating the sweep, point A goes suddenly to more than 6 slugs negative with respect to ground, and hence points B, C, and D also go negative with respect to ground. V1 ceases to conduct and grid G would go positive were it not for the action of tubes V4 and V5. V4 receives the initial negative pulse through capacitor C1, and hence its plate goes positive, causing the grid of V5 to go positive by virtue of the neon-tube coupling. The plate of V5 then goes negative, causing grid G to remain negative for a short length of time determined by the time constant R1C1. It should be noted that this section of the circuit also serves to blank the tube during tank sparks or other transient conditions, thus eliminating spurious traces or "ghosts." It may be noted in Fig. 5.17 that the trace is blanked completely during the interval $t_0$ to $t_1$.

At time $t_1$, point D reaches ground potential since it is 6 slugs positive with respect to point A, and point A is now 6 slugs negative with respect to ground. The grid of V7 then goes to ground and V7 begins to conduct, causing a negative signal to appear on the grid of V1, which in turn puts a positive signal on grid G of the cathode-ray tube. This voltage causes the trace to appear at normal brilliancy, adjustable, of course, by the intensity control. This brilliancy persists for the Fig. 5.17 time interval $t_1$ to $t_2$.

At time $t_2$, point C reaches ground potential. V3 now conducts, placing a negative signal on the grid of V6. V6 then places a positive signal on grid G, which greatly intensifies the trace over the interval $t_2$ to $t_3$ in Fig. 5.17. At time $t_3$, point B reaches ground potential and V2 begins to conduct, thus placing a negative signal on grid G about equal to the positive signal of V6, thereby returning the intensity to normal. This normal brilliancy then remains for the interval $t_3$ to $t_4$ in Fig. 5.17. At time $t_4$, point A again returns to zero, V1 conducts and blanks the trace by its negative signal on grid G, and the same conditions are reached as those just preceding time $t_0$.

A horizontal amplifier is connected to point A of Fig. 5.16. This serves the purpose of converting the voltage variations at point A to current variations of the same waveform and of sufficient magnitude to be used with the horizontal deflection coils of the magnetic-deflection cathode-ray tube. This amplifier is shown in Fig. 5.18. The direct-coupling network R5, R6, and R7 establishes the operating bias
for V2, still allowing d-c transmission from V1 to V2. The output coils are connected in the cathode of V2.

To neutralize the effects of the leakage inductance of the deflecting coils, negative-feedback stabilization is used via resistor R1. Any increase in the impedance of the coils with frequency will cause the current in V2 to tend toward a lower value, thereby decreasing the drop across R1 and decreasing the voltage fed back to V1. The output of V2 therefore increases to compensate for the original change in coil impedance. Since the loop gain is high, the stabilization is very good out to about 2000 cycles, which is about 20 times the sweep frequency if the sweep were repeated continuously; therefore a faithful reproduction of the sweep in the horizontal deflecting coils is obtained. Resistors R2 and R3 also apply a positive voltage to V1 in order to start the trace from one side of the screen.

The beam current under observation is fed directly into the vertical deflection coils, which have a sensitivity of approximately 100 ma for full vertical screen deflection. No provision is made in the unit for amplification of smaller currents, such as probe currents, for example. This renders the instrument incapable of handling small currents where a small voltage drop is required. However, shunts are provided for higher values of current.

A detailed drawing of this six-slugger circuit is shown in Fig. 5.19. As noted above, this circuit was used with hot-source operation and also with no deceleration of the beam at the collector.

When a six-slugger device is to be used with cold-source units and collector deceleration of the beam, various problems arise. From Fig. 5.20 it is seen that the source is now grounded and therefore an r-f regulator is necessary because of the d-c potential difference between input and output. To use a d-c regulator here requires the 893 tube to be in the grounded side of the rectifier, which is unsatisfactory because the G drain and arc drain flow through the 893 tube.

A number of possibilities for using the type of six-slugger available (Fig. 5.19) in cold-source operation and with collector deceleration were considered. One method which appeared satisfactory is the circuit shown in Fig. 5.21. Here the sweep portion of the slugger is left as in Fig. 5.20, and the vertical input is coupled to the receiver by means of an r-f link or transformer insulated for deceleration (C) voltage. This coupling link can also be inserted at A, with some disadvantage due to the capacitance to ground of the deceleration supply, but with some advantage from surge protection on collector to C sparks and collector to C capacitance. The slugger can be used with or without deceleration with no changes required for the tran-
sition and no changes required in the slugger except the addition of the r-f link or transformer, which could be external.

The six-slugger was found to be a satisfactory method for the study of the intensity distribution in the beam. Some of the results of such studies are included in the National Nuclear Energy Series, Division I, Volume 6.

3. FLURRY AND SPARK SUPPRESSORS

During the operation of a process tank, it was found to be a common occurrence for sparks, or arc discharges, to effectively short-circuit the decelerating voltage (or C) supply. This condition was detrimental to operation, and there was the possibility of serious damage to power supplies and other associated equipment.

The first solution used was a series tube running at a low enough filament temperature to limit definitely the maximum surge current, as seen by the rectifiers, to a safe value. The action can be explained with the aid of Fig. 5.22. A high-voltage supply is connected in series with a current-limiting device and some sort of spark gap. When voltage is impressed, there is no drop across the limiter until a spark occurs. When this happens, the current rises rapidly until the entire voltage of the supply is across the limiter, and there is not enough voltage left at the gap to maintain the discharge, which then extinguishes. Immediately the current drops to zero and the voltage again appears across the gap. Now it is found that the gap breaks down much easier, owing to the fact that the gas is not entirely deionized between its points. In this theoretical case the rate of rise of voltage is limited only by the capacitance of the output capacitor and the series impedance of the power supply.

Actual oscillograms showed that this was essentially what happened in the tanks. The voltage dropped to zero, rose very rapidly, and then dropped again; this action was repeated at a high rate of speed. To prevent this, some device is required that will cut off the high voltage when a spark appears and restore it only after a definite time, which must be long enough to ensure complete clearance of the gap area of any ions formed by the discharge. The high-voltage systems and the regulators being used lent themselves well to this method since it was only necessary to put a steep negative wavefront on the grid of the 893 series regulating tube (see Chap. 2), drive it to cutoff, and hold it there for the desired time interval. The first devices for accomplishing the job did exactly that.

Early experiments showed that if the current surge is brought to zero too rapidly, exceedingly high voltages can be developed across
the various components of the rectifier system. Transients were experienced which appeared to be two or three times the d-c output voltage and caused flashovers on bushings and terminals in the cubicles. It was found that on placing an RC network in the loop, so that it was effective in slowing the rate of change of current to a reasonable value, the transient trouble disappeared.

As was pointed out earlier, a modified square wave of known duration should be applied to the grid of the series tube. A 'locked' multivibrator can be best used to supply such a waveform. Such a unit appeared desirable also from a standpoint of ease of firing. One tube in such a multivibrator is normally biased to a point below cut-off, leaving the other tube conducting. If the cutoff grid is lifted above cutoff even for a very short period of time, the unit will go through one complete cycle of operation in the usual multivibrator fashion and then return to the cutoff condition. (In a later model this means of exciting the circuit was not used because of the desirability of keeping the capacitance to ground on this grid as low as possible.)

As the maximum voltage and power output of the device is strictly limited, it is worked into a power amplifier, the plate load of which is the grid resistor of the 893 tube. A 304TL triode was used in the first model, but this was changed in the later units to an HK-257 (RCA 8001). The power requirements did not justify the use of such a large tube as the 304TL, and a pentode appeared desirable from a standpoint of frequency response. The pentode power amplifier also has an advantage in that the plate is effectively shielded from the grid by the screen. This prevents any steep wavefronts which might appear on the grid of the 893 tube owing to regulator characteristics from being coupled back into the multivibrator circuit through grid-plate capacitance in the amplifier.

The circuit for a model based on the above principles is shown in Fig. 5.23. Looking into the unit from J1, the first element encountered is C11. This is simply a d-c blocking capacitor. It is followed by a 50,000-ohm resistor (R17), which when considered in connection with the grid capacitance of the first tube serves to increase the time duration of the applied pulse, ensuring operation even on the fastest pulses. The sensitivity or threshold is controlled by means of the variable resistor R1, which forms part of a voltage divider. The other parts are the reactance of C11 and the resistance of R17.

The input tube serves as an inverter, delivering a minus signal to the screen grid of V3, which then operates the multivibrator circuit. The plate of V3 is coupled to the grid of the other multivibrator tube through a network (consisting of C7 and R7) and a neon lamp. When the plate of V3 goes positive rapidly, as it does when a signal is
applied, the steep wave is transferred to the neon lamp through the capacitor C7. The neon lamp V5 breaks down, driving the grid of V4 positive as in the usual multivibrator. Grid current is limited by means of R7, which also provides d-c coupling. The time during which the multivibrator stays in this position is determined primarily by the time constant of the grid circuit of V3, which is variable. (This variable resistor R6 provides the means of changing the time duration of the negative voltage applied to the grid of the 893 tube.)

When the multivibrator starts its return action, the plate of V3 again becomes relatively negative, and the neon lamp V5 extinguishes. Now the capacitance on the grid of V4 is made up of only the wiring capacitance and the tube capacitance, and therefore the time constant is very short. This is essential if the circuit is to be made capable of clearing itself for another operation in a minimum of time.

It may be noted that the plate resistor of V3 is in series with the grid circuit of the power amplifier V6. V3 is normally conducting, and the drop across its plate resistor represents the bias applied to the power amplifier. As this is in a negative direction, there is normally no plate current flowing in the power amplifier tube V6 and no drop across its plate resistor (which is the grid resistor of the 893 tube).

Now R16 and C10 in the grid circuit of the power amplifier V6 may be considered. As was pointed out earlier, there is danger of severe transient troubles if the current in the 893 tube is cut off too rapidly. This condition is avoided by introducing the time constant of the correct magnitude in the grid circuit of V6. The drop across the resistor R11 in the low-voltage power-supply circuit provides the cutoff bias for V1 and V3.

Figure 5.24 is a simplified wiring diagram of the high-voltage supply and regulator system with the suppressor in the circuit. The input terminal of the unit is connected to the Beta split network as shown on this diagram. Input capacitance of the unit will not noticeably affect the usual functions of the network because the reactance of the 100-μf capacitor in series with the input of the unit is very large as compared to the reactance of C1. Assuming the unit to be in the quiescent condition, the final amplifier tube V6 is drawing no plate current, and there is no drop across the grid resistor of the 893 tube except that of the regulator amplifier.

A later model of the suppressor unit, designed for use on Alpha II units, will now be described. In the case of the Alpha II system, the source units were operated at a high voltage above ground and were maintained at this voltage by the M voltage supply. The accelerating or G slits were maintained at their voltage by the G voltage supply.
In the case of the present suppressor unit, the design is such as to combine M flurry suppression with G flurry suppression. Reduction of the current in a selected arc is incorporated in the system.

From Fig. 5.25 it is seen that thyratrons V1 and V2 comprise a dual timing circuit operated by a signal from the G voltage supply. Thyratrons V3 and V4 comprise a similar circuit, which is operated by a signal from the M voltage supply. One set of contacts on relay Re-2 is used to operate a magnetic switch which inserts a fixed resistance in the primary circuit of the regulator-tube filament transformer. A second set of contacts on relay Re-2 is used to operate thyratron V2. Relay Re-3 is operated from the control circuit of the M and G rectifier supplies.

The operation of various parts of the circuit is as follows: The combination of variable resistor R1 and capacitor C1 supplies a peak alternating voltage from 56 to 0 volts with a phase advance of 69.5 to 90 deg with respect to the plate voltage on V1. Approximately one-fourth of this voltage is made available at the grid of V1 through the combination of resistors R2 and R3 and capacitor C2, one end of which may be considered as connected to ground. Capacitor C2 and resistor R4, in parallel with the series resistors R1, R2, and R3, comprise a timing circuit with a constant of 0.25 sec. Thyratron V1 and the aforementioned phasing and timing circuits comprise the G trigger circuit.

Capacitor C4 and resistor R5 comprise the release timing circuit with a maximum time constant of 5 sec. Capacitor C4 may be charged to a peak value of about 145 volts through one-half of diode V5 and voltage divider R17 and R6 when V1 is not conducting. When V1 is conducting, divider resistors R17 and R6 place the cathode of V5 at a potential more positive than the lower end of C4, preventing C4 from becoming charged. The lower end of C4 is brought to a peak potential of about +3.4 volts by the divider resistors R7 and R19 to ensure positive firing of V2. Without this positive bias, the current produced by the negative space charge of V5 would tend to keep a minimum of -1.5 volts across C4, which would prevent positive firing of most type 2050 tubes when used in position V2.

In the output circuit of V2, resistor R8 limits the current through relay Re-1, and capacitor C5 filters the pulsating current from V2 and provides sufficiently pure direct current for quiet operation of Re-1.

The combination of variable resistor R9 and capacitor C6 supplies a peak alternating voltage from 31 to 0 volts with a phase advance of 79 to 90 deg with respect to the plate voltage on V3. Approximately 43 per cent of this voltage is made available at the grid of V3 through
the combination of resistors R11 and R12 and capacitor C7. Capacitor C7 and resistor R10 in parallel with the series resistors R9, R11, and R12 comprise a timing circuit with a time constant of about 0.2 sec. Circuits connected with V4 are identical to those described for V2.

The operation of the G flurry circuit is as follows: Resistor R1 is adjusted so that a peak of +5 volts is applied to the grid of V1. Normally relays Re-2 and Re-3 will be closed when the high voltage is on the tank and both V1 and V2 are conducting.

The signal voltage to trigger V1 is obtained from a resistor of such a size that 7 volts is developed across it when G flurry current is flowing. This 7-volt signal is applied between terminals 1 and 7 of the flurry suppressor unit. Since C2 is discharged, or very nearly so, both cathode and grid will be made positive by the amount of the signal voltage (7 volts) at the instant the signal is applied. In about 0.25 sec, C2 will charge up to 7 volts, leaving a net -2 volts on the grid of V1, and V1 is extinguished at the end of the next positive half-cycle of plate voltage.

