

**INDUSTRIAL ELECTRONICS
HANDBOOK**

INDUSTRIAL ELECTRONICS HANDBOOK

BY

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WITH A FOREWORD OF

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FOREWORD

The word electronics has, only very recently, made its first appearance in the Oxford English Dictionary. Few modern words are, however, more firmly established in use, either colloquially or in the literature of applied science. The utility of the word is unquestionable, but its very usefulness is already rendering it difficult to limit it satisfactorily in scope and application.

Essentially the subject of electronics deals with electric circuits in which the current is controlled in one respect or another by devices such as thermionic valves, gas filled relays, or photo-electric cells, but there are now, in addition, the rapidly developing techniques of current control by semi-conductor devices such as the transistor. This book is concerned with industrial electronics and, by accepted convention, this means electronic circuits in applications other than those for purposes of telecommunication. Such a division is not only a practical necessity but it has long been recognised in the general field of electrical engineering that power and telecommunication constituted the two main and distinct branches of the applications of electricity. Overlapping territories have, however, become a common feature of the modern scientific scene and so with electronics there can be no clear boundaries of demarcation. Automatic computers which have been described as being concerned with the "processing" of information are already often associated with electromechanical systems of considerable power and no doubt in the future will be increasingly utilised in the automatic control of industrial processes. The field taken by the author of this book as his province is one which, important as it already is, must increase in influence as time goes on, for, within it, lies the most potent factor for the increasing of productivity in industrialised communities. That is a matter, the importance of which to the progress of civilisation, needs no argument or demonstration.

The reader will find in the present book first a sound treatment of the scientific principles of vacuum and gas-filled electronic tubes and their basic circuits. It is the second part of the book which deals with applications and these have been judiciously selected to illustrate in a clear fashion the more important principles of application which have so far been developed.

The book may be said to constitute a most satisfactory and trustworthy introduction and guide to a branch of electronic work which already has an extensive specialised literature. Coming, in authorship, from an organisation which has been associated in such a distinguished way with research and development in the electronic field, it is natural to expect to find a treatment of the subject which is both illuminating and authoritative and in this expectation the reader will not be disappointed.

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PREFACE

Developments during the past few years have abundantly proved how great are the advantages, both technical and economic, which can be gained by the adoption of electronic methods, that is to say processes and methods of control based upon the application of electronic tubes in their many and diverse forms.

It must be admitted, however, that in certain industries the introduction of applied electronics is but slow. This tardiness is almost certainly due to the fact that there are still many engineers and technologists who have not, as yet, been fully informed as to the possibilities, or even the nature, of industrial electronics. It is quite natural that plant and production engineers in, for example, the textile, steel and chemical industries, should view with a certain reserve, amounting almost to suspicion, processes and methods with which they are quite unfamiliar, and will show some reluctance to the adoption of an electronic aid in place of some older and well-understood device.

It was mainly this reason that induced me to write this book, which is primarily addressed to works engineers engaged in every branch of industrial production. It is also hoped that those studying for entry into the technical professions will also find the book a useful elementary introduction to the whole subject of industrial electronics, a knowledge of which is today essential to every industrial engineer.

To this end, much thought has been devoted to the methods of exposition and presentation. It has, for example, been considered especially important to include detailed descriptions and explanations of a number of circuits which have been used in industrial practice and have proved their worth. In many cases very comprehensive data and specifications of the values of the component parts have been included.

I am greatly indebted to numerous firms engaged in industrial electronics for permission to reproduce photographs and to quote from their technical publications. Special thanks are due to Philips, Eindhoven, for their valuable support, without which this book would never have been published. I also wish to express my thanks to Mr. Harley Carter, A.M.I.E.E. of Mullard Ltd., London, who prepared the English text, and to Mr. H. E. Kater of the Electronic Tube Division of Philips, Eindhoven, for his assistance in supervising the general lay-out and the production of the illustrations.

Hamburg, October 1953

The Author

TO THE SECOND EDITION

This second edition has been considerably enlarged and is again entirely up to date. The recently introduced E 1 T decade counter tube, for example, is described in detail. Several counter circuits equipped with this tube and various applications of these circuits are discussed. Full particulars are given on the design of industrial

rectifier circuits, including the design of controlled rectifiers. The section on electronic motor control has been extended by including, for instance, a description of electronically controlled Ward-Leonard systems. Finally various types of electronic apparatus for special purposes have been dealt with, such as stabilized supply units, electronic dust precipitators, and so on.

As a result of inserting this new material, the present edition is considered to be so exhaustive as to justify the addition of the word HANDBOOK to its title.

I wish to make a special acknowledgement to Mr. H. E. Kater of the Electronic Tube Division and Mr. D. J. Mitchell of the Translation Department – both of Philips Eindhoven – under whose care this edition was prepared.

Hamburg, October 1956.

The Author

PREFACE TO THE THIRD EDITION

As the second edition of the Industrial Electronics Handbook, like the first, has sold out within so short a time, the Philips Technical Library has decided to issue a third edition. In view of the increasingly important role played by active semiconductor elements in industrial electronics, it seemed a suitable opportunity to add a short paragraph regarding the operation and the elementary circuit technique of transistors. A few additional special circuit arrangements have, moreover, been included. I hope that the third edition will meet with the same success as the two previous editions.

Hamburg, October 1958.