The instant V1 is extinguished, V2 is extinguished, since its plate voltage is removed, and relay Re-1 opens. Now there is a positive potential on the plate of V5, and in one or two cycles C4 will be charged to a potential of about 145 volts negative toward the grid of V2. When the flurry signal is removed, C2 discharges and V1 again fires. This brings the cathode of V2 within 10 to 15 volts (V1 tube drop plus running G signal current) of the ground-bus potential. Now V5 cannot conduct because divider resistors R17 and R6 put its cathode about 20 volts rms from the ground-bus potential. C4 discharges through R5 until its voltage approaches zero, when V2 will fire and relay Re-1 will be operated.

Control of arc current in any one of the four arcs is obtained by means of 1-megohm resistors in the filament circuits. These 1-megohm resistors are arranged so that any one of them may be connected across the contacts of relay Re-1 in the flurry suppressor unit while each of the remaining three is shorted. With Re-1 in the operating position, all arcs will operate normally. When Re-1 is non-operating, owing to a G flurry, the arc selected will have its current dropped to about one-quarter of its normal value. Since relays Re-2 and Re-3 are connected in series with V1 and V2, an M flurry or a high-voltage recycle will cause reduction of current in the selected arc.

In the case of the M flurry circuit, resistor R9 is adjusted so that a peak of +5 volts is applied to the grid of V3. The signal voltage to trigger V3 is obtained from a resistor of such size that 7 volts is
developed across it when M flurry current is flowing. This 7-volt signal is applied between terminals 5 and 7 of the flurry suppressor unit. Since C7 is discharged at the instant of application, it will take about 0.2 sec to reach a negative potential high enough to overcome the fixed positive bias on V3 and extinguish it. From this point on, operation is identical with that of V2.

Relay Re-2 and a magnetic switch, not shown in the figure, are used to control the emission of the 893 tube. The magnetic switch has normally closed contacts. With relay Re-2 in operating position, the magnetic switch is deenergized and the 893 tubes have normal emission. When relay Re-2 is dropped out by an M flurry, the magnetic switch is energized, inserting the resistors in the primary circuit of the 893 filament transformers, and the emission is reduced to a predetermined figure.

4. DECELERATION CIRCUITS

Deceleration collectors were designed to collect ions at a few hundred volts energy. In view of the fact that the operation and construction of such collectors are treated in detail in National Nuclear Energy Series, Division I, Volume 6, only enough will be said here to make the discussion of the associated circuits intelligible. The treatment given here is for hot-source operation since all deceleration collectors used were designed for this type of operation. This section is divided into three parts, which treat power supplies, circuits for rejecting ions, and beam-monitoring equipment.

A typical deceleration collector consists of the following parts: shaving vanes ("barbershop"), shield ("pigpen") with decelerating grid, and collector ("honeycomb"). Two other grids are included in the equipment, one at ground potential before the decelerating grid and one before the final collector at shield potential. The nature of these parts can be visualized by reference to Fig. 5.26, where A refers to shaving vanes, B to decelerating grid, C to shield, and D to final collector. The ion deceleration is, of course, produced by means of the source voltage. However, since the deceleration is applied in steps it is necessary to provide additional power supplies. Two such supplies are normally used, one for the shield (Ω) and the other for the collector (Ψ).

4.1 Power Supplies. The schematic diagram of Fig. 5.26 shows the connections to the various power supplies. The shaving vanes are at ground potential, and the shield and collector are at source voltage with the shield and collector voltages superimposed. It has not been deemed necessary to include details of the shield (Ω) and collector (Ψ) supplies in view of the fact that they are of standard...
design. The shield supply normally is designed to give voltages up to 5 kv with currents up to about 100 ma. The final collection of ions at the collector (or honeycomb) is carried out at a few hundred volts negative with respect to ground. A collector supply giving up to 300 volts and 100 ma is typical of the type required. The leads labeled "to source volt." in Fig. 5.26 are connections for the beam-monitoring system. The other circuits in the figures comprise a system for avoiding contamination when sparking occurs to the source. This is discussed in Sec. 4.2.

4.2 Ion-rejection Circuits. When sparking occurs between the source and ground, undesired ions reach the collector, resulting in contamination. A number of circuits were devised to avoid this difficulty. The first one to be discussed is a manually operated device, the elements of which are shown in Fig. 5.26. In this figure it is seen that the positive terminal of the collector voltage supply (V1 and associated circuit) is tied to the positive terminal of the source high-voltage supply, and the negative terminal is tied through R1 and tube V3 to the collector. The power supply, consisting of V2 and associated network, is connected across R1 through the grid-control gas tetrode V3. The negative terminal of this supply is tied to the negative terminal of the collector supply, and the positive terminal is tied to the anode of V3, the cathode of which is connected to the collector. Also V3 has a cutoff bias applied by a voltage source B1. The control grid of this tube is tied to the negative terminal of the bias supply through resistors R2 and R3 and is also connected to ground through capacitors C1 and C2.

In normal operation of the calutron, the plate of C1 is at a high positive potential. When a spark between source and ground occurs, this capacitor discharges through R3, applying a positive signal to the control grid of V3 and thus firing it. This then applies the positive potential of the power supply (V2, etc.) to the collector, raising its potential high enough to reject the ions of the beam. Alternatively, V3 may be fired by depressing the push button D1, which puts a firing signal on the control grid of V3, thereby connecting the positive terminal of the power supply to the collector. To extinguish V3 and return the unit to normal operation (after sparking has ceased), switch D2 is opened, which removes power from the supply of the rejecting network and therefore voltage from the plate of V3.

A completely automatic ion-rejecting circuit is shown in Fig. 5.27. It should be noted that reference is made in the Project literature to "electric fences." This term was used to denote ion-rejecting circuits of the type to be described here. In referring to Fig. 5.27, it is seen that the negative terminal of the fence power supply (V2 and
circuit) is connected to the negative terminal of the collector voltage supply, and its positive terminal is normally connected through the contactor and arm of relay Re-1 and through the solenoid of relay Re-2 to the anode of the gas tetrode V4. The cathode of this tetrode is tied to the collector. The control grid of V4 is tied to ground through resistor R3 and capacitor C2 and to the bias supply B1 through resistors R3 and R4. This tetrode is normally at cutoff. The cathode of tube V3 is connected to the negative side of the fence power supply; its anode goes through the solenoid of Re-1 and resistor R5 to the positive terminal of the supply. The control grid of the tube has an RC network (R1 and C1) in its circuit.

During normal operation of the calutron, the collector is connected through R2 and the collector voltage supply to the positive side of the source high-voltage supply. The circuit between the fence power supply and the collector is open since V4 is cut off. Also a high voltage is normally applied to C2, and upon sparking between source and ground this capacitor discharges through R4, thus firing V4 and completing the circuit between the fence power supply, the collector, and the source voltage. This raises the collector voltage so as to reject the beam ions. Moreover, relay Re-2 is energized, thus connecting the control grid of V3 through R1 and C1 to the cathode. It should be noted that, during normal operation, charge is accumulated by C1, so that the voltage across this capacitor is essentially equal to the voltage drop across the tapped portion of the potentiometer P1. Therefore when Re-2 is actuated, V3 is driven to cutoff and remains in this condition until sufficient charge has leaked off C1 through R1. When V3 conducts again, relay Re-1 is energized, which breaks the circuit between the positive terminal of the fence power supply and the anode of V4, thus extinguishing this tube. Also the fence power supply is disconnected from the collector, thus restoring the system to normal operating conditions. Furthermore, relay Re-2 is deenergized so that the control grid of V3 is connected through the R1 and C1 network to a positive potential in order to recharge C1.

Another type of ion-rejecting circuit is shown in Fig. 5.28. In normal calutron operation, capacitor C1 is charged to source voltage (positive) through R1. When a spark occurs between source and ground, C1 is discharged through R1 and R2. In this process, capacitor C2 in the RC network (R4 and C2) is charged and a signal is applied to the tetrodes V2, V3, and V4 through capacitor C3 and resistors R8, R9, and R10. Moreover, the charge on C2 puts a positive signal on the control grid of V5 which is amplified by this tube and V6 and applied as positive signals to the control grids of the gas tetrodes V2, V3, and V4. The positive bias on these tetrodes is maintained so long as
V5 is conducting, which time interval is determined by the charge on C2 and therefore by the time constant of the R4 and C2 network. So long as there is a positive signal on the grids of the tetrodes V2, V3, and V4, there is a substantially constant current through the potentiometer P2, and the voltage drop across the selected portion of P2 is applied in series with the collector power supply and the collector. This raises the potential of the collector sufficiently to reject beam ions. Moreover, following a sufficient leakage of charge from C2 through R4, the signal input to V5 is of such value that the signals applied to the grids of the tetrodes V2, V3, and V4 are insufficient to fire them. Thus there is no voltage drop across the potentiometer P2, so that the collector is returned to its normal operating potential.

4.3 Beam-monitoring Equipment. A type of monitoring circuit which was commonly used with deceleration collectors is shown in Fig. 5.29. The two leads, F1 and F2, go to feeler electrodes in the shaving-vane region, as indicated in Figs. 5.26, 5.27, and 5.28. The monitor consists of parallel circuits (Fig. 5.29) including the adjustable resistors P1 and P2 and a comparing circuit adapted to supply a modifying potential to the source-voltage regulator. The resistors are adjusted to set a selected ratio between the currents to the respective feelers in accordance with the desired focus of the beam. Any unbalance in the currents is utilized in the comparing circuit to apply a corresponding potential to the source-voltage regulator and to provide an appropriate adjustment of the accelerating voltage. The circuit is essentially a d-c ratiometer, such as that discussed in Sec. 9, with provision for automatically adjusting the source voltage.

The circuit described above was used extensively with one-arc sources at the Radiation Laboratory. However, in the work at Oak Ridge where two-arc sources were studied, other monitoring schemes were devised. Most of these methods involved only the simplest types of electrical circuits (metering and control). Consequently the descriptions to follow are quite brief.

Perhaps the first fairly satisfactory method of monitoring involved measurement of the current to a door between the first two grids and the honeycomb (R can). The ratio of U 238 ion current to U 235 ion current was obtained by noting the currents to the rear door before and after using a three-slugger circuit. This corresponds closely to methods used with conventional collectors. Another method involved the use of two vanes ("S" vane method) placed in the region of the barbershop; these were operated in conjunction with a three-slugger much as in the above rear-door method. Attempts were made to monitor by means of a satellite beam (U++) from one of the arcs, but not too much success was achieved owing to the necessity for mechanc-
ical readjustment of the collectors' electrodes (or "targets") used when the arc conditions changed. The current intercepted by grids placed between the first deceleration grids and the honeycomb was also used for monitoring the beam. This was not too successful because of an unexplainably large loss of beam current. A final method to be mentioned is that involving the use of a so-called "ferret" circuit. In view of the fact that this method showed considerable success and also because the electrical circuits are of interest, it will be described in more detail.

This method permitted the use of two insulating electrodes per beam in the collector. A 0-deg defining vane was attached electrically to the jail (or pigpen), so that only leads to the jail and to the R can were necessary. The jail current was used as an indication of the size of the U 238 beam both before reception, when the source voltage was lowered with a coarse control to put the U 238 beam on the 0-deg defining vane, and during reception. The quality of the beam and source-voltage settings was determined by means of the ferret circuit (see Fig. 5.30). By periodic modulation of the source voltage, the beam was swept back and forth almost 2 slugs on either side of normal setting for reception. The circuit (Fig. 5.30) was connected so that it would indicate source-voltage excursions on the horizontal sweep of an oscilloscope. On the vertical sweep of the same instrument another circuit indicated the corresponding changes in R ion current. The result was to cause the oscilloscope to trace a curve in the neighborhood of the plateau in Fig. 5.31a.

The modulation was applied to the source voltage as shown in Fig. 5.31b. Excursions of about 5 cps gave a steadily visible trace but rapid enough (0.0017-sec time-constant decay on each half-cycle) to cause an inappreciable amount of contamination, even if operated steadily. The appearance of the oscilloscope trace for good operation is shown in Fig. 5.31c. By observing the trace and switching from one R can to the other, it was possible to (1) determine the beam quality by the width and flatness of the R plateau as well as by the hash indicated, (2) set the source voltage by adjusting it until the spot was just on the plateau, and (3) set intercollector adjustment by moving No. 1 jail until the source voltage could be set correctly for both beams simultaneously.

5. ELECTRIC-SHIM CIRCUITS

In the electric-shim method of beam focusing, described in National Nuclear Energy Series, Division I, Volume 4, a source of constant electric potential must be available. The voltage regulators used in this connection were simplified variations of the high-voltage
regulators used with source units. Considerable simplification was possible owing to the fact that the electric shims required only about \( \frac{1}{2} \) amp current at about 2000 volts, as compared with the much larger output requirements of the usual high-voltage regulator. A typical voltage-regulator unit for electric shims is shown in Fig. 5.32.

It is seen from this figure that the power supply using rectifier tube V10 provides the voltage standard. A Beta divider network comprising resistors R12 and R13 and capacitors C11 and C12 is also included. A d-c power supply consisting of a suitable rectifier was connected between binding posts BP1 and BP2 with binding post BP2 at ground potential. The actual load, in this case the electric shims, was connected between binding post BP2 and the chassis terminal marked "Ch. Ter." An external control rheostat for the purpose of selecting the desired output voltage of the system was connected between binding posts BP4 and BP5 and the chassis terminal, with the arm of the potentiometer going to binding post BP5. An external voltmeter consisting of a 200-\( \mu \)a meter movement was connected between binding post BP3 and chassis terminal to indicate the output voltage of the system.

The operation of this system is essentially similar to that of certain of the high-voltage regulators described in Chap. 2. The output voltage of the system is also applied across the Beta divider network, and a certain portion of the output voltage is developed across resistor R12. This voltage is bucked against the standard voltage developed by the standard voltage supply using rectifier tube V10, a filter network including inductance L3 and capacitors C9 and C10, and two stages of voltage regulation using voltage-regulator-type tubes V7, V8, and V9 and including resistors R14 and R15. The difference between the voltage developed across resistor R12 and the selected portion of the voltage standard, as determined by the setting of the arm of the control potentiometer (not shown), is applied to the control grid of vacuum tube V1. The signal is further amplified by means of vacuum tube V2. In order to improve stability, a negative feedback network including resistors R4 and R7 and capacitor C4 is used. The output of vacuum tube V2 is coupled to the grid circuit of main regulator tube V5 through two voltage-regulator tubes V3 and V4 arranged in a circuit to reduce the d-c grid voltage on vacuum tube V5 to a suitable operating value. Vacuum tube V5 acts as a series losses tube to reduce the voltage of the rectifier-type power supply (not shown) to the value desired for application to the load.

This circuit is mainly distinguished from the high-voltage regulators by its noteworthy simplicity. It has not been deemed advisable to include a drawing of the rectifier-type power supply in view of the
fact that it is of quite standard design for the power requirements of
the system.

Not too much attention was paid to beam-regulating methods for
electric shims in view of the fact that this method of focusing was not
developed to the point where it could be used in the plant. However,
the work done on regulating methods follows fairly closely the general
methods outlined in Sec. 1, which deals with beam regulators. One
method was to insert a monitoring circuit between a collector elec-
trode in the R pocket and the electric shims’ power supply. An in-
crease in collector current increased the negative potential applied
to the shims, thereby increasing the radius of curvature of the beam.
In a second method, a slight ripple is imposed on the shim voltage,
which gives a corresponding ripple in the collector current. If the
shim voltage and collector-current ripples are in phase, the negative
potential of the shims is decreased, whereas when the ripples are out
of phase, the potential of the shims is increased. This is the general
method used in connection with three-sluggers.

6. BEAM INTEGRATORS

Although devices for measuring the magnitude of an ion current as
a function of time had been developed for use on cyclotrons prior to
the inauguration of the electromagnetic process, they were designed
for very small currents. In a cyclotron the currents are of the order
of microamperes, whereas in the electromagnetic process the cur-
rents obtained are measured in milliamperes. Consequently it was
necessary to develop new types of beam integrators or to modify the
existing designs. The principal types of integrators which were con-
sidered were (1) d-c motors, (2) capacitor discharge, and (3) elec-
trolytic deposition (coulometers or voltmeters).

With regard to the first method, it was found that two types of d-c
motors were available which could be used as current integrators.
One of these was manufactured by the General Electric Company, the
most sensitive model being rated at 10 amp and designed to read cur-
rents above 1 amp accurately. The second type of ampere-hour meter
was manufactured by the Sangamo Electric Co. and was rated at 5
amp. It was decided to be impractical to use these meters to integrate
the low currents encountered in calutron operation.

With regard to the capacitor-discharge types of integrators,* a
simple unit which was used consisted of a low-leakage capacitor
shunted by a neon glow lamp with the whole placed in series with the
beam. Current flowed into the capacitor and raised its voltage pro-
portionately. Then at the firing voltage point of the neon bulb, the
capacitor was suddenly discharged and the glow lamp extinguished
just as in an ordinary relaxation oscillator. The cycle then started again. A mechanical register for counting the number of discharges completed the device, the number of counts being the integration of the beam current over a given period of time. Such a simple device is limited in actual practice by collector leakage paths, either ionic or poor insulation, so that it is necessary to keep the collector voltage from varying away from zero without, at the same time, affecting the flow of beam current.

The first attempt to compensate for collector voltage variations in the milliampere region is shown in the circuit of Fig. 5.33. The compensation is effective to the extent of the effective gain of the 6F5 high-gain triode. Thus, if the voltage variation in the capacitor-discharger circuit (consisting of the 28-mfd capacitor and VR-75 glow tube) is from 90 to 75 volts, or 15 volts, and the effective gain of the 6F5 is 50, then, over the range of firing for a particular beam current the collector voltage varies only 15/50, or 0.3 volt. A mechanical register is connected to the terminals marked "output" to record the number of capacitor discharges. It will be noticed that there is a highly unorthodox arrangement of the 6F5 triode amplifier, i.e., the minus side of the circuit power supply is returned to the beam collector lead itself. This means that the beam current must flow through the power supply and capacitor circuit before it can return to ground through the tube. In fact, the tube plate current is exactly equal to the beam current. If no beam current exists, there is likewise an absolute cutoff in the tube current, and positively no count could be recorded by the register no matter how long that condition existed. If 1 ma of beam current flows, exactly enough bias automatically appears on the 6F5 to allow 1 ma of plate current to flow at the effective plate voltage.

The range of the circuit would be from absolute zero beam current to that current which would bring the grid of the 6F5 to the zero bias level. As long as the 6F5 grid works in a minus region the tube constants other than the amplification factor fall out of the picture entirely. Any source of nonlinearity is then confined to the glow-lamp–capacitor combination not always firing at the same voltage or to poor collector insulation. This latter problem is minimized as the circuit input approaches zero. The purpose of the 3-volt bias battery is to make the collector run at near zero voltage and at the same time to bias the tube to cutoff in case the collector shorts to ground completely. This bias could also be put in the cathode lead of the tube, and, by making it variable, it becomes an easy means of making the collector voltage run at any level.

It was apparent that a single-tube amplifier could never be expected to give much more voltage compensation than the original triode cir-
cuit. Consequently a multistage amplifier was applied to the same basic idea, and the resulting circuit is shown in Fig. 5.34. This integrator worked from zero beam current to 20 ma and kept collector voltage variations to within 0.02 volt of zero over the entire range. The previous explanations for the original circuit (Fig. 5.33) should serve to make the latter circuit self-explanatory. It should be added that one section of the 6SN7 tube is used simply as a capacitor-coupled pulse amplifier to drive the mechanical register. It is normally biased to cutoff and receives a steep positive pulse every time the strobotron (used as the new capacitor-discharging element) flashes, and the resulting pulse of plate current drives the mechanical register one step more. A rapid discharging of the strobotron is a very important factor since the faster the discharge occurs and is completed, the less will be the time during which the integration cycle is disturbed. It will be noted that a close parallel exists between beam integrators and beam sniffers. The last circuit (Fig. 5.34) would make a suitable sniffer merely by adding a variable resistor in series with the 6V6 plate and using the drop across this resistor as the signal for modulating the high-voltage supply in order to degenerate out any beam-current changes.

Electrolytic deposition is inherently the most accurate of the types of integrators, being used as a basis for the International Standard of electrical current. A number of different types of coulometers have been used to measure quantities of electricity, including zinc, sodium, silver titration, iodine, water, silver, copper, and mercury. There are disadvantages associated with all these coulometers when used for beam-integrating purposes. Zinc coulometers require weighing in order to make a reading. Although sodium coulometers can be made exceedingly accurate by using hot sodium glass as the electrolyte, readings must be made by weighing, and the resistance is fairly high. Iodine and silver titration coulometers have the disadvantages of requiring titration, limited capacitance, and high resistance. Water coulometers can be made direct-reading and have a number of convenient features, but large errors are introduced by very small amounts of double-valence salts. Silver coulometers are the most accurate of all the types. Unfortunately this type has a disadvantage in the nature of silver deposits on the cathode and the subsequent care necessary in weighing. Furthermore, it is not easily modified for direct reading. When tried a number of times in the laboratory, it proved to be impractical for these reasons. Although copper coulometers are quite accurate, sturdy, and easily constructed and require no extraordinary care in handling and weighing, the cathode plate must be quite large for the current ranges desired. The mercury coulom-
eter is not so accurate as the copper or silver forms, although the error is less than 0.5 per cent. However, it can easily be converted into a direct-reading instrument, and trained personnel is not necessary for the handling and operation of the device. Furthermore, the voltage drop for the range of current used is less than 0.3 volt. In view of these advantages some development work on mercury coulometers was carried out at the Radiation Laboratory.¹⁰

Mercuric iodide (HgI₂) was used as the principal ingredient of the electrolyte, which dissociates into Hg⁺⁺ and I⁻ ions. Under the influence of an electric current the Hg⁺⁺ ions migrate to the cathode and are neutralized to form metallic mercury. The iodine ions (I⁻) migrate to a mercury-pool anode, and the resulting I₂ combines with the mercury of the anode to form anew the salt mercuric iodide. The whole process can be operated in a closed cycle by transferring the mercury formed at the cathode to the anode pool.

Difficulties were first encountered in polarization of the anode. If the concentration of the mercuric iodide at the anode becomes too high, the salt precipitates, causing polarization and resulting in an increase in the voltage drop of the coulometer. To prevent this, two methods were devised, both making use of gravity as the agency for mixing the solution. The scheme used on the XA coulometer depended on the curvature of the mercury surface to drain the high concentration of mercuric salt from the anode, as shown in Fig. 5.35a. The second scheme is shown in Fig. 5.35b. In this case the mercury was supported by a nylon diaphragm because of the high surface tension of the metal. Owing to the characteristics of the nylon, the area of mercury contacting the solution was a large percentage of the total mercury surface. The voltage drop of this type of coulometer was not excessive.

The factors involved in an ideal cathode material are nonamalgamation, stability, smooth surface, purity, and availability. The literature reports the use of mercury, platinum, iridium, magnetic iron oxide, carbon, and palladium. All these materials appeared to have undesirable features; consequently tests were conducted on two additional metals, molybdenum and tungsten. These were found to be very satisfactory after proper polishing.

The most convenient way to measure collected mercury at the cathode is in a calibrated capillary tube. This method is quite accurate so long as the fine droplets falling from the cathode coalesce into an unbroken column of mercury. The finer the capillary, the more difficult this becomes. The first method used to overcome this difficulty was by applying an a-c voltage intermittently between the cathode and the column of mercury (for capillaries of 2 to 3 mm), which broke
down the surface tension and coalesced the mercury droplets. This scheme was not found to be practical on tubes smaller than 2 mm in diameter. Introducing a syringe into the capillary tube to coalesce the droplets mechanically was tried and found to be so successful with even the smallest available capillary tubes (0.25 mm) that the electrical scheme was abandoned. In the course of the experiments it was observed that coulometers, after being run for a few days, cleaned up and the bubbles coalesced of their own volition. This led to the obtaining of purer chemicals, which decreased the debubbling trouble. Chemicals were cp KI and HgI₂ and triple-distilled mercury. The sensitivity and capacitance in ampere-hours determines the bore and length of the capillary tubing. A rate of deposit of mercury of 3.7418 grains/amp-hr was adopted as a standard.

On the basis of the tests which were carried out, a mercury coulometer was designed for use on the XA unit. This coulometer, which is shown in Fig. 5.36, gave quite satisfactory results. Referring to the figure, it will be observed that a 3-way stopcock and syringe have been incorporated to pump the collected mercury back to the anode after measurements have been made in order to have a closed system. The syringe also serves the purpose of coalescing stubborn droplets. Since a low voltage drop is important, the anode and cathode areas were made as large as was convenient, the limitations being the stability of mercury and the conveniently available materials. A graph of the voltage drop as a function of current is shown in Fig. 5.37. The tungsten rod used as a cathode was ground and polished carefully in order to prevent excessive sticking of the mercury. The anode terminal (see Fig. 5.36) was tungsten tipped, with nickel (amalgamated with mercury) at the point of contact with the mercury anode. Pyrex was used in the construction of the glass parts because of its low temperature coefficient, ease of annealing, and workability. It was necessary to calibrate the scale because of the nonuniformity of the capillary bore. The results obtained showed up very favorably compared with the old method of taking current readings at 10-min intervals and estimating the outage time.

7. VAPOR-VALVE CIRCUITS

In the operation of a calutron it is important to provide suitable control of the vapor entering the arc chamber from the charge reservoir. If insufficient vapor flows, beam hash results, whereas too much vapor causes excessive drains. It was observed that the G-slit drain provided a good index of the vapor flow. Consequently automatic control systems for vapor were devised using this drain.
A very simple system which was used successfully for some time involved no mechanical valve but controlled the vapor flow by controlling the heating of the charge reservoir. The G drain was passed through a current-responsive device (in the high-voltage supply line) which could be a simple relay. Operation of this device, owing to the G drain becoming too low or too high, would actuate a relay in the temperature-control unit so as to close or open the heating circuit.

A somewhat more complex control system involving a mechanical vapor valve is shown in Fig. 5.38. Again the G-slit drain is used as the controlling agency. From the figure it is seen that the motor M, which controls the valve, is of the reversible type. The direction of rotation depends on the relative phase between the field and armature voltages. Windings 1 and 2 of the saturable-core reactors T1 and T2 and windings 3 and 4 of transformer 5 form a Wheatstone bridge. The input is an a-c voltage to points a and b, which are terminals of the center-tapped secondary of transformer 5, the primary being connected to a source of a-c voltage. Terminals c and d are the output terminals of the bridge and are connected through transformer 6 to the armature of motor M. The field of this motor is connected to the same source of alternating current that feeds the bridge. Therefore the relative phases of the motor field and armature voltages and consequently the direction of rotation of M are determined by the relative impedances of arms 1 and 2 of the bridge. Also, the greater the differences in the impedances of windings 1 and 2, the higher the value of the bridge output and the greater the speed of M.

Impedances of the arms 1 and 2 of the bridge are controlled by current through the primaries of T1 and T2, which are responsive to the current input to terminal J1. V1a has its output coupled directly through the gas diodes LP1 to LP3 to the input of V2 and by way of R1 to R8, etc., to the input of the phase-inverting triode V1. The output of V1 is directly coupled through LP4 to LP6 to the input of tube V3. The input to V1a is the voltage drop across R4 and part of potentiometer R2, which adjusts the control-grid bias of V1a. Thus the input signals to V2 and V3 respond in opposite directions to the input current to J1 which is the G drain current. That is, a rise in drain current gives a drop in input voltage to V2 and a rise in input voltage to V3.