The Author

CONTENTS

Foreword	V
Preface	VI
Introduction	3

Part I. The Tubes and their Basic Circuits

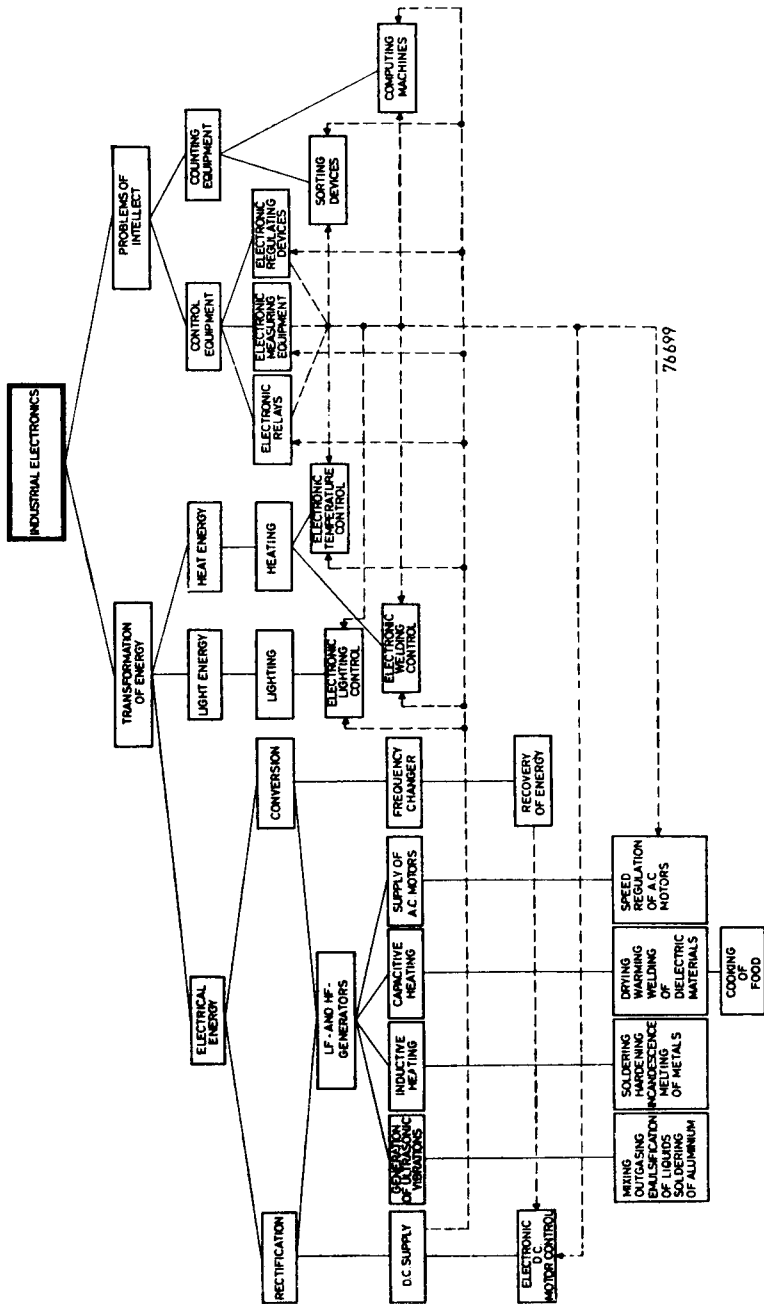
1. AMPLIFYING AND TRANSMITTING TUBES	5
2. RECTIFYING TUBES	17
3. THYRATRONS	29
4. SENDITRONS.	46
5. IGNITRONS AND EXCITRONS.	50
6. VOLTAGE STABILISING TUBES	59
7. PHOTOCELLS.	65
8. TRIGGER TUBES.	71
9. CATHODE-RAY TUBES.	75

Part II. Electronic Devices for Industrial Purposes

10. ELECTRONIC RELAYS	85
Simple photo-electric relay.	85
Selective photo-electric relay.	86
Relay with pre-amplifier	87
Photo-electric door control.	89
Automatic control of oil-fired boilers.	91
Electronic smoke detector.	93
Automatic examination of food cans	94
Automatic reversing control for electric motors	94
11. ELECTRONIC COUNTING CIRCUITS	95
"Scale of two" circuit.	96
Ring counting circuit	96
The Eccles-Jordan ("flip-flop") circuit.	98
Circuits with decade counter tubes	101
Measuring shutter speeds of cameras	107
Predetermined counter	109
Multiplying circuit	112
Subtracting and dividing by means of decade counter tubes	115
Reading out of the number of counts	115
12. ELECTRONIC TIMERS	116
Timer for photographic enlarger	117
Timer giving two successive intervals.	119
Continuously acting timer	120
Another timer giving two successive intervals.	120
Another single-stroke photographic timer	120
Timer for small electric spot-welder	121
Alternative welding timer circuit.	122
Timer for controlling four stages in the welding operation.	122
Another welding timer producing four time intervals	125

13. INDUSTRIAL RECTIFIER CIRCUITS	126
Two-phase half-wave rectification	127
Controlled two-phase half-wave rectification	130
Three-phase rectification	135
Three-phase controlled rectification	137
Calculation of the value of the smoothing choke	138
Influence of the grid control	139
Simple two-phase battery charger	142
A larger battery-charging equipment	144
Rectifier with constant output voltage	144
Automatic voltage control with saturable core reactor	146
Transformerless 3-phase rectifier	151
Large rectifiers for electronic motor control systems	154
14. ELECTRONIC DIMMING OF LAMPS	157
Electronic circuit for dimming fluorescent lamps.	160
Installation for keeping the lighting constant	164
Electronic stage lighting circuit	166
15. SPEED AND TEMPERATURE CONTROL.	168
Paper register control.	168
An alternative register control	170
Lateral register control	173
Variable high-speed drive	177
Temperature control.	180
Temperature regulation	183
16. ELECTRONIC CONTROL OF RESISTANCE WELDING	185
Simple electronic welding contactor	186
Electronic control of small spot welders	188
Four-time spot welding control	190
Control of seam welding	193
Obtaining an even number of half cycles during welding	198
Pulsation welding.	201
Three-phase welding equipment	204
17. ELECTRONIC MOTOR CONTROL	208
Simple single-phase motor control systems	213
Three-phase motor control with armature and field control.	216
Constant-torque drive	219
Constant-power drive	220
Reversal and braking	221
Electronic control of Ward-Leonard systems	233
18. HIGH-FREQUENCY INDUCTIVE HEATING OF METALS.	236
19. HIGH-FREQUENCY CAPACITIVE HEATING OF DIELECTRIC MATERIALS	245
5 kW high-frequency generator.	252
H.F. generator with high frequency stability.	254
Automatic load circuit tuning device	258

20. ELECTRONIC APPARATUS FOR SPECIAL PURPOSES	266
Electronic stabilisation of direct and alternating voltages	266
Ultrasonic soldering iron	269
Photo-electric revolution counter	271
Low-power two-phase inverter	276
Electronic dust precipitators	278
21. TRANSISTORS.	282
General design and working principle	283
Multivibrator circuit with transistors	287
Periodic power switch	287
Phase-sensitive power switch	288
Phase-independent power switch	289
Photoelectric relay	290
Stabilized power-supply unit for transistor circuits	292
Conclusion.	294
Principal data of some preferred types of Philips tubes	296
Subject matter index	301
Bibliography.	307



INTRODUCTION

This book deals in particular with electronic circuits employing highly evacuated or gas-filled discharge tubes, i.e. devices in which electrons move through free space instead of in a metallic conductor. Such devices include not only the familiar amplifying tube so widely used in radio and television engineering, but also photocells, cathode-ray tubes, X-ray tubes, and gas-filled rectifying tubes with or without grid control; even fluorescent lamps should be classed as electron tubes.

During the last ten or twenty years the potentialities of electronic techniques have gained increasing recognition, especially in industry. These techniques make possible improved and more efficient manufacturing methods and the accurate control and regulation of almost every kind of process.