Suppose that the bridge is initially balanced so that there is no output to the field of M and therefore no rotation. If a rise in drain current occurs, the voltage input to V3 increases, thus increasing the current through the T2 primary and decreasing the impedance of the arm 2. The voltage input to V2 decreases, thereby decreasing the current through the T1 primary and increasing the impedance of
bridge arm 1. The resulting unbalance of the bridge is in such a
direction that M causes the vapor valve to reduce the vapor flow,
decreasing the drain current. The drain current may be set by adjust-
ing the potentiometers R2 and R13, which vary the grid bias on V1a
and V1. By presetting the grid biases on V2 and V3 so that they pass
very little current at the equilibrium position, a small change in
drain current from the equilibrium value gives negligible unbalance
of the bridge (no rotation of M). This prevents hunting of the motor.

In view of the fact that the vapor valves were necessarily located
in the calutron magnetic field, it was necessary to locate outside the
calutron field those driving motors provided with their own field of
coils. However, operation was also carried out with motors using the
calutron field itself. The control circuits used were quite straight-
forward and will not be reproduced here. Another type of valve, called
"electromagnetic" or "magnetic," was used. In this type, the
throttling member had an electromagnetic force produced on it by
the interaction of a direct current through it with the calutron mag-
netic field. This force tended to open or close the valve, depending
on the direction of flow of the current, and was counterbalanced by
the natural spring of the valve itself. The electrical system for this
type of vapor valve is very simple, a schematic diagram being shown
in Fig. 5.39. With this valve, optimizing of the Q current to G drain
ratio, which is a rough measure of process efficiency, was rapid and
accurate. Other types of valves which were tried included thermal
valves, control in this case being provided by heating (or cooling) a
heater element which exerted a mechanical force on the valve proper,
and various designs of valves operated by mechanical linkages.

PART II. AUXILIARY MEASURING EQUIPMENT

8. LOW-IMPEDANCE ELECTRONIC CURRENT METER \(^{11}\)

In many cases it is desirable to measure and observe by use of an
oscillograph currents of the order of a few microamperes without in-
troducing excessive voltage drops into the circuit being measured.
In this section an amplifier which was developed for this purpose will
be described.

8.1 Basic Principle. An amplifier with a gain A connected as
shown in Fig. 5.40 makes it evident that the current i can flow in one
path only, that is, through the series combination of R, power supply
E, and tube V4. Also the input voltage \(e_s\) is equal to the difference
between the drop through resistance R and the sum of E plus the drop
across output tube V4. But this voltage $e_s$ is also applied as the input to the amplifier of gain $A$. Thus, if a change $\Delta e_1$ occurs across resistance $R$ by virtue of a change in current $\Delta i$, the voltage $e_s$ tends toward a change which appears at the output terminals as a voltage $\Delta e_0 = A\Delta e_s$. This latter voltage, being applied to the grid of tube V4, causes the drop across tube V4 to change in the opposite direction from the original change $\Delta e_1$, thereby reducing voltage $e_s$ to very nearly what it was before. Thus the input voltage $e_s$ remains virtually zero for all values of current $i$ within the limits of the amplifier, being approximately equal to the voltage across resistance $R$ divided by the total gain of the amplifier including tube V4. Thus, if the drop across resistance $R$ is 10 volts and the total amplifier gain is 5000, then the input voltage $e_s$ would be about $(10 \times 1000)/5000 = 2$ mv.

It is to be noticed, however, that the voltage $e_1$ appearing across $R$ is very closely equal to $iR$ and that, since $e_s$ is very small, the voltage from the top of $R$ to ground is also very closely $iR$. It is this voltage that is actually measured when the amplifier is used. It is evident, therefore, that the only function of the amplifier is to produce sufficient gain to keep the voltage $e_s$ constant and small and that if the amplifier is within its operating range and is operating correctly, the measured voltage $e_1$ is negligibly dependent on the characteristics of the amplifier over a frequency range extending from zero to the middle audio range to a point where the gain seriously reduces.

8.2 Amplifier Details. The detailed circuit is shown in Fig. 5.42; an elementary circuit for the input stage is shown in Fig. 5.41, where V1 and V2 represent the two identical sections of the 6J6 tube. It is clear from Fig. 5.41 that if the signal voltage $e_s$ increases positively, the plate current $i_{pl}$ of tube V1 will tend to increase causing an increased drop in resistance $R_k$. This drop also appears as an increased negative bias on tube V2 which then causes plate current $i_{p2}$ of tube V2 to decrease by almost exactly the same amount as $i_{pl}$ increased, and hence the net current through $R_k$ is practically zero. The net effect for tube V1, therefore, is as though the resistance $R_k$ was not there, and the gain of the stage is about the same as that of a single tube with no cathode resistor.

An analysis of this circuit for the class A operating condition (neglecting effects of characteristic curvature) shows that the output voltages $e_{o1}$ and $e_{o2}$ are given by the expressions

$$e_{o1} = e_s \frac{P R_k}{1 - P R_k} \quad e_{o2} = -\frac{P R_k e_s}{1 - P R_k} (PR_k)$$

where $P = \frac{K}{1 + KR_K} = \frac{g_m}{\frac{R_L}{R_p}}$.
Examination of these expressions shows that the output voltages will be equal and of opposite phase if \( PR_K = \frac{KR_K}{1 + KR_K} = 1 \). This is obviously true if \( KR_K \gg 1 \) or if \( \frac{g_m}{1 + R_L/R_p} > 1 \). For the latter condition to obtain, a tube with a high transconductance is needed and a large value of \( R_K \) with a low ratio of \( R_L/R_p \) is required. Under these conditions the gain of the amplifier stage is \( KL_R \) or \( \frac{g_m R_L}{1 + R_L/R_p} \) for each section. This gain is exactly that of a conventional tube amplifier with zero cathode impedance. These gain expressions are, of course, for the mid-frequency range.

The chief reason for using this type of input circuit in preference to other systems is because of the elimination of cathode drift. The construction of the 6J6 tube is such that the cathode drift effect is present in each triode section in very nearly equal magnitude and is therefore neutralized by virtue of the common cathode resistor. Slight unbalancing is neutralized by the potentiometer R22. Some reduction in input capacitance also results from the use of this circuit. It will be noted in Fig. 5.42 that the bottom end of R24 in the cathode of the 6J6 is returned to \(-105 \) volts from ground. This voltage bucks out the d-c drop across R24 in order to allow the tubes to operate with only a few volts of negative bias.

The second stage of the amplifier is made up of two 6SJ7 tubes with the grids directly coupled to the plates of the 6J6. The operation is very similar to that of the first stage, but the gain is considerably higher. With the conclusion of the 6SJ7 stage, the double-ended operation is discontinued and the output of one 6SJ7 is not utilized. The output of the other 6SJ7 is coupled to V4 (a 6AC7) by a conventional neon-tube coupler. The neon tubes operate to give a constant-potential difference of about 120 to 150 volts between the grid of the 6AC7 and the plate of the 6SJ7, an additional \(-105 \) volts being inserted to determine a satisfactory operating bias on the 6AC7. Capacitor C4 and R30 form a filter to eliminate high-frequency components of hash caused by the nature of the glow discharge of the neons.

The plate of the 6AC7 is fed through a second power supply of about 200 volts to the resistance network R1 to R20 and thence back to the grid of the input 6J6. The resistance R1 to R20 provides the actual calibration of the unit since it is the voltage developed across these resistors that is the actual output of the system.

8.3 Special Features. The purpose of the 6H6 diode, V6, is to limit the maximum current that can be fed through the circuit by the input. The voltage drop across the resistors R1 to R20 is negative
with respect to the input side of these resistors, and the diode is con-
nected across these resistors but is biased up to about 120 volts
negative with respect to its plate, so that the drop across R1 to R20
cannot exceed about 120 volts plus 1 volt or so drop across the 6H6.
The incorporation of the diode allows the observation of very small
R peaks during the beam-scanning process without difficulty arising
from the much larger Q currents which also flow into the input circuit
during the cycle.

The resistor R35 is connected from the power-supply end of R1 to
R20 to the +255-volt side of the main power supply. This causes a
small amount of negative current to flow through the feedback circuit
in order to eliminate the necessity of operating the output tube V4
completely to cutoff. The 6SH7 tube, V5, serves as a phase inverter
with unity gain to supply the plates of an oscilloscope.

Since the series combination of R1 to R20 plus the effective input
resistance of the amplifier forms a divider across which is applied
what may be a high-frequency signal, it is necessary to compensate
the divider for frequency. This is accomplished in the usual manner
by adding a capacitance split in parallel with the resistance divider
of value inversely equal to the resistance split, so that the time con-
stants of both sections are the same and are sufficiently long so as to
be effective.

8.4 Conclusions. This amplifier has a frequency characteristic
virtually flat from zero (direct current) to the middle audio range,
and when properly zero-adjusted for the middle range, it will give a
virtually constant input voltage drop of not more than about 10 mv.
The output voltage of the instrument behaves almost exactly as though
a resistance of from 0 to 55 megohms, variable in 0.5-megohm steps,
were being inserted directly in series with the circuit being measured.

9. DIRECT-CURRENT RATIOMETER

It was considered desirable to have available a meter that would ac-
curately measure the ratio of two direct currents, or more specifi-
cally, the ratio Q to R. This ratio is a measure of the purity of U 235
being obtained.

Several systems were tried involving logarithmic amplifiers and
variable-gain amplifiers, but without success. The principles of an-
other method which showed promise will be described in some detail
here. The essence of this method is a bridge scheme of balancing,
involving two currents flowing through two resistance branches. Any
voltage unbalance in the branches is fed to an amplifier with a bal-
ancing motor, which operates a potentiometer in one of the resistance
branches, as shown in Fig. 5.43. In this figure, the currents I₁ and I₂
flow through $R_1$ and $R_2$, respectively, producing definite voltage drops in both branches. If these drops are unequal, the amplifier sees the difference and actuates the balancing motor which moves the contact on $R_1$. This contact is wired so as to short-circuit out a portion of the resistance, and the voltage drop will be changed in a direction such that the drop in the $R_1$ branch will be equalized to that in the $R_2$ branch. The potentiometer $R_1$ is linked to a dial which indicates ratios linearly. The condition for balance is $I_1 R_1 = I_2 R_2$. Thus

$$R_1 = R_2 \frac{I_2}{I_1}$$

The ratio which is desired is $I_2/I_1$, and evidently the magnitudes of the currents do not enter the picture. Also $R_1/R_2 = I_2/I_1$, so that if $I_2/I_1$ is considered as a constant for each ratio, $R_1$ is linearly proportional to $R_2$. Thus the ratio points on the dial can be made linear if the potentiometer is linear for resistance vs. rotation.

The ratiometer was to have the following specifications: (1) a range of 10 to 55 (ratio), (2) accuracy of 1 ratio unit division, (3) a voltage drop of not more than 0.15 volt for Q currents up to 150 ma, and (4) a panel of standard rack size (5½ in. high). To meet the specifications of range and voltage drop, proper values of resistance must be chosen. Let $R_2$ be the large current (Q) branch (Fig. 5.44), this resistance being fixed, and $R_1$ the small current (R) branch, which thus has a higher resistance. It is preferable to make the large resistance $R_1$ the variable in order to minimize contact resistance in the potentiometer. Even though the specifications call for a range of 10 to 55, it is desirable to have a slight leeway on both ends so that the potentiometer contact does not have to go to the extreme ends of its travel. For this reason 9.5 and 59.5 were taken as the limits of the range. In order to satisfy specification 3, the value of $R_2$ in the heavier current branch must be 0.15/0.15, or 1 ohm. The equation $R_1/R_2 = I_2/I_1$ determines the value of $R_1$, which for a ratio 9.5 is 9.5 ohms. For the maximum ratio of 59.5, the value of $R_1$ is 59.5 ohms.

The resistance $R_1$ was composed of two resistances in series, one variable ($R_v$) and the other fixed ($R_f$). The $R_f$ had a value of 9.5 ohms for the low-ratio end of the range, while $R_v$ was a 50-ohm potentiometer. Although a General Radio Type 214 potentiometer proved satisfactory for $R_v$ in tests, it might be desirable to use a larger-diameter potentiometer with many turns of fine resistance wire, such as was used on Micromax recorders, in order to obtain smoother and more precise variations in resistance. Both $R_f$ and $R_v$ were made accurately with manganin wire. Since $R_2$ was very low, it was necessary to use
heavy and short connections to the ground point and amplifier input, as indicated by the heavy lines in Fig. 5.44. In order to cancel out all possible thermoelectric effects, there must be an even number of resistance wires to copper-wire junctions.

The amplifier unit (model No. 76020-1) with balancing motor (model No. 76750-2) was made by Brown Instrument Company and gave satisfactory operation in tests. A worm-gear system, with the worm gear on the potentiometer and the worm on the motor, was used with satisfactory results but with room for improvement. It appeared that a 10 to 1 gear ratio was best from the standpoints of quick scale coverage, quick balancing, and fineness of control. With this gear ratio, it was possible to cover the whole scale in about 20 sec. It was found that in order to reduce motor hunting to a minimum and to increase over-all accuracy, the gear setup must be quite rigid, have little backlash, and yet move with little friction over the scale. Stops were provided to limit the movement of the potentiometer contact in case the ratio went below 10 or above 55.

A resistor method of calibrating the device was used. In shop tests the instrument was found to operate quite satisfactorily.

10. ARC AND BEAM OBSERVATION EQUIPMENT

Under certain conditions of calutron operation, a random variation, or "hash," is found superimposed on the direct current of the ion sources or arcs. The arc hash results in a lowering in enrichment of the collected material, and consequently various efforts were made to find its cause and to develop methods for eliminating it. A theoretical discussion of hash, as well as experimental observations, are contained in National Nuclear Energy Series, Division I, Volume 5 (work of Massey's group). This section is concerned partly with arc-hash observation equipment used by L. E. Reukema\textsuperscript{13} and his associates and partly with some general beam-arc hash observation equipment.

Arc hash is primarily a function of four variables, namely, gas pressure, nature of gas in the arc, arc current, and strength of the magnetic field. In general, hash is increased by reducing gas pressure, reducing arc current, or increasing the strength of the magnetic field. For the magnetic fields and arc currents normally used, appreciable hash appears only at low pressures. Although the hash is eliminated by raising the pressure, a limit is set on this by the requirements of the process. It was felt that the hash might be reduced, or even eliminated, by preventing change in the arc current. Two methods of doing this were suggested: (1) the use of a large inductance in series with the arc and (2) negative feedback.
10.1 Equipment in Preliminary Experiments. To test the above principle, preliminary experiments were carried out, using the unfiltered output of a single-phase full-wave mercury-vapor-filled rectifier to supply both direct current and superimposed alternating current. The use of a mercury-vapor rectifier ensured the presence of high-frequency components because of the steep waveform of the current at the instant of electrical breakdown of the vapor every half-cycle. The ratio of alternating to direct current from this source was considerably higher than would ever be found in an arc. Direct currents ran as high as 3 amp, and a-c voltages were as high as 100 volts peak.

The type of circuit used in the work is shown in Fig. 5.45. Suppressor resistors in control grid, screen grid, and anode circuits were necessary to prevent r-f oscillations between tubes. In all tests an attempt was made to have the direct current passed by the 6L6's and 815's at least approximately neutralize the d-c ampere-turns of arc current in order to minimize magnetic saturation effects in the transformers. Such effects were found to be less troublesome than anticipated. The waveform and amplitude of alternating current across $R_1$ was studied by means of a calibrated cathode-ray oscilloscope. Frequencies as high as 15,000 cycles were present in the output of the unfiltered rectifier. Any a-c voltage drop across $R_1$ stepped up 2 to 1 by the input transformer, was amplified approximately 130 times by the first stage and about 4.3 times by the power stage, giving a voltage amplification of 1100, of which half was lost in the step-down ratio of the output transformer. The step-down ratio was necessary to match direct currents. The a-c voltage induced in the transformer secondary was approximately of such phase as to tend to buck any alternating current to the load. By means of this negative feedback it was possible to reduce the a-c component to 3 to 5 per cent of its value without feedback for direct currents as high as 3 amp, the smaller percentage holding for the smaller currents. The input transformer was included for use in later tests, where it would be necessary to insulate the feedback circuit from the source voltage.

Although high frequencies were present in the output of the unfiltered rectifier, they could not be considered as random noise. Therefore a buzzer was connected directly across $R_1$. Since considerably higher than rated d-c voltage plus the a-c voltage was thus impressed across the buzzer, it sparked violently, resulting in an abundance of random noise. The oscilloscope was connected directly across $R_1$ and therefore across the buzzer. The negative feedback reduced the random noise to about a fifth of its value without the feedback.
10.2 Equipment Used with the R1 Unit. The type of circuit finally used in the R1 studies is shown in Fig. 5.46. The three Gardiner 0.5-kva transformers were housed in a single case and were doubly shielded for 50,000 volts. One shield was connected to the source voltage and the other shield was connected to ground. This prevented any kickback to the adjusting controls in case of an arc to ground inside the tank.

Even when the potentiometer in the negative feedback circuit was set for zero amplification, making the feedback circuit inoperative, the presence of the three transformer windings in the arc-current circuit definitely decreased the alternating current shown by the oscilloscope. This was to be expected since the filter used to decrease a-c ripple in the output of the arc-current rectifier is tuned for antiresonance at 360 cycles, which unfortunately allows all the higher harmonics to get through with little opposition. The inductance of the transformers keeps these high audio frequencies out of the arc and also tends to reduce current variations due to impedance variations in the arc itself.

When the setting of the potentiometer was changed to produce ample negative feedback, the alternating current and random hash were further decreased by about 60 per cent, and the source-voltage hash was decreased by a somewhat smaller amount. Note that this was not nearly so good a result as found in the preliminary tests. Moreover, the decrease noted seemed to be nearly, if not completely, restricted to the frequencies below about 3000 cycles. If too much negative feedback was used, the hash again increased to a value far larger than the value when no feedback at all was used. Evidently for the higher frequencies the phase angle of the feedback departed very materially from 180 deg.

The frequencies present in the arch hash were studied with a Sky-rider all-wave radio receiver. By adjusting the vapor pressure of the arc to a critical value, the hash could be made to appear and disappear at fairly constant intervals, giving a motorboating effect, which could be speeded up or slowed down to almost any desired rate. In this way it was possible to check all frequencies from about 500 kc to 42 megacycles. Over this entire range a pronounced signal was picked up by the receiver for every period during which hash existed. Since hash signals came through for all frequencies throughout the frequency range of the receiver, apparently the hash originated in the form of exceedingly steep waveform pulses, which were able to shock-excite the receiver into oscillation, regardless of the frequency for which it was tuned. It was concluded that the hash consisted of high harmonics, up to possibly 30 kc, probably originating in the arc-
current rectifier, and a superimposed random variation of very steep waveform, probably caused by random accumulations of space charge in either the arc or the source and sudden discharges when these random accumulations reached critical values.

10.3 Further Experiments with Dummy Arc. In view of the fact that it was not feasible to make the R1 unit hashy, except for short periods of time, later tests were carried out on the dummy arc (Fig. 5.45) in cooperation with Massey and his associates (see National Nuclear Energy Series, Division I, Volume 5). Since the higher frequencies of the hash were the ones which the negative feedback system seemed unable to decrease appreciably, the various components of the test equipment were studied in order to isolate the difficulty. First the cutoff frequency of the transformers was checked. The transformers in this case were of the same type as those used with the R1 unit except that they were housed in three separate units rather than in a single unit. It was found that the transformers were series resonant at 65,000 cycles (no load in secondary) and the response was practically as good at 100 kc as at the low frequencies. It appeared, therefore, that the difficulty could not be attributed to the cutoff frequency of the transformers. Next the effect of magnetic saturation of the transformers was investigated. In this case it appeared that difficulties should be experienced at low frequencies rather than at high frequencies and any injurious effect could be eliminated completely by the use of more tubes in parallel or by using tubes which would pass more direct current.

Finally, the effect of phase angle introduced by the transformers was studied. To offer as low an impedance as possible to the d-c arc current, it appeared desirable to use transformers at both input and output of the negative feedback circuit. For frequencies ranging from a few hundred to a few thousand cycles, it is a simple matter to make the voltage fed back practically 180 deg out of phase with the voltage causing the undesired a-c variation. As the frequency is increased, however, the increased leakage reactance drops in the transformers, together with the increased effect of the charging currents in the internal capacitances of the transformer windings, and the capacitance across the arc terminals introduces a phase angle which increases rapidly with frequency and may exceed 180 deg for very high frequencies. As soon as this phase angle equals 60 deg, negative feedback loses all its beneficial effects. At a frequency of 3000 cycles (obtained from an oscillator) it was found that the feedback circuit could just about cut the impressed voltage in half.

10.4 Equipment for Observing Beam and Arc Hash. The equipment described above was used directly in attempts to eliminate arc
hash. Other equipment was built for observation of phenomena associated with beam and arc hash, simultaneously if necessary, in order to facilitate research on the subject. Such equipment will be described here briefly without attention to circuit details.

The equipment was designed in order that observation could be carried out by means of cathode-ray oscilloscopes, suitable means for photographic recording being made available. The general method used can be made clear by an examination of Fig. 5.47a. The rectangular-wave generator generates waves of the form shown in Fig. 5.47b. An integrating circuit, consisting of R and C, converts the rectangular pulses to triangular pulses to provide the horizontal sweep, while the rectangular pulse itself is used to provide voltage for a blanking circuit.

The triangular waves are produced in the following manner: If a square wave is applied to the input terminals (see Fig. 5.47c), the capacitor will charge on the positive half-cycles and discharge on the negative half-cycles, assuming the negative side of the circuit is grounded. If the time constant of the RC network is long compared to the period of the square wave, the capacitor will only charge to a small fraction of the total voltage of the square wave. Since the exponential charging curve of a capacitor is almost straight over its lower portion, the result is an almost linear rise in voltage across the capacitor. When the cycle reverses, the capacitor discharges in a similar manner. With a square-wave input, such an arrangement produces almost perfect triangular waves. The only portion of the triangular wave used is the rising part which is linear, the oscilloscope being blanked over the rest of the cycle. The resistance of the circuit in the charging cycle is the output impedance of a 6SN7 cathode follower (about 300 ohms), whereas in the discharging cycle it is about 10,000 ohms. Consequently the voltage across the capacitor tends to decay very slowly. To overcome this effect, a 6H6 tube is connected in order to short-circuit the R part of the circuit on the discharging cycle. Without this tube the triangular wave discharges as shown by the dotted line of Fig. 5.47b. The operation then proceeds in the following manner: When the rectangular pulse comes along (see Fig. 5.47b) it turns on the oscilloscope, at the same time generating a linear sweep voltage as shown. At the end of the pulse, the trace is blanked out and the sweep voltage falls to normal, as described above.

The rectangular-wave generator referred to is a multivibrator with several interesting features evolved to assure flat tops and square corners on the rather fast rectangular pulses used (up to 50 μsec). The most obvious innovation is the use of cathode followers (6SN7's) between plate and grid (see Fig. 5.48) on both sides of the circuit.
The cathode followers provide impedance transformation from the high-impedance pentode plate circuits to the grid-coupling circuits, providing in effect a low-impedance generator for charging the grid-coupling capacitors, thus shortening their charging period and assuring a steep rise time. A second feature of the circuit is the use of the clipper tube (6H6). This tube cuts off the positive half of the rectangular pulse when it reaches the potential of the clipper cathode, normally held at 105 volts by the voltage-regulator tube. This is done to prevent rounding off of the rising corner of the rectangular pulse as the plate current of the pentodes approaches its maximum value. A third point of interest is the returning of the pentode grid resistors to a positive voltage rather than to ground. This has a similar purpose to the foregoing. It prevents the bending off of the grid-coupling capacitor voltage as this capacitor discharges and approaches its final value, thus giving a squarer pulse. Also, the circuit is made more stable by making the range of voltages where the tube begins to conduct occur in a shorter period of time. Finally, by returning the grid of 6SH7 No. 2 to the 100,000-ohm potentiometer, a convenient means is provided for varying the sweep repetition rate.

Du Mont 208 oscilloscopes were used with the circuit to make the observations. These scopes were modified by using 4000 volts or more on the intensifier electrode rather than the 1500 volts provided. In this way sufficient brilliancy could be maintained for photographing with the 50-μsec sweep. A possible type of pulse generator which was proposed for use in simultaneously observing arc and beam phenomena is shown schematically in Fig. 5.47d. With PB open, the thyratron grid is biased beyond cutoff of the maximum power-supply voltage. C charges to full voltage through R until PB is depressed. The thyratron fires, discharging C suddenly through the two transformer primaries. The very fast incidence of the large currents builds up large pulse voltages across the transformer secondaries which are delivered to the sweep units to initiate the sweeps. Because of the time delay of the beam in the tank (of the order of 25 μsec), elaboration of the circuit was considered.

11. ECCENTRICITY TESTER

It is often of importance to be able to make accurate and rapid measurements of the amount of eccentricity of rods, where the term "rods" is used to include concentric lines, heaters, heater leads, etc. For this purpose an eccentricity meter was designed and built. The principle of operation of the instrument is the following: A 60-cps electric current is passed longitudinally through the inner lead of the
rod and returned through the outer conductor by means of a short-circuiting chuck at the far end of the rod. The meter responds to the external magnetic field that is produced by the 60-cps current in the concentric line when the line is not truly concentric. The external field produced by a current in the outer conductor is the same as though the current were concentrated at the geometric center of the conductor. If the inside conductor is off center by an amount \( X \), the external magnetic field produced by the total line current is the same as would be produced by a parallel-wire line having a separation between wire centers equal to \( X \). The shielding effect of the outside conductor is quite small for 60-cps fields.

If two pickup coils are placed as shown in Fig. 5.49 (coils \( A \) and \( A' \)), the sum of the induced voltages will be very nearly proportional to the product of \( X \) and the cosine of the angle \( \theta \) between the plane of the pickup coils \( AA' \) and the plane of the equivalent parallel-wire line. If a second set of coils, \( B \) and \( B' \), are placed at right angles to \( A \) and \( A' \), the voltage induced in this set will be proportional to the sine of \( \theta \). Since the sine squared plus the cosine squared is equal to unity, a meter that reads the sum of the square of the voltage from coils \( AA' \) and the square of the voltage from coils \( BB' \) will give a reading that will be independent of orientation of the plane of the equivalent parallel-wire line with respect to the coil. Using four coils in this manner makes it practical to design a clamp-on type of instrument, and the measurement obtained for a given eccentricity will be quite independent of the position of the pickup instrument.

A simplified functional diagram of the instrument is shown in Fig. 5.50. The instrument consists essentially of two sets of pickup coils, two amplifiers, and two square-law detectors, the combined outputs of the latter being indicated by the meter. Current to the concentric lines being tested is supplied by a current supply unit. In the tests made in the laboratory, this unit was provided with two voltage ranges, a 0- to 6-volt range for testing low-resistance heater leads, and a 0- to 150-volt range for testing heaters. The maximum current rating was 4 amp.

The coils \( A \) and \( A' \) are connected in series (see Fig. 5.49) so that their voltages are additive. Coils having 1100 turns each and a resistance of 190 ohms, which had negligible 60-cps reactance, were found to be suitable. Coils \( AA' \) are connected to channel 1 of the amplifier unit, which has an input resistance of 3000 ohms. An input voltage of approximately 580 \( \mu \)V gives full-scale reading of the meter.

The amplifier unit consists of two similar channels, one for amplifying the voltage from coils \( AA' \) and one for amplifying the voltage from coils \( BB' \). The schematic diagram of the amplifier unit is shown in
Fig. 5.51. Each channel consists of a two-stage feedback amplifier having a voltage gain of approximately 1500, a triode amplifier (tube V5) having a 60-cps voltage gain of approximately 9, and a square-law detector (tube V6).