For example, by means of electronic control, mechanical drives can be given almost any desired speed-torque characteristics; the control apparatus being to all intents and purposes inertialess, it is practically instantaneous in action. Moreover it consumes negligible power. Feed drives of machine tools; multi-motor drives in rolling mills; spinning machines; wire drawing mills; lifts and many other drives may be given the required characteristics by means of electronic control. Electronic circuits incorporating photocells or other pick-up devices can be employed for signalling, sorting, inspection or counting in mass-production plants, in chemical processes, in transportation equipment and in other industrial operations. Electronic timers in conjunction with electronic switches can control the process timing of welding machines with high precision, and thus contribute to improved quality. Electronically generated H.F.-energy offers new possibilities in the woodworking and plastic industries for the economical production of furniture, plywood and plastic articles. Steel hardening and soldering or smelting of metals by H.F.-energy increases the economic production of metal goods and leads to improved quality.

Generally speaking, electronics can assist the industrial engineer in the solution of two fundamental types of problem; (1) the transformation of electrical energy into other forms of energy; and (2) the execution of processes analogous to the exercise of intellect, for instance, measuring, counting, sorting, etc.

A similar division of problems into two fundamental groups can equally be found in other fields, for example, in medicine (therapeutics and diagnosis), in lighting (illumination and indication) and in telecommunications (carrier generation and modulation). For industrial electronics, these two groups of problems are depicted in the diagram on the opposite page, which also indicates a large number of practical applications and their mutual relationships. Most of these applications are dealt with in this book.

As result of the growing importance of electronics in the industrial field, it has been necessary to develop tubes specially designed for industrial use. Radio and telecommunications call mainly for voltage amplifying tubes, for

power amplifiers of comparatively low output, and for tubes capable of generating unmodulated or modulated H.F. oscillations of medium power. These requirements are easily met by high vacuum amplifying and transmitting tubes. In industry, however, control or switching operations on comparatively large currents will very often be needed. For these purposes high vacuum tubes are less suitable, and gas-filled tubes and valves of various kinds are then employed.

Both vacuum and gas-filled tubes and many other electron devices are used in industrial applications, and the first part of this book is therefore devoted to these tubes and their basic circuits. In the second part practical circuits and their applications are described.

PART I

The Tubes and Their Basic Circuits

1. AMPLIFYING AND TRANSMITTING TUBES

The essential elements of a high vacuum tube are a cathode and an anode, which are mounted facing another and sealed in an evacuated bulb.

The cathode consists of a material which readily emits electrons when heated. The cathode of transmitting tubes often consists of a wire of pure or thoriated tungsten which is heated by a low voltage source to the temperature required for electron emission. Such cathodes are said to be directly-heated. The cathodes of amplifying tubes are usually formed of a nickel tube coated externally with a mixture of alkaline oxides — particularly barium and strontium oxide. Heating of this type of cathode is by a tungsten filament located within the nickel tube, which is therefore termed an indirectly-heated cathode.

The anode consists of metal or occasionally graphite. Its function is to collect the electrons emitted by the cathode. To ensure this, the anode is maintained at a positive potential with respect to the cathode.

In most types of tube one or more “grids” are located between the cathode and anode. These are wire spirals of suitable shape and size. Such a grid will govern the density of the electron stream flowing from the cathode to the anode, according to its potential with respect to the cathode. If, for instance, the grid potential is negative, the electron current may be greatly reduced or even entirely suppressed.

Fig. 1-1 shows the conventional diagrammatic representation of typical tubes, that on the left being the symbol for a directly-heated triode and that on the right an indirectly-heated pentode.

The construction of a pentode, i.e. an amplifying tube containing three grids surrounding the cylindrical cathode, is shown in fig. 1-2.

If the anode of a three-electrode tube (containing cathode, anode and one grid) is maintained at a positive potential (the anode voltage V_a) and a variable negative voltage V_g is applied to the grid, the value of the current I_a flowing to the anode can be plotted as a function of the grid voltage. A typical “ I_a-V_g ” diagram is reproduced in fig. 1-3. From this curve the amount by which anode current will change, if the grid voltage

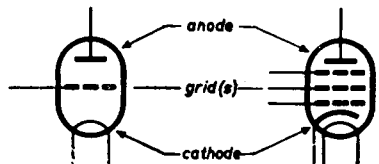


Fig. 1-1. Tube symbols (left: directly heated triode, right: indirectly heated pentode).

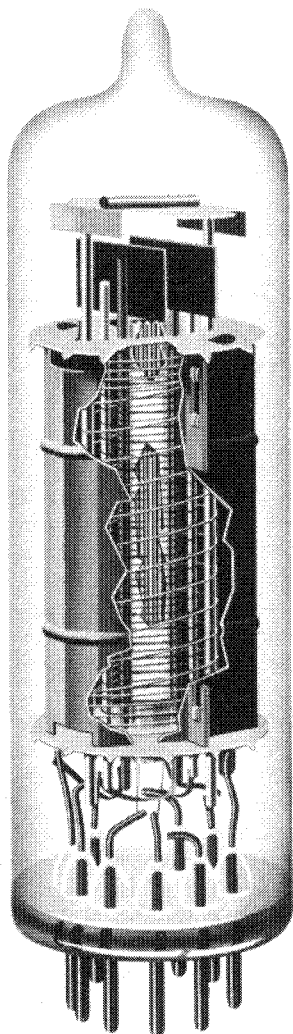
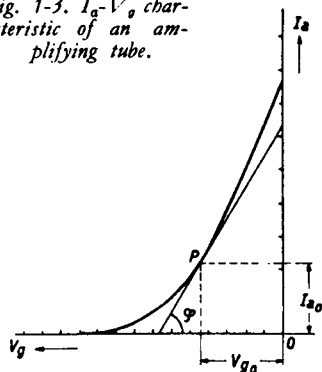


Fig. 1-2. X-ray photograph of a pentode amplifying tube (EL 84).

Fig. 1-3. I_a-V_a characteristic of an amplifying tube.



is modified by a given amount, can be ascertained. Obviously this relation is equal to the slope of the curve, i.e. the tangent of the angle formed by the positive x -axis of the graph and the tangent to the curve constructed at the point under consideration. The slope S may thus be defined as:

$$S = \tan \varphi = \frac{dI_a}{dV_a}, \quad (1.1)$$

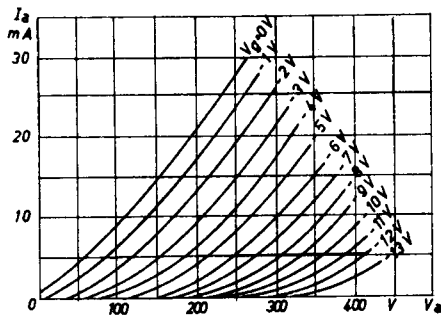


Fig. 1-4. I_a-V_a characteristics of an amplifying triode.

assuming the anode voltage to be held constant.

The slope has the dimension of a conductance, i.e. the reciprocal of a resistance, and is usually expressed in mA/V.