The grid bias of the detectors is determined by resistors R40, R41, and R45, and their values are such that the plate current of each section of the detector tube is proportional to the square of the input voltage of the corresponding channel. Since the meter reads the sum of the plate currents of the two sections of the tube, it measures the sum of the squares of the voltages from coils AA' and coils BB'. The meter has a squared scale which indicates directly the amount of eccentricity. The zero of the meter is set by means of potentiometer R41 which controls the grid bias of the detector tubes. The grid bias of −5 volts exceeds the peak value of the maximum grid signal voltage by only a fraction of a volt. This combination and resistors R28 and R29 prevent excessive overloading of the meter due to large input signals.

Since the amplifiers are intended for amplifying only 60-cps voltages, the high-frequency response is reduced by means of capacitors C15 and C16. Gain control is provided by potentiometers R22 and R23, and a calibrating voltage is available through switches D1 and D2. These switches also provide for short-circuiting the input of either channel. The plate power supply is regulated and sufficiently filtered so that hum voltages are negligible. A 1-ohm resistor, R30, placed in the heater lead of the detector tube delays the emission from this tube after the instrument is turned on until the d-c potentials of the amplifier have reached their normal values. This protects the meter during the warming-up period. The meter is short-circuited by the on-off switch when the amplifier is turned off.

Owing to a combination of errors in the unit itself (mainly a slight inherent hum pickup and inability of the meter circuit to yield accurately an output proportional to the square of the input), a maximum error of ±3 per cent is probable. Another ±2 per cent error may be attributed to error in reading the meters. Thus the over-all maximum probable error may be ±5 per cent when the instrument is operated in a location free from stray magnetic fields.

Figure 5.52 shows the complete eccentricity unit with connecting cords and sample rods. Figure 5.53 shows more details of some of the component parts. The clamp-on pickup head is designated by C in Fig. 5.52. The head contains the two sets of pickup coils and the meter which indicates the amount of eccentricity. The flush lead power chuck which makes the connections between rod and current supply unit is designated by E. The outer conductor of the rod is clamped
in the spring collet jaws when the knurled conical tube is tightened. The knurled knob A (Fig. 5.53) is attached to an extension which makes contact with the inner conductor. The flush lead short-circuiting chuck, which makes the connection between inner and outer conductors, is designated by F. These chucks were designed to operate only for rods where the conductor extended less than $\frac{3}{8}$ in. beyond the outer conductor and with rods that were sawed off flush. Other chucks (extended lead power chuck and extended lead short-circuiting chuck) were available for the cases of rods having the inner conductor extending $\frac{3}{8}$ in. or more beyond the outer conductor.

12. PARTICLE COUNTERS AND DETECTORS

No extensive development work on electronic circuits for detecting radioactive particles was carried out in the Radiation Laboratory in connection with the electromagnetic separation process. However, it may be of interest to discuss briefly a few of the circuits which were designed. Two ionization-chamber circuits (designed primarily for fission counting), a photoelectric alpha-particle detector, and two Geiger-tube circuits will be treated in this section.

12.1 Completely Portable Linear-amplifier Counting Unit. Alpha and other heavily ionizing particles produce ionization in a typical ionization chamber, the pulses from which are amplified in a built-in four-stage linear amplifier. The amplifier first stage consists of a low-filament-drain acorn-type tube 959 (V1 in Fig. 5.54). This is followed by three stages of amplification featuring low-filament-drain pentode-type tubes 1N5-GT (V2, V3, and V4 in Fig. 5.54).

The amplified pulses, 50 to 80 volts, are then applied by means of the switch S1 to either a scale-of-1 counting circuit or to a cumulative charge-type scaling circuit. The scaling circuit embodies the tube V5 (1N5-GT), which charges capacitor C3 to the breakdown potential of the neon bulb NE. The scaled pulses, or the pulses direct from the amplifier, are applied to the discriminator tube V6 (1N5-GT), which is not sensitive to voltage disturbances from the amplifier, having a peak voltage less than approximately $22\frac{1}{2}$ volts positive. The pulses passing the discriminator then actuate a self-contained register by means of the discharge of the cold-cathode thyratron V7 (OA4-G). The discriminator was included in order that fission fragments could be counted without interference from beta particles, gamma rays, etc.

Voltage for collection of ions in the ionization chamber is obtained from the cold-cathode thyratron relaxation oscillator V9 (OA4-G) and associated circuit. The high voltage from transformer T1 is rectified by the low-filament-drain tube V8 (1G4-G) and applied to the
chamber through the filter network consisting of capacitors C1 and C2 and resistor R1. The collection voltage is maintained around 1000 volts. The unit was completely self-contained in a metal case measuring 10 by 12 by 6 in. The weight of the complete unit, including batteries, was approximately 37 lb. Circuits similar to the above were later designed for stationary use in fission counting, in which case lead cells rather than dry cells were used to avoid voltage changes with time.

12.2 Circuit for Low-input-capacitance Ionization Chamber. The ionization chamber is normally connected in the grid circuit of a high-gain amplifier tube. Since the voltage pulse produced by the chamber (if it is due to a fission) is inversely proportional to the capacitance of the chamber, it is desirable to make this capacitance as small as possible. The smallest attained at the time this work was done was about 20 $\mu$μf. However, the effective grid capacitance of the amplifier is normally around 200 $\mu$μf. Therefore the voltage pulse in the grid circuit due to a fission is less than the voltage pulse that would be produced across the ionization chamber alone in the absence of the effective grid capacitance. To increase the sensitivity and signal-to-noise ratio of a counting circuit connected to such an ionization chamber, a cathode-follower circuit was included between the ionization chamber and the amplifier.

From Fig. 5.55 it is seen that the plate of the cathode-loaded tube is effectively grounded by the large capacitor C1, so that the grid-to-ground capacitance C2, effectively connected in parallel with the input resistance R1, is equal to that of the grid-to-plate capacitance $C_{pg}$ of the tube. With this arrangement, the effective input capacitance can be made as low as 2 or 3 $\mu$μf (less than that of the ionization chamber). The pulses appearing at the output of the cathode-loaded tube are applied to a high-gain amplifier (fairly standard design) and then to a counter. Consideration was also given to the use of a grid between the ionization-chamber electrodes so as to reduce the capacitance of the chamber.

12.3 Photoelectric Alpha-particle Detector. (a) Introduction. Various methods of detecting and counting α-particle sources are well established. The oldest is the method of scintillations, the α particle in this case producing a short pulse of light when it impinges on a thin screen of zinc sulfide crystals. Recently this method has been replaced in most work by that of the shallow ionization chamber coupled to a linear high-gain amplifier, the output of which is fed to some form of scaling circuit and mechanical recorder (see Sec. 12.1). A variation of this method is occasionally adopted—the use of the proportional counter. Here the original ionization in the chamber is
multiplied some $10^3$ times by collision processes in the gas, and the required tube amplification consequently is reduced. Less-sensitive devices of an integrating nature are occasionally employed. In these the mean ionization current produced by the entry of $\alpha$ particles into a chamber is amplified by one or two tubes and recorded as a small current. This type of detector is essentially similar to the electrometer since the small ion current is used to charge a very small capacitance, e.g., $10 \mu\text{uf}$, which is almost completely insulated (leakage resistance of the order of $10^{12}$ ohms). The time constants necessary at high sensitivity make it difficult in practice and usually ion currents of $10^{-12}$ amp are recorded as the lower limit. Finally there is the method of detection by means of a cloud chamber.

Features which are generally considered important in any form of detecting apparatus may be listed as follows:

1. High sensitivity. The limit is reached when each particle is detected.
2. Resolving time. This determines the maximum permissible rate of counting and defines to a large extent the range of intensity which can be measured.
3. Background. It is important that the equipment should be capable of detecting $\alpha$ particles in the presence of disturbing radiations, such as $\beta$ and $\gamma$ rays.
4. Robustness. This term includes such things as freedom from microphony and spurious effects from magnetic or electric fields.
5. Serviceability. This usually implies simplicity in the circuit arrangements since this generally leads to greater reliability.

It is believed that the short description of the combination of the method of scintillations with the use of photoelectric current amplification which follows shows that an instrument can be produced which for most purposes satisfies the five requirements listed above.

(b) General. The sensitivity of scintillation apparatus is the maximum when each $\alpha$ particle is recorded. It is known that with the correct screen materials and observation conditions a single $\alpha$ particle produces a scintillation which is clearly observable as a flash. Charlton and Lea showed that these scintillations were recorded by the eye of an experienced observer with 100 per cent certainty when they corresponded to 30 or more quanta entering the eye. This result held for flashes of all durations less than 0.01 sec. They further established the fact that good zinc sulfide screens converted an average of 15 per cent of the $\alpha$-particle energy into luminous energy, a high process efficiency.

The relation $1.57 \times 10^{-6}$ erg $= 10^6$ ev is now considered. The energy of an $\alpha$ particle of $2 \times 10^6$ ev is therefore $3.1 \times 10^{-6}$ erg. The average
The luminous energy emitted by the absorption of this particle in the zinc sulfide is $0.15 \times 3.1 \times 10^{-6}$ erg, or $0.47 \times 10^{-6}$ erg. The energy of a quantum of wavelength 4500 A = $4.3 \times 10^{-12}$ erg, and this may be taken as representing approximately the energy of the quanta emitted during a scintillation produced in a blue phosphor material. Hence one $\alpha$ particle releases an average of $(0.47 \times 10^{-6})/(4.3 \times 10^{-12})$ quanta, that is, $1.1 \times 10^5$ quanta per scintillation.

Referring the above result to the manufacturer's data on the sensitivity of a photocell of the electron-multiplier type it can be shown that there is a high possibility of detection. For the RCA tube type 1P21 there are the following data (at 110 volts per stage): (1) luminous sensitivity is 11 amp/lumen, (2) current amplification is $5.5 \times 10^5$, and (3) sensitivity at 3750 A is 9700 $\mu$a/$\mu$w.

Considering the third factor in the above data, it is known that $10^{-6}$ watt produces 9700 $\mu$a, or since 1 watt = $10^7$ ergs/sec, we have 10 ergs/sec = $9.7 \times 10^{-3}$ coulomb/sec, i.e., 10 ergs releases $9.7 \times 10^{-3}$ coulomb. Since each quantum of wavelength $\lambda = 3750$ A has an energy of $5.3 \times 10^{-12}$ erg, the data show that, if $n$ is the number of quanta involved in the release of $9.7 \times 10^{-3}$ coulomb, then $n \times 5.3 \times 10^{-12} = 10$, or $n = 1.9 \times 10^{12}$. Therefore $1.9 \times 10^{12}$ quanta release $9.7 \times 10^{-3}$ coulomb and the sensitivity is $5.1 \times 10^{-15}$ coulomb/quantum.

The sensitivity of the tube type 1P21 varies with wavelength but it is still high at $\lambda = 4500$ A, the wavelength used in determining the number of quanta emitted per scintillation produced by an $\alpha$ particle of energy $2 \times 10^6$ ev, namely, $1.1 \times 10^5$. If all the quanta so emitted fell on the sensitive surface of the photocell, each scintillation would release $5.1 \times 10^{-15} \times 1.1 \times 10^5$ coulomb, that is, $5.6 \times 10^{-10}$ coulomb.

The rate of release of the energy is an important quantity and the visual duration of the flash has been determined experimentally and estimated at approximately $50 \times 10^{-8}$ sec. It is known that the initial intensity of the blue phosphor material builds up very rapidly and decays approximately according to the $1/T$ law, $T$ representing time. Assuming that buildup takes a fraction of 1 $\mu$sec, during the first 2 $\mu$sec a number of quanta approximately equal to half that emitted during the succeeding 50 $\mu$sec have been emitted. Thus in 2 $\mu$sec about one-third of the quanta are emitted and therefore one-third of the charge released, that is, $1/3 \times 5.6 \times 10^{-10}$ coulomb or $1.9 \times 10^{-10}$ coulomb. This would be equivalent to a pulse current of amplitude $(1.9 \times 10^{-10})/(2 \times 10^{-6})$ amp, or 95 $\mu$a.

The above calculation shows that it is possible to obtain current pulses of peak amplitude approximately 95 $\mu$a in the anode load of a 1P21 tube exposed to $\alpha$-particle scintillations if it is assumed that all the quanta emitted fall on the photosensitive surface.
(c) Experimental Results. A weak collimated beam of $\alpha$ rays was directed at a small screen of zinc sulfide placed about 1 mm from the surface of a 1P21 tube. The beam was of circular cross section of 6 mm diameter. The geometry (see Fig. 5.56a) was such that the sensitive photosurface subtended at the screen a solid angle of about $0.4 \pi$. From the calculations above, $\alpha$ particles of energy $2 \times 10^6$ ev should produce current pulses of amplitude $9.5 \times 10^{-6}$ amp.

The circuit used with the 1P21 is shown in Fig. 5.56b. The output from the $5 \times 10^5$ ohm load was examined on a cathode-ray tube. Sharp voltage pulses considerably greater than the noise level were observed. Calibration of the oscilloscope indicated that the average amplitude of the pulses produced by the scintillations was approximately 6 volts. The variation in the amplitude was rather great. With the arrangement shown in Fig. 5.56b, the expected voltage output from the 1P21 was $5 \times 10^5 \times 9.5 \times 10^{-6}$, or 4.75 volts.

In view of the somewhat approximate nature of the calculations, the measured value of 6 volts may be taken as in fair agreement with the calculated one of 4.75 volts. The larger measured output could be ascribed to various factors, two of the most probable being:

1. Pulse time. A pulse duration of less than the expected value of $50 \mu\text{sec}$ would increase the output.

2. Conversion efficiency. This might be greater than the expected value of 15 per cent.

The apparatus should be capable of resolving these questions in due course.

(d) Absolute Efficiency of Detector. Various types of screen material were examined and a blue phosphor of zinc sulfide, silver sensitized (RCA sample No. 3), was found to be the most suitable. With screens of this material and the geometry given above, the pulses corresponding to the scintillations were counted when they reached a minimum amplitude of between two and three times the average peak noise level. With this criterion for bias adjustment, the background was measured and was found to be approximately 0.5 pulse/min. The average amplitude of the pulses produced by the $\alpha$-particle scintillations was roughly 6 volts, or twice the bias voltage of 3 volts. A large percentage of the pulses were in excess of 12 volts amplitude. The apparatus, adjusted in this manner, was checked by counting different sources and comparing these counts with those produced by the same sources when used in a shallow ionization chamber coupled to a linear amplifier and scaling circuit. The results showed that the zinc sulfide screens could easily be made to record between 80 and 100 per cent of the number of $\alpha$ particles falling on the screen.
It was thus established that the 1P21 can be a detector of α particles of 99 to 100 per cent efficiency when screens of a suitable zinc sulfide powder are prepared by a proper technique. The screen must be such that no holes for α particles are exposed, and its depth has to be less than would render it opaque. Tests showed that these conditions can be met.