If the grid voltage is held constant, and the anode current is plotted as function of the anode voltage, the " I_a-V_a " diagram can be obtained (fig. 1-4) with V_g as parameter. Obviously the I_a-V_g diagram of a tube can easily be derived from its I_a-V_a diagram. If a given point is chosen on an I_a-V_a diagram and the tangent to it is constructed, the cotangent of the angle formed by the tangent

and the positive x -axis of the graph represents the internal resistance of the tube at the given working point. It can be defined as

$$R_i = \frac{dV_a}{dI_a}, \quad (1.2)$$

the grid voltage being held constant. It should be pointed out that this resistance is the a.c. resistance of the tube, and must not be confused with the d.c. resistance which is equal to the quotient of the anode voltage and anode current at the working point, i.e.

$$R_{i-} = \frac{V_a}{I_a}. \quad (1.3)$$

The basic circuit of a tube operated as an amplifier is shown in fig. 1-5. The

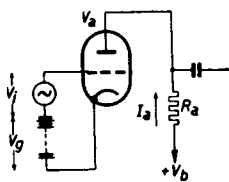


Fig. 1-5. Basic circuit of an amplifying stage.

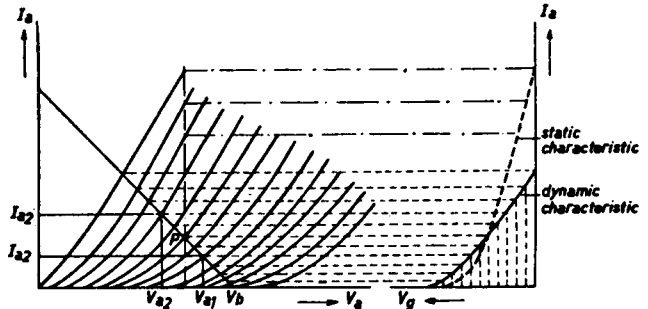


Fig. 1-6. I_a - V_a characteristics of a triode with load line and dynamic I_a - V_g characteristic.

control grid is given a fixed negative bias V_g ; in addition the alternating input voltage V_i , which is to be amplified by the tube, is applied to the grid. The anode circuit contains an external resistance R_a across which there will be a voltage drop $I_a \cdot R_a$ depending upon the instantaneous value of the anode current. This voltage therefore varies in sympathy with the changing value of the input voltage V_i . The effective anode voltage is therefore:

$$V_a = V_b - I_a \cdot R_a, \quad (1.4)$$

where V_b is the direct supply voltage.

Equation (1.4) shows that I_a is a linear function of the anode voltage V_a , which may be plotted in the I_a - V_a diagram as a straight line (fig. 1-6). Evidently V_a is equal to V_b only when $I_a = 0$, and the effective anode voltage decreases as the anode current increases. It is therefore clear that the "static" I_a - V_g characteristic is not applicable when a signal is being amplified, and must be replaced by a "dynamic" characteristic of reduced slope.

The gain obtainable with a circuit as shown in fig. 1-5 is:

$$G = \frac{I_a \cdot R_a}{V_i}. \quad (1.5)$$

The alternating anode current I_a *) flows through the tube and the anode re-

*) Henceforth I_a and V_a symbolise the *alternating* anode current and voltage without direct component.

sistance R_a , and at the anode appears an alternating voltage V_a which is shifted in phase by 180 degrees with respect to I_a and V_i . The alternating anode current caused by V_i is equal to $S \cdot V_i$ according to equ. (1.1); furthermore the alternating anode voltage produces an alternating current V_a/R_i corresponding to equ. (1.2). These currents are additive, so that

$$I_a = S \cdot V_i + \frac{V_a}{R_i}. \tag{1.6}$$

Now

$$V_a = -I_a \cdot R_a \tag{1.7}$$

(by reason of the 180° phase shift), so that

$$I_a = S \cdot V_i - \frac{I_a \cdot R_a}{R_i}, \tag{1.8}$$

and from equation (1.5) the gain is:

$$G = S \cdot \frac{R_i \cdot R_a}{R_i + R_a}. \tag{1.9}$$

If $R_a \gg R_i$,

$$G = S \cdot R_i = \mu. \tag{1.10}$$

μ is called the amplification factor of the tube, and represents the theoretical limit of voltage amplification obtainable from the tube. This limit can, however, never be achieved because in practice $R_a/(R_i + R_a)$ is always less than unity.

To obtain a large voltage gain a tube with a high amplification factor, i.e. with high slope and high internal resistance should be chosen. The slope depends almost entirely on the geometry of the electrode system and cannot be increased indefinitely under usual production conditions. The internal resistance, however, can be considerably increased if a second grid, the screen grid, is mounted between the control grid and the anode and is maintained at positive potential. Fig. 1-7 shows the I_a-V_a characteristics of such a tetrode. The flat trend of the curves in the range of higher anode voltages shows that in that region the internal resistance has a considerably higher value than that of the triode as indicated by fig. 1-4.

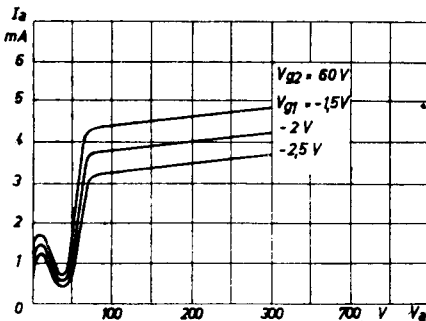


Fig. 1-7. I_a-V_a characteristics of a tetrode.

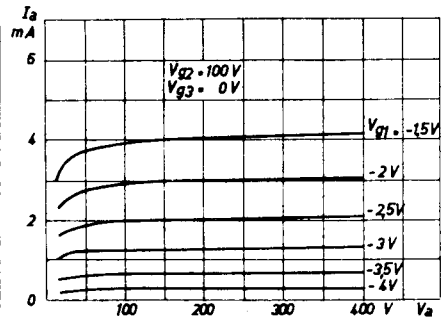


Fig. 1-8. I_a-V_a characteristics of a pentode.

On the left hand side of the diagram the internal resistance becomes negative within a certain range. This is due to secondary emission from the anode. To avoid this effect, a third grid is placed between screen grid and anode, thus forming a pentode. This "suppressor" grid is maintained at cathode potential, thus preventing the secondary electrons from flowing to the screen grid. Fig. 1-8 shows the " $I_a - V_a$ " diagram of a pentode.

As for pentodes $R_i \gg R_a$, equation (1.9) may be simplified to

$$G = S \cdot R_a. \quad (1.11)$$

In fig. 1-9 is shown a typical pentode amplifying tube and the construction of its electrode system.

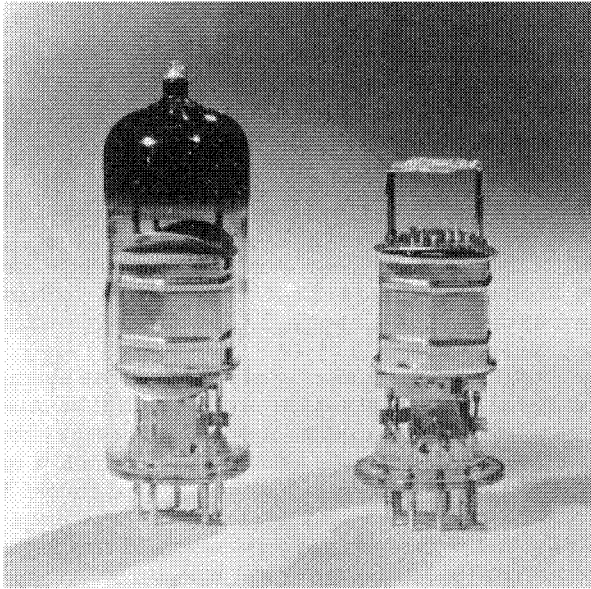


Fig. 1-9. E 80 F pentode amplifying tube. On the right the bulb has been removed to show details of construction.

High vacuum tubes are used in industrial equipment not only for voltage amplification, but also for generating oscillations — often of considerable power output. For this application triodes are generally used or, in special cases, tetrodes. Generation of oscillations is achieved by compensating the attenuation of an oscillatory circuit, tuned to the desired frequency, and connected in the anode circuit of the tube. This so-called tank circuit consists of a capacitor C and an inductor L , furthermore the circuit possesses ohmic resistance R which symbolizes the circuit losses (see fig. 1-10).

Assuming the capacitor is charged with the polarity as shown, it can discharge via R and L , the inductor producing an electromotive force (e.m.f.) which tends to maintain the flowing current. This current then charges the capacitor

with reverse polarity and the process repeats *). The current flowing during each discharge of the capacitor produces a voltage drop across the resistance R and the energy dissipated in R "damps" or progressively decreases the amplitude of the oscillations. The larger the resistance R , the faster the oscillations decay. Fig. 1-11 shows such a damped oscillation. The resulting frequency is termed

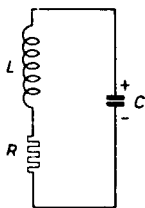


Fig. 1-10. Basic oscillatory circuit.

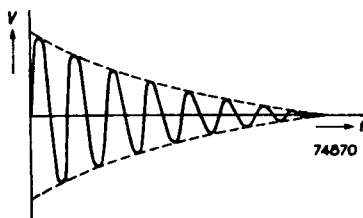


Fig. 1-11. Damped oscillations.

the "resonant frequency" of the oscillatory circuit; it is given by the Thomson formula:

$$\omega_0 = \frac{1}{\sqrt{L \cdot C}} \quad (1.12)$$

For the sake of simplicity, however, the loss resistance of the circuit has been neglected in this formula. As the influence of this resistance upon the generated frequency is only small, formula (1.12) is sufficiently accurate for most practical cases.

If the voltage drop across R is compensated by applying an alternating voltage of equal amount but of opposite phase, the oscillations will be maintained. This can be done by a circuit such as that shown in fig. 1-12. The tank circuit consisting of L_1 and C_1 is connected in the anode circuit of the tube, and a second coil L_2 is inductively coupled with L_1 and is included in the grid circuit. The magnetic alternating field produced by L_1 induces an alternating voltage in L_2 . This voltage controls the grid of the tube, and produces the alternating anode voltage required to compensate the attenuation of the tank circuit.

The tank circuit presents an impedance for the alternating anode current I_a . At resonance (i.e. if the frequency of the alternating current is equal to the resonant frequency of the tank circuit), this external anode load will be

$$R_a = \frac{L}{C \cdot R} \quad (1.13)$$

It then has the character of an ohmic resistance, i.e. the current and the voltage across it are in phase. For the sake of simplicity this condition will be assumed in the following explanation. A part of the alternating voltage V_a appearing across the external load R_a will be fed back to the grid circuit. This feed-back voltage can be written

$$V_i = -k \cdot V_a \quad (1.14)$$

*) The oscillatory circuit consisting of capacitance and inductance can be compared with the balance of a watch, C corresponding to the helical spring and L to the flywheel. R is comparable to the friction resistance.

The negative sign is necessary if the oscillations are to be maintained because, as already explained, the phase of the input voltage has to be shifted by 180 degrees with respect to the anode voltage. The feed-back factor k denotes the amount of coupling between anode and grid circuits necessary for maintaining the oscillations. Its value can be calculated from equations (1.6) and (1.7).

Thus
$$k = \frac{1}{S} \cdot \frac{R_i + R_a}{R_i \cdot R_a}, \tag{1.15}$$

or

$$k \cdot S \cdot \frac{R_i \cdot R_a}{R_i + R_a} = 1. \tag{1.16}$$

It is essential that, when the self-excited oscillations start, the left side of equation (1.16) is greater than unity. This is ensured automatically by the fact that, when first switched on, the tube has no negative grid bias, and therefore operates on a part of the $I_a - V_g$ characteristic at which the slope is comparatively high. Due to the occurrence of grid current the capacitor C_g is charged by the positive half cycles of the alternating grid voltage with the polarity shown in fig. 1-12, and will then discharge slowly through the resistor R_g . The voltage drop across this resistor constitutes the negative grid bias of the tube. This bias will increase if

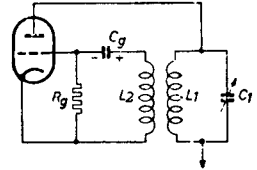


Fig. 1-12. Basic circuit for a tube as oscillator.

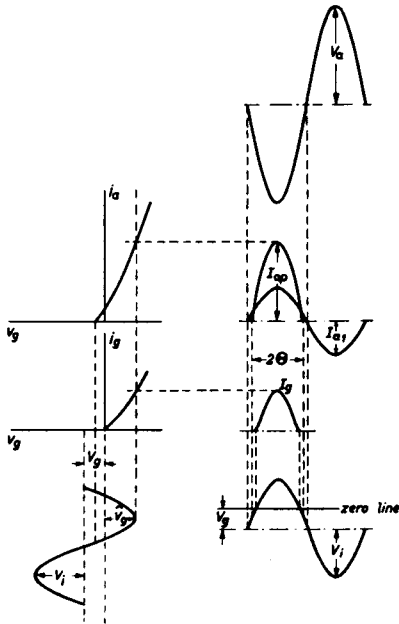


Fig. 1-13. Class C operating condition of a self-excited tube oscillator.

the amplitude of the alternating grid voltage increases, and the working point on the $I_a - V_g$ characteristic is therefore shifted into a region of lower slope until equation (1.16) is satisfied, and the alternating grid and anode voltages assume constant values. Usually the working point is then situated so far down in the region of negative grid voltage that only the crests of the positive half cycles of grid voltage drive the grid positive, causing grid current to flow (see fig. 1-13). Similarly, anode current flows only during a part of the positive half cycles of the alternating grid voltage.