12.4 High-speed Geiger-Mueller Circuit and Integrator. For various reasons it was desirable to have a design of Geiger-Mueller counting circuit which would be free of some of the difficulties associated with others and which at the same time would afford at least equal or greater advantages.

The Neher-Harper circuit (and various versions and adaptations of it) gives the counter a fast response characteristic even if the counter tube itself is of the slow type. The drop of voltage across the tube is not, however, of square-wave form, and the return voltage sweep is slow compared with the drop time—it is in fact exponential in form. It seemed desirable to modify this so that the tube was dropped rapidly below a counting voltage and brought back to the normal counting point almost equally as rapidly. This means that each count occupies a definite interval of time, and errors at different counting rates can be easily calculated. The dead time of the tube could be varied over a number of different values. This will be shown to have two distinct advantages: (1) different counter types can be handled, good counters having a short dead time and poorer counters a longer dead time, and (2) at slow rates of counting the sensitivity of the integrator is easily increased by making the time of the pulse long.

Some circuits have both electrodes of the counting tube electrically live. This is a disadvantage in certain circumstances such as when the counter is being operated in the presence of static interference. The ideal arrangement is to have the counter wall in the form of a grounded metal screen since it thereby forms its own interference protection. The circuit to be described observes this principle.

In many cases counters have to be operated in positions remote from the main part of the associated circuits. The pulse in these circumstances has to be transmitted along a length of cable of low impedance. The circuit described accomplishes this by having a pre-amplifier with cathode-follower stage closely attached to the tube itself. The cathode-follower action following amplification means that quite weak tube signals can be transmitted without appreciable distortion over distances of 20 to 30 ft.

(b) Circuit Arrangement. This is shown in Fig. 5.57. The case of the Geiger-Mueller counter tube is grounded, and the remote positive high-voltage supply is applied via two resistors, each of value $5 \times 10^4$. 
ohms, one at the supply and one at the tube. The self-capacitance of
the cable is thus isolated by the resistors. The signal is taken from
the wire by means of a small capacitor (125 \( \mu \)F) to the grid of the
first amplifier and inverter. This grid is protected against complete
failure of the counter (or capacitor) by the \( \frac{1}{25} \)-watt miniature neon
lamp strapped between it and ground. The cathode follower V2 is
part of the same double triode (6SN7), and signals are transmitted
along the cable without distortion to the main circuit. The voltage
fed to the head amplifier stage is normally about 140 volts. Tube V3
acts as an amplifier and inverter, and its anode is directly coupled to
the anode of V7, which, together with V4, forms a multivibrator. The
negative drop in potential at the point of common connection of the
anodes is fed via switch \( S \), and one of the three capacitors, to the grid
of V4 which is normally conducting. This pulse serves as a trigger
to the biased multivibrator. The positive output of V4 drives the tube
V7 into the conducting state by overcoming the applied bias. This
bias level is not at all critical and may be set approximately in the
range 25 to 65 volts. The double diode 6H6 serves to remove back-
swing pulses which produce undesirable effects. The positive output
pulse from V4 is passed through the cathode follower V8 to the power
output tube V9 and the integrator V10.

The output tube V9 is normally nonconducting and the positive input
square wave applied to the grid produces a negative square-wave out-
put at the anode. This is applied to the high-voltage cable by means
of a coupling capacitor. The tube V9 is of adequately low impedance
to drop the voltage on the line in a very short time. Thus, for 500 \( \mu \)F
of cable self-capacitance there is a charging current of about 0.15 amp,
giving \( i \, dt = C \, dV \), or 0.15 \( dt = 5 \times 10^{-10} \times 10^2 \), or \( dt = 3.3 \times 10^{-7} \) sec
for a voltage drop of 100 volts. The counter itself generally has a
self-capacitance of, say, \( 5 \times 10^{-12} \) farad at most. The time constant
for the tube is then \( RC = 5 \times 10^4 \times 5 \times 10^{-12} = 2.5 \times 10^{-7} \) sec. Thus
a rapid drop in the voltage across the tube to a value below the ex-
tinction potential can be accomplished. The positive voltage on V9
can be chosen to suit particular requirements as to length of "plateau."
It was operated usually with an applied voltage of 600 volts.

(c) Sensitivity. The sensitivity of the circuit to negative pulses
produced across the tube resistor was measured and found to be about
0.1/volt for full output. Thus a current of 2 \( \mu \)a through the tube pro-
duces the reactive effect, and this could be dropped to even lower
values by increasing the series resistor to higher values. This would
have no deleterious effects provided that the self-capacitance of the
counter and the negative pulse produced by V9 were maintained at
values sufficiently small and large, respectively, to satisfy require-
ments as to the rate of drop of voltage across the tube.
A simple count integrator was included in the circuit to make it independent of external scale-of-2 arrangements. Tube V10 is a pentode of sharp cutoff characteristics (6SH7), and it is biased well beyond cutoff. Generally the residual current to the anode in these circumstances was of the order of $10^{-9}$ amp. The screen was given a high voltage (+300 volts) to increase the peak currents passed by the tube when pulsed. The grid was pulsed effectively to zero by driving it through a blocking resistor tied to the grid of V9. This grid thus was made to serve effectively as a diode since its voltage could not rise appreciably above ground potential. Under these circumstances, assuming a steady 300-volt supply to the screen of V10, the amplitude of the pulse current through the tube remains constant and of the order of 30 ma. Its value is independent of anode voltage over a considerable range. This was kept at about 150 volts since higher values produced higher leakage currents. An external microammeter could be plugged into the circuit to read the count rate. It was damped by means of a resistance-capacitance network as shown. The sensitivity of the instrument can be illustrated as follows: With the pulse length at 100 $\mu$sec (suitable for weak sources) we have $30 \times 10^{-3} \times 10^{-4}$ coulomb/pulse, i.e., $3 \times 10^{-6}$ coulomb/pulse. Thus, 20 pulses/min (average natural count for a medium-sized counter tube) would give a steady current of $(3 \times 10^{-6} \times 20)/60 = 10^{-6}$ amp = 1 $\mu$A. This means that the natural or background rate of counting is easily detected on a commercial microammeter with a full scale corresponding to 20 to 50 $\mu$A. For very fast rates of counting it is possible to (1) reduce the pulse length and (2) shunt the meter. A wide range of rates can thus be covered, the pulse length being varied from, say, 20 to 200 $\mu$sec.

12.5 Counting-rate Meter. The counting-rate meter described here was designed to give linearity of output to counting rate and to include a simple arrangement for calibrating the different ranges in a single operation. From the schematic diagram of Fig. 5.58 it is seen that positive pulses from the Geiger-Mueller tube are impressed on the grid of tube V1. When the corresponding negative pulses appearing in the output of V1 are applied to the trigger circuit, including tubes V2 and V3, V3 is changed from the nonconducting to the conducting state. Charge passing through V3 is impressed on the integrating network, including capacitor C7 and one of the resistors R11 to R17. The voltage produced across C7 is proportional to the counting rate. Since the integrating network is external to the grid circuit of V3, the voltage across the capacitor does not affect the strength or length of the current pulse through V3.

The voltage across C7 depends on the value of the precision scaling resistor connected across it by the selector switch D5 and is meas-
ured by a vacuum-tube voltmeter coupled to it by R6. Resistor R6, together with C8, protects the meter M1 from swinging violently. To calibrate the unit, D1 is switched from the operating position to the standard position, and the reading on M1 is brought to a standard value by adjustment of R20. This calibration is correct for all values of R11 to R17.

13. VOLTAGE-REGULATOR-TUBE SCANNER

The circuit to be described here was devised to check the performance characteristics of commonly used voltage-regulator tubes, primarily to save time. The general method consists in passing direct current of any desired value through the tube being tested. This current is then modulated with a low-frequency triangular wave at any value up to 100 per cent. The voltage drop across the voltage-regulator tube is impressed on the vertical plates of an oscilloscope whose horizontal sweep is synchronized to the modulation frequency. The sweep voltage is also of triangular shape, thus providing two traces on the screen, one for rising current and one for decreasing current. Means are provided to separate these traces so that they are not superimposed. In effect, a dynamic graph of voltage across the voltage-regulator tube as a function of the current through it is provided.

It is seen from Fig. 5.59 that tube V1 to V4 and the associated network comprise a voltage-regulated power supply. The double triode V5, with the circuit connected to it, is a multivibrator having neon-tube coupling, which is locked with the ripple output of V4. The output of the multivibrator is a steep-sided square wave which is fed to triode V6, the output of which is applied to a series RC network. The voltage across the potentiometer P4 is a triangular wave and is directly coupled to the control grid of V8, pentodes V8 and V9 comprising a 2-stage degenerative feedback d-c amplifier, the current output of which closely follows the voltage on the control grid of tube V8. The output of V9 is fed through the voltage-regulator tube to be analyzed, which may be plugged in socket S1 or connected to the corresponding binding posts BP1 and BP2. The voltage across the regulator tube is applied through the RC matching network (consisting of R16 and C9), jack J1, and a coaxial cable to the vertical plates of an oscilloscope. Also applied to the vertical plates is the synchronized square-wave voltage output of triode V7 through the RC network consisting of R17, C10, and C11, so that the oscilloscope will have different horizontal references for rising and falling currents through the test regulator tube.

The oscilloscope horizontal sweep voltage is triangular, corresponding to the current through the test regulator tube and is taken from
BP3, which is connected to the junction point of the series RC connected to the output of V7, the input to which is a square wave. Thus traces on the oscilloscope screen correspond to the voltage vs. current characteristics of the test regulator tube for a rising and falling current. The amplitude and reference line of the triangular wave of current may each be independently varied by means of potentiometer P4 and potentiometer P1, respectively, or both may be varied so that they maintain a fixed relationship by means of potentiometer P2. It was found that a low frequency was necessary for operation of the instrument (between 12 and 15 cps). It was desirable to synchronize it with the 60-cps line frequency so that any residual hum pickup would not cause walking patterns on the screen. The low frequency required made it essential that the oscilloscope have a good low-frequency response (down to at least 10 cps). Considerable gain in the vertical deflection amplifier is required, which means careful shielding of the input circuit wiring.

14. DROPPING-MERCURY POLAROSCOPE

In the polarograph method of analysis of liquid mixtures, use is made of an electrolytic cell having a dropping-mercury electrode to analyze a liquid mixture with respect to reducible ions and compounds. The dropping-mercury electrode is generally in the form of a capillary tube projecting beneath the surface of the liquid mixture undergoing analysis. To analyze the mixture an electric potential is applied between the dropping-mercury electrode and the liquid mixture, i.e., across the surface of a mercury drop being formed in the liquid at the tip of the capillary. By varying the voltage across the cell, an irregular stepped current-voltage curve is obtained in which the voltage at which each current step appears is characteristic of a reducible component in the mixture and the height of each step is a measure of the quantity of such reducible component present.

With the polaroscope developed in the Radiation Laboratory a polarogram (i.e., a graph of cell current vs. cell voltage) is automatically obtained with a dropping-mercury cell during the period of formation of each drop and is presented for view on the screen of a cathode-ray oscilloscope. In this polaroscope a controller is provided in which the generation of a sawtooth voltage wave is initiated in timed relation to the birth of each new drop. The controller for accomplishing this result is arranged to do three things: (1) to detect and respond to a change in a characteristic of a dropping-mercury cell occurring when each drop falls off the dropping-mercury electrode and a new drop begins to form there, (2) to generate a voltage wave to be impressed on the cell, and (3) to delay the beginning
of the sawtooth voltage wave so that the voltage wave is applied to the newly formed drop at a time when the drop has reached a substantial size and its rate of growth is relatively small and nearly the same as for previous and succeeding drops. Thus with this controller the formation of new drops of a dropping-mercury cell is used to initiate the operation of a voltage generator, the output of which is applied across the cell for a short interval of time during the formation of each drop. The apparatus is so designed that the amplitude of the voltage wave applied across the cell is constant regardless of the dropping rate, and the period of the voltage applied across the cell is a predetermined fraction of the average dropping rate. The cell voltage is applied across one pair of deflecting plates of a cathode-ray oscilloscope, and a voltage proportional to the cell current is applied across the other pair of deflecting plates of the cathode-ray oscilloscope.

14.1 General Operation. Figure 5.60 shows that the polaroscope comprises a negative-feedback amplifier (1) having a negative-feedback loop including a feedback circuit in which a dropping-mercury cell (C) is so arranged that the voltage across the cell bears a simple relationship to the voltage applied to the input of the amplifier. A relatively large feedback resistance (R) is included in series with the cell in the feedback loop, and the voltage across the cell is applied to the amplifier input in series with the applied input voltage supplied by the controller unit. With such an amplifier having a high value of round-trip transmission coefficient \( s13 \), the voltage across the cell is proportional to and almost equal to the amplifier input voltage, and the current flowing through the cell passes through the feedback resistance R, so that the voltage across it is proportional to the cell current. Thus the voltage appearing at the output of amplifier 1 equals the sum of the cell voltage plus the potential drop through the feedback resistance R. Accordingly, in order to produce on a cathode-ray oscilloscope a graph of cell voltage vs. cell current, the amplifier input voltage is applied through a reproducing amplifier (2) to the horizontal deflecting plates (H) and the voltage appearing at the output of amplifier 1 is applied in opposite phase to a suitable pair of terminals of a balanced amplifier (3), so that the output of the balanced amplifier is proportional to the cell current. The output of this balanced amplifier is applied to the vertical deflecting plates (V).