This condition is termed "class C operation". It has the advantage of high efficiency and is therefore used in practically all H.F. generating circuits for industrial service.

It should be distinguished from what is termed "class B operation", in which the working point lies approximately at the lower bend of the $I_a - V_g$ characteristic, so that

anode current flows during the whole of the positive half cycles of the grid voltage. The efficiency of class B operation is much lower than that of class C, but

the generation of harmonics, i.e. the distortion of the amplified voltage caused thereby, is considerably smaller. Class B operation is therefore used in power amplifiers for modulated H.F. (pre-stage modulated radio-transmitters).

The working point of a "class A" amplifier lies on the straight part of the I_a-V_g characteristic. In this case anode current flows during the full cycle of grid voltage. Such amplifiers are used for H.F. or L.F. voltage amplification. In class A operation the amplitude of the grid voltage is usually limited to a value at which grid current does not flow and therefore no driving power is needed.

However, there is no need to employ the comparatively inefficient class A mode in oscillators and power amplifying transmitting circuits, because sufficient driving power for class C or class B operation is always available either by feedback from the tube itself or from a separate driver stage. Transmitting tubes are therefore constructed in such a way that they may be operated at higher grid voltages at which grid current flows.

Operating conditions for transmitting tubes may be well illustrated by V_g-V_a diagrams which are known as "constant current" characteristics, they show the curves of constant anode and grid current for the tube as a function of the grid and anode voltages. An idealized diagram is plotted in fig. 1-14. For the sake

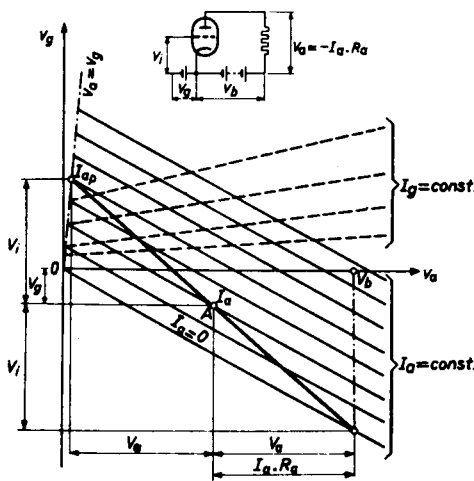


Fig. 1-14. Idealized V_g-V_a characteristics of a transmitting tube.

of simplicity an ohmic resistance R_a is assumed as the load in the anode circuit of the tube. Across R_a appears a voltage drop corresponding to the value of the anode current I_a . If this current is zero the voltage drop is also zero and the first point of the load line can be plotted on the anode current curve $I_a = 0$ at the anode (supply) voltage V_b . The further points may be constructed according to equation (1.4). To prevent the anode voltage from dropping below the grid voltage, which would give rise to excessive grid currents (see below), the maximum peak anode current is limited by the line $V_a = V_g$. The load line, which is a straight one in this case, will be traversed once in each direction during each cycle

of the alternating grid voltage. Point A is the working point of the tube under class A conditions. For every instant the corresponding values of voltages and currents may be taken from the graph; the instantaneous values of grid current and anode current are determined by the intersections of the load line and the curves; the corresponding values of grid voltage and anode voltage can be read off the y - and x -axis of the graph.

The conditions under class C operation are shown in fig. 1-15 by the V_g-V_a diagram of the Philips triode transmitting tube TBL 6/6000 (see fig. 1-17). The anode load is formed by a tank circuit tuned to the operating frequency; it

therefore represents, at this frequency, an ohmic resistance R_a according to equation (1.13). The load line is therefore again a straight line. In class C operation the working point, A in fig. 1-15, corresponds to such a large value of negative grid voltage V_g that anode current flows during only a part of the positive half cycles of the grid voltage, i.e. from B to C . The anode current impulses during this time contain the fundamental wave of the anode current I_{a1} and a large number of harmonics. The angular duration of anode current flow Θ , which corresponds to half the time during which current flows, measured in degrees, can be ascertained easily from the diagram. It would obviously be 90 degrees if current flowed during the complete half of the load line AC . As, however, current flows only during the part BC ,

$$\cos \theta = \frac{AB}{AC} = \frac{DE}{DF}. \quad (1.17)$$

In the case of fig. 1-15, Θ is 67 degrees.

From Θ and the anode peak current I_{ap} the amplitude of the fundamental current wave I_{a1} can be determined. Normally,

$$I_{a1} = I_{ap} \cdot f_1(\theta). \quad (1.18)$$

In the case of a tube with an idealized linear $I_a - V_g$ characteristic (this simplifying assumption can be made in most cases) Fourier analysis of the anode current impulses gives:

$$f_1(\theta) = \frac{\theta - \frac{1}{2} \cdot \sin 2\theta}{\pi (1 - \cos \theta)}. \quad (1.19)$$

This function is plotted in the graph of fig. 1-16.

The alternating anode voltage V_a may be taken easily from the diagram fig. 1-15; in this case it will be nearly 5.6 kV. With $f_1(\theta) = 0.42$ and $I_{ap} = 4.8$ A, $I_{a1} = 2$ A. The output power of the tube

$$W_o = \frac{I_{a1} V_a}{2}, \quad (1.20)$$

which in the case considered gives 5.6 kW approximately.

The grid direct current I_g can be determined

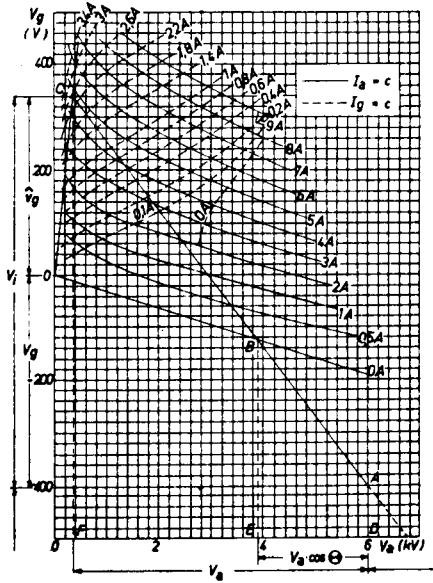


Fig. 1-15. Constant current characteristics of the Philips TBL 6J6000 transmitting triode with load line.

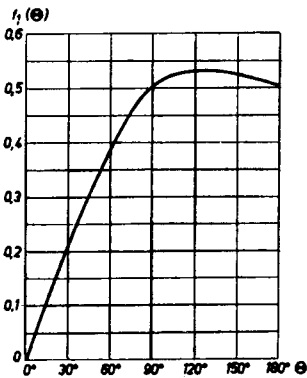


Fig. 1-16. Graph showing $f_1(\theta)$ as a function of θ .

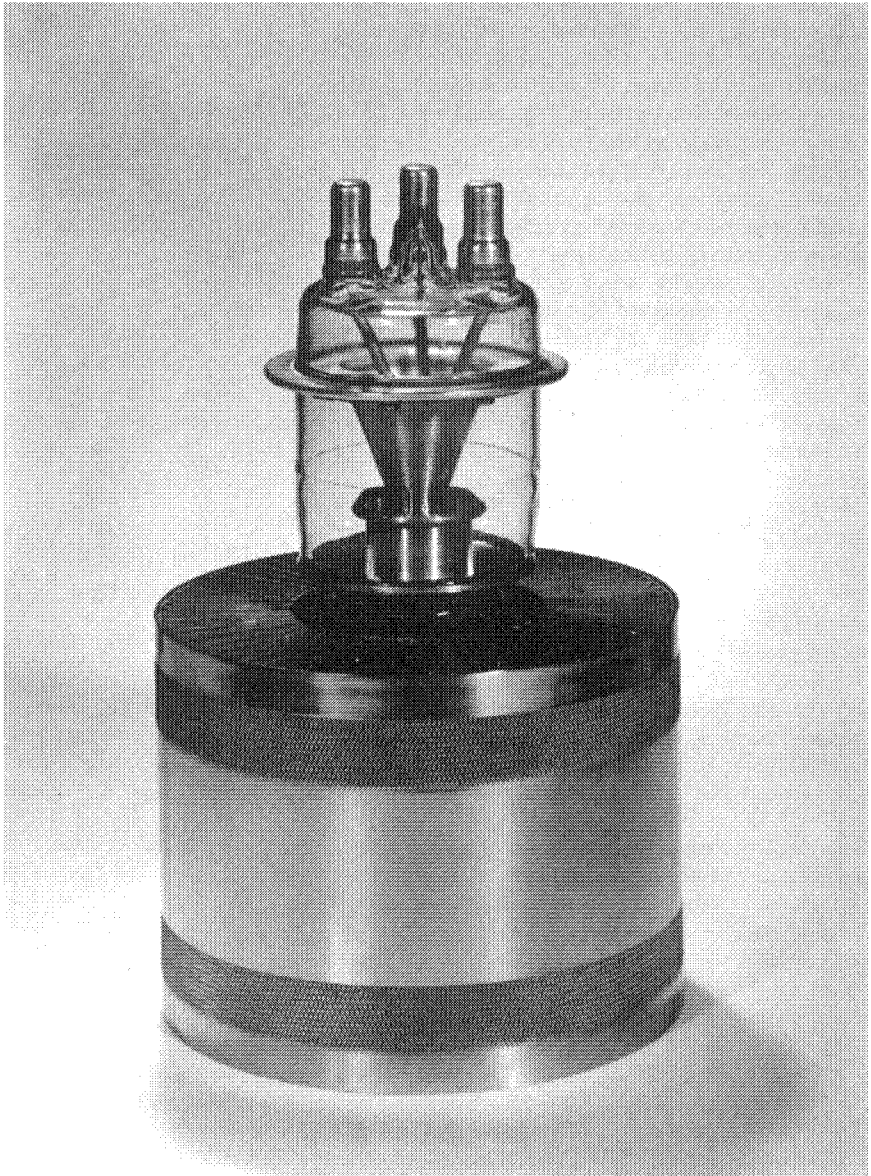


Fig. 1-17. Philips transmitting triode TBL 6J6000.

by integrating the grid current curve which can be constructed from the instantaneous values of i_g derived from the graph:

$$I_g = \frac{1}{2\pi} \cdot \int_0^{2\pi} i_g \cdot d\varphi. \quad (1.21)$$

Similarly the necessary input power will be

$$W_{i_g} = \frac{1}{2\pi} \cdot \int_0^{2\pi} i_g v_g \cdot d\varphi \approx 0.9 V_g I_g, \quad (1.22)$$

where v_g denotes the instantaneous value of the alternating grid voltage.

When an electrode, anode or grid, is positive with respect to the cathode, the possibility of current flowing to this electrode exists. When electrons strike the electrode they release their kinetic energy in the form of heat. This heat thus generated is called the power dissipation and must be removed to avoid overheating and finally destruction of the tube. This can be done in several ways. Small tubes can be cooled sufficiently by natural heat radiation which may be facilitated by giving the electrode a suitable form and surface. To dissipate larger quantities of heat this form of cooling does not suffice, and forced cooling by water or air must be used. For this reason large transmitting tubes are often so constructed that the anode, which has the largest dissipation, forms a part of the vacuum-tight envelope of the tube. On the external side of the anode a cooling radiator may be mounted. Fig. 1-17 shows a commercial air-cooled transmitting triode with external anode for industrial service (Philips TBL 6/6000).

For every tube the maximum ratings of anode and grid dissipation are published by tube manufacturers. These ratings should never be exceeded; otherwise the life of the tube will be greatly reduced. The danger of overloading occurs, for example, with transmitting tubes operating in industrial H.F. generators if the load varies greatly, i.e. if the generator is operated under no-load condition or short-circuited. In the latter case the resistance of the tank circuit becomes very small and the efficiency of the tube decreases until oscillations are no longer maintained. Consequently the anode dissipation increases to a point at which the maximum rating may be exceeded. On the other hand, under no-load conditions the anode impedance will rise considerably, and in consequence the amplitude of alternating anode voltage also increases. Due to feed-back, the alternating grid voltage rises and grid current may exceed the limiting value. This effect will be aggravated by the fact that grid and anode voltages are out of phase by 180 degrees, so that the anode may be at a lower potential than the grid during the positive half cycles of grid voltage. The grid current may then increase until the grid wires melt and the tube is destroyed.

Even if, as in modern industrial H.F.-equipment, fuses and protection circuits are included to switch off the generator in case of overloading, it is recommended that the above possibilities are kept in mind when operating a generator under severe conditions.

Estimation of the grid loading is comparatively simple. Obviously the grid dissipation is equal to the difference between the H.F. driving power and the power losses in the grid voltage source or the grid resistor. Thus:

$$W_g = W_{i_g} - I_g V_g, \quad (1.23)$$

I_g being the d.c. grid current and V_g the grid biasing voltage.

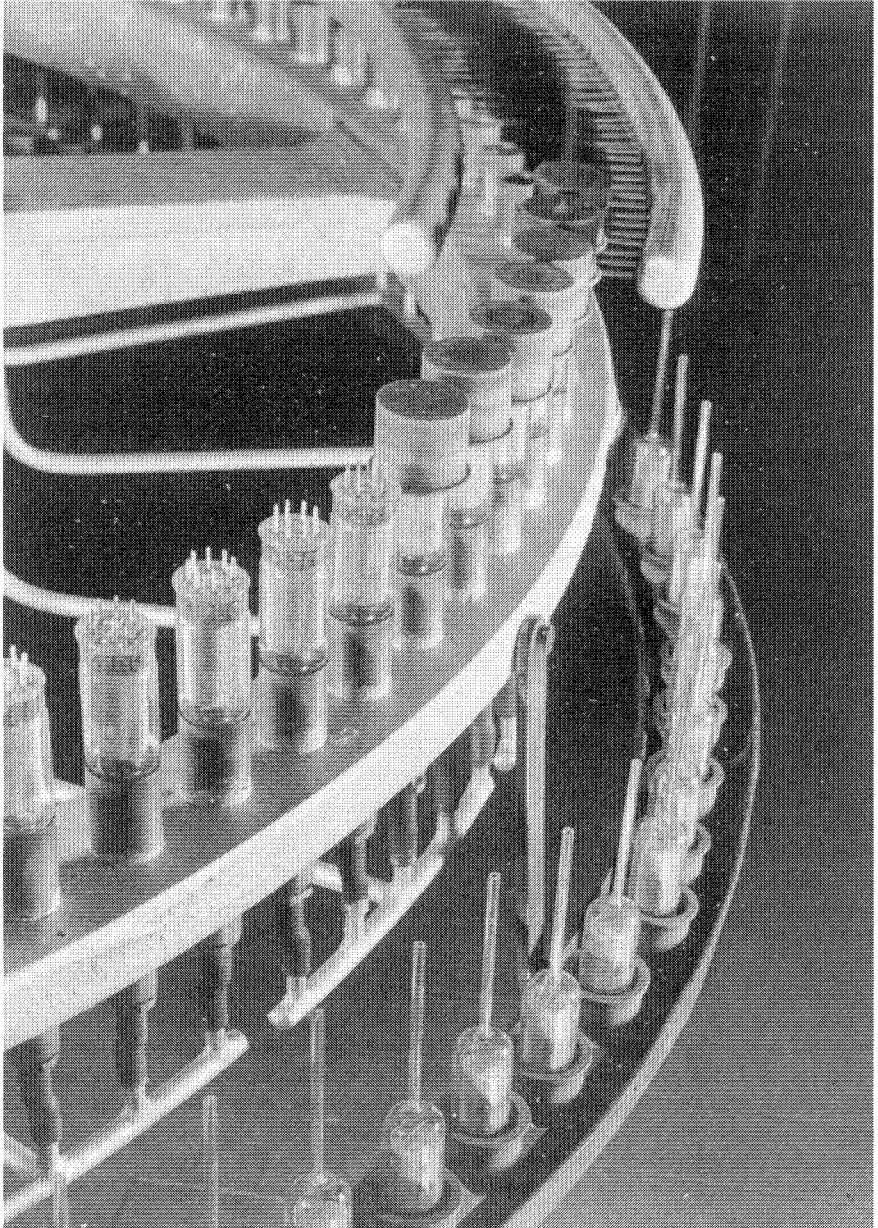


Fig. 1-18. Automatic sealing of miniature amplifying tubes.

From equation (1.22)

$$W_g = I_g (0.9 V_i - V_g) \approx I_g \cdot \hat{v}_g, \quad (1.24)$$

where \hat{v}_g denotes the peak value of positive grid voltage (see fig. 1-15).

The cathode heating supply for tubes in industrial apparatus is usually provided by a transformer from the a.c. mains. Maximum and minimum values of heater voltage are usually published by tube manufacturers; in the interest of long useful tube life they should not be exceeded. In many cases these tolerances are so large that normal fluctuations of the mains voltage may be disregarded. Where this is not the case voltage stabilizing devices should be provided in order to reduce the fluctuations to a permissible level.

Sometimes a hand-operated adjustable resistor or variable-core transformer with voltmeter etc. will suffice for heater regulation, but in this case careful and continuous supervision is necessary in order to maintain the nominal values of heating power.

Fig. 1-18 is a picture of an automatic machine employed in the manufacture of amplifying tubes, and shows the automatic sealing of miniature tubes.

2. RECTIFYING TUBES

Rectifying tubes are diodes or two-electrode tubes, containing only a directly or indirectly heated cathode and one or more anodes. In industrial equipment such tubes are mostly used for changing the alternating voltage and current delivered by the mains into d.c. power. Rectifying tubes are either highly evacuated or gas-filled. Vacuum tubes are used for the rectification of comparatively small powers; they are found chiefly in single- or two-phase circuits.

In fig. 2-1 the conventional symbols for a two-anode, directly heated tube and a single-anode indirectly heated tube are shown. The gas filling is indicated by a spot beside the cathode *).



Fig. 2.1. Tube symbols (left: two-anode directly heated gas-filled rectifying tube, right: single-anode indirectly heated gas-filled tube).

The basic circuit of a single-phase rectifier using a single-anode tube is shown in fig. 2-2. The secondary of transformer Tr produces an alternating voltage V_{tr} which is applied to the rectifying tube in series with the reservoir capacitor C across which the rectified voltage appears. The current can pass only in one direction as the negative electrons can flow only from the cathode to the anode when the anode potential is positive. This is the case only during one half of each cycle of the alternating voltage (half-wave rectification); during these periods the capacitor is charged with the polarity shown in fig. 2-2. In the following half cycle no current can flow and the capacitor will be discharged partially during this period by the load resistance R (for the sake of simplicity an ohmic resistance is assumed). In the following positive half cycle the process is repeated.

*) Instead of marking the tube symbols by a spot, they are often hatched to indicate the gas filling.

In fig. 2-3 this voltage variation is shown as a function of time. Obviously the voltage across the capacitor is opposed to the positive half cycles of the voltage delivered by the transformer, so that charging current can flow only

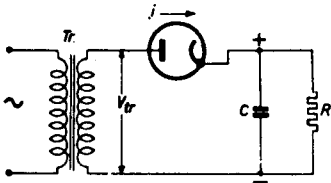


Fig. 2.2. Basic circuit of a single-phase half-wave rectifier.

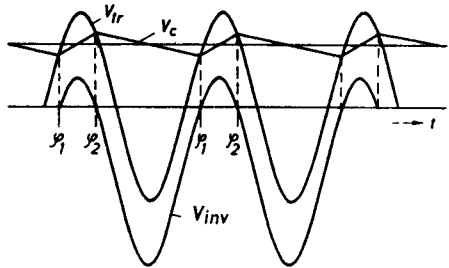


Fig. 2-3. Output voltage of a single-phase half-wave rectifier.

during the time $\varphi_2 - \varphi_1$ when the transformer voltage is predominant. During the negative half cycles