In order to produce cyclic variations of the amplifier 1 input voltage, and hence the cell voltage, use is made of the controller unit. This consists of an initiator responsive to a change in cell characteristic occurring at the time each new drop is formed, a voltage generator for producing a sawtooth voltage wave, and a timing circuit for delaying the application of the sawtooth voltage wave to the cell and
for timing incidental operations. A complete cycle of operation occurs in the period between the instants of breaking off of successive drops. This cycle can be broken up into three successive time intervals: The first interval begins when a drop falls and a new drop begins to form and ends shortly before a voltage wave appears at the output of the voltage generator. The second interval includes the time during which a polarogram is produced (the time the variable voltage is on the mercury cell). The third interval covers the remaining portion of the cycle of operation (from the time the output of the voltage generator goes to zero). The timing circuit produces at the output terminals A and B, respectively, a master control voltage and an auxiliary control voltage. The master control voltage is of the periodic square-wave type which has a positive value during the first and second intervals and is zero during the third interval. The auxiliary control voltage is also of the periodic square-wave type, having a positive value in the second interval and a negative value in the first and third intervals.

The initiator unit, as shown in Fig. 5.60, consists of derivator, drop detector, impulse generator, and dropping-rate meter, connected in tandem. The derivator applies the changes in output of amplifier 1 to the drop detector to render this detector insensitive during the second interval and sensitive during the third interval. When the drop detector has been rendered sensitive, a drop falling from the capillary tip in cell C produces a positive pulse at the output of the detector. This pulse trips the impulse generator, producing a negative pulse of standardized magnitude at the output of the generator. This pulse is applied to the dropping-rate meter, which is designed so that its output is a voltage proportional to the frequency with which pulses are applied to its input. The negative pulse from the generator is also applied to the trigger circuit of the timing unit to initiate a new cycle of operation.

The voltage generator consists of a sawtooth-voltage-wave generator and output circuit connected in tandem. The wave generator produces sawtooth waves during the first and second intervals of the cycle, being rendered operative by the master control voltage from the timing circuit. The period of operation of the wave generator corresponds to the period during which the drop detector is insensitive to the output of the derivator. The sawtooth generator produces waves of constant amplitude irrespective of the dropping rate and of a frequency proportional to the average dropping rate, this frequency being controlled by the voltage output of the dropping-rate meter. The sawtooth waves are applied to the trigger circuit of the timing unit to initiate and terminate the second interval. They are also applied to the output circuit, which is controlled by the auxiliary control volt-
age in order to render it inoperative while the first sawtooth wave is being generated in each cycle (first interval) and operative during the second interval (when it transmits the sawtooth wave).

The timing circuit consists of two trigger circuits (1 and 2), circuit 1 generating the periodic auxiliary control voltage, and circuit 2 generating the periodic master control voltage. Circuit 1 has two input terminals and a pair of output terminals. Alternate negative pulses applied to the input terminals upon termination of alternate sawtooth waves (from the voltage generator) cause an auxiliary control voltage to appear at the output terminals of trigger circuit 1, this voltage being positive during the second interval and negative during the remainder of the operating cycle. Trigger circuit 2 has an output terminal, at which the master control voltage appears, and two input terminals responsive, respectively, to negative pulses from the trigger circuit 1 and from the impulse generator. In this way the master control voltage is changed from a zero value to a positive value by a negative pulse (trigger circuit 1) and from the positive value to zero by a negative pulse from the impulse generator, the zero value being maintained during the first and second intervals and the positive value being maintained during the third interval.

14.2 Individual Circuits. (a) Amplifier. It will be noted from Fig. 5.61a that amplifier 1 comprises three stages including, respectively, amplifying tubes $T_1$, $T_2$, and $T_3$ and a coupling circuit (not shown in Fig. 5.60) including an amplifying tube $T_0$ similar to the tube $T_1$ of the first stage. This coupling circuit is provided in order to facilitate application of the sawtooth-wave voltage in the cathode circuit of the tube $T_1$ of the first stage and may be considered a part of the first stage. The output of amplifier 1 is applied to the signal grid of the tube $T_1$ by means of a suitable connection taken from the feedback network.

The coupling tube $T_0$ and the tube $T_1$ of the first stage are provided with a common cathode resistor; the two screen grids of the coupling tube $T_0$ and the tube $T_1$ of the first stage are connected directly together and are provided with a suitable operating voltage through a common screen resistance, and the two cathodes of tubes $T_0$ and $T_1$ are connected to a common heater supply to eliminate effects due to fluctuations of emission of the cathode of tube $T_1$. The master control voltage is applied through a dropping resistor $R_0$ to the signal grid of the second amplifier tube $T_2$.

In operation, when the voltage applied to the signal grid of the coupling tube $T_0$ changes by a given amount, an equal change occurs in the voltage at the cathode of this tube, thus producing an equal change in the voltage at the cathode of the tube $T_1$ of the first stage.
Because of the negative-feedback connections of the amplifier, an equal voltage change occurs across the cell. Thus, it is seen that a sawtooth voltage wave applied to the coupling tube $T_0$ produces a corresponding voltage wave across the cell. The circuit is such that the feedback ratio caused by changes in the resistance of the cell during the application of a variable voltage adjusts itself automatically to amplify the variable input voltage to a variable output voltage which is equal to the sum of the input voltage and the $IR$ drop through the feedback resistance $R$.

The master control voltage produced by the timing circuit is applied through $R_0$ to the signal grid of tube $T_2$ of amplifier 1, whereby a positive voltage is produced at the output and also across the cell in the third time interval. When the master control voltage is positive, the amplifier output voltage varies slightly as the drop at the tip of the capillary enlarges, and, at the moment that the drop breaks off, the resistance of the cell suddenly increases by a very large amount, causing the effective gain of the negative-feedback amplifier to be reduced and, in turn, causing the voltage appearing at the output of the amplifier to drop.

(b) Initiator. This unit is shown in Fig. 5.61b. The derivator is of simple capacitor-resistance type and operates to produce at its output a voltage proportional to the rate of change of voltage impressed on its input. The master control voltage is applied to the anode of pentode $T_4$ in the drop detector, so that this tube then amplifies signals applied to its input during the third time interval. When a drop falls in the cell C, the input voltage of the derivator is dropped, causing a negative pulse to be applied to the input of the drop detector. The positive pulse output of the detector is applied to the impulse generator, which is of the self-restoring trigger type. The dropping-rate meter consists of an amplifier having a negative-feedback connection for feeding all the output current back to the input through the impedance $R_1C_1$. The apparent input impedance of the amplifier has the same time constant as the $R_1C_1$ network. The input of the amplifier is coupled to the output of the impulse generator through a unilateral impedance (diode $T_9$), so that negative current pulses from the impulse generator through $R_1C_1$ produce a voltage across $R_1C_1$ proportional to the rate of generation of the pulses. In effect, an $R_1C_1$ network of very low impedance is connected between the anode of $T_9$ and ground, so that the voltage between these points is at all times low enough to allow transmission of uniform pulses to the RC network regardless of the dropping rate. The output voltage and the input pulse rate of the amplifier are therefore proportional.

(c) Voltage Generator. The voltage generator (see Fig. 5.62a) consists of a sawtooth-wave generator and an output circuit. The
wave generator is an amplifier with a negative-feedback loop formed by a feedback resistance $R_2'$ and an input resistance $R_2''$ of equal magnitude connected in series with the effectively low output impedance of the dropping-rate meter. Since the $\mu \beta$ of this amplifier is always large throughout the cycle of operation and the feedback ratio is one-half, the output voltage is equal and opposite to the input voltage. Thus the output voltage is proportional to the dropping rate. The amplifier has four stages ($T_4'$, $T_{10}$, $T_{11}$, and $T_{12}$), the first two of which are conventional. The second stage is coupled to the third through $C_3$, across which thyratron $T_{13}$ is connected. The third and fourth stages form an amplifier in which all the output is fed back to the input through $R_3$. Thus the effective resistance between $C_3$ and ground is $R_3$ divided by the $\mu$ of this amplifier. When $T_{13}$ conducts, the voltage across $C_3$ is constant, independent of the dropping rate. When $T_{13}$ is not conducting, $C_3$ is charged by current through $R_3$ at a rate proportional to the dropping rate until the thyratron fires. Since the current to $C_3$ is proportional to the dropping rate and since there is a constant voltage drop between the anode and control grid of $T_{13}$ when this tube is nonconductive, the sawtooth wave generated is inversely proportional to the dropping rate.

The master control voltage is applied through a parallel RC network to the grid of thyratron $T_{13}$, so that sawtooth waves are generated so long as the applied voltage is zero and are not generated so long as this voltage is sufficient to maintain $T_{13}$ conducting. Thus the output voltage of the generator is constant while the master control voltage is positive (third time interval). When $T_{13}$ becomes nonconducting (master control voltage zero), the generator voltage output increases as a linear function of time ($C_3$ charging) until the thyratron fires. This reduces the generator voltage to its initial value. $C_3$ charges and discharges periodically so long as the grid voltage on $T_{13}$ is sufficient to bias the thyratron normally to cutoff. When thus operated, the generator output voltage periodically changes from a relatively low negative value to a relatively high positive value, the total swing being of the order of 100 volts.

The output circuit consists of a cathode-loaded triode $T_{14}$. The input circuit of $T_{14}$ is a potential divider including $R_4$ and $T_{15}$. A second diode $T_{16}$ is included between the grid of $T_{14}$ and the auxiliary control voltage applied at the output of trigger circuit 1. When this voltage is positive the diodes are nonconducting and the sawtooth waves are fully impressed on the grid of $T_{14}$ and thus transmitted to the cathode circuit of $T_{14}$ and the output of the voltage generator. However, when the control voltage is negative, diodes $T_{15}$ and $T_{16}$ are conducting, reducing the resistance of $T_{15}$ and effectively short-
circuiting the input of T14 so that it cannot transmit signals to the output of the voltage generator. T15 short circuits the input to T14 while the output of the sawtooth-wave generator is negative.

(d) **Timing Circuit.** This circuit is comprised of two trigger circuits (1 and 2), as shown in Fig. 5.62b. Circuit 1 consists of a pair of triodes, T17 and T18, which are rendered alternately conducting by alternate negative pulses simultaneously applied to both their grids at the termination of sawtooth waves. The output of T18 is applied to the grid of a cathode-loaded triode T19. When the auxiliary control voltage (taken off the cathode circuit of T19) is positive, it is changed to negative by a negative pulse on the grids of T17 and T18. In the same way negative control voltage is changed to positive. While T19 is highly conductive, a positive voltage is applied across T15 and T16 of the output circuit.

Trigger circuit 2 consists of a pair of triodes, T20 and T21, which are rendered alternately nonconducting by negative pulses applied to their grids. The negative pulses are obtained from trigger circuit 1 (for T20) and from the impulse generator (for T21). The output of T20 is applied to the grid of cathode-loaded triode T22. When the conductivity of T22 is low, the master control voltage (taken from cathode load of T22) is zero, and when its conductivity is high the control voltage is positive.

14.3 **Cycle of Operation.** A cycle of operation will be considered that starts an instant shortly before a mercury drop falls from the tip of a capillary and a new drop starts to form, i.e., shortly prior to the end of the third interval of a normal cycle of operation. At this moment the master control voltage is positive and the auxiliary control voltage is negative. Therefore thyratron T13 is conducting and the voltage output of the sawtooth-wave generator is steady. The drop detector is operative and a steady current flows in the amplifier 1 output circuit, a small portion flowing through cell C. At this moment application of a negative auxiliary control voltage to amplifier 2 deflects the oscilloscope beam out of the polarogram area. As the drop continues to enlarge, the resistance of the cell gradually decreases and the voltage output of amplifier 1 slowly increases.

When the drop breaks off, the resistance of the cell suddenly increases, so that the voltage output of amplifier 1 suddenly decreases, impressing a positive pulse on the drop detector. A negative pulse is produced by the drop detector, which is impressed on the impulse generator. The corresponding standard pulse produced by this generator is applied to trigger circuit 2, reducing the master control voltage to zero and initiating a new cycle of operation. When this happens thyratron T13 becomes nonconducting; consequently the output of am-
plifier 1 becomes zero and renders the drop detector insensitive to changes in the output of this amplifier. At the same time the output of the sawtooth-wave generator begins to increase as a linear function of time. This voltage is applied to the output circuit, which at this moment cannot transfer voltage to the input of amplifier 1 since the auxiliary control voltage is still negative.

When the output of the sawtooth-wave generator reaches its maximum value, thyatron T13 fires, discharging C3, and the second interval is initiated. This reduces the output voltage of the generator to a slightly negative value and puts a negative pulse on the input of trigger circuit 1. This changes the auxiliary control voltage to a positive value. This change renders the output circuit transmissive so that the positive portion of the next sawtooth voltage wave is transmitted through the output circuit to the input of amplifier 1, and at the same time brings the oscilloscope beam back on the screen. During the generation of the second sawtooth wave, a voltage is applied to the input of amplifier 1, and an equal voltage is produced across cell C. The current in the cell depends on the nature and amounts of reducible components in it. As this voltage varies, the voltage across R varies correspondingly. Therefore by applying the input and output of the amplifier in phase opposition to the balanced amplifier 3, a vertical deflecting force is applied to the oscilloscope beam which is proportional to the cell current. At the same time a horizontal deflecting force, proportional to the amplifier input, is applied to the beam, so that a trace is produced which represents cell voltage vs. cell current. During this time the drop detector is insensitive.

When the voltage output of the sawtooth generator falls to zero at the end of the second interval, a negative pulse applied to trigger circuit 1 lowers the output of this circuit. This changes the auxiliary control voltage to a negative value and simultaneously puts a negative pulse on trigger circuit 2 (T20), swinging the master control voltage to a positive value. The output circuit is then rendered nontransmissive and the electron beam is deflected off the screen. At the same time the sawtooth-wave generator is rendered inoperative and the drop detector sensitive, and a small positive voltage appears at the output of the amplifier, placing the entire system in a condition ready for the next cycle of operation. A clearer picture of the whole cycle of operation can be obtained by reference to Fig. 5.63.
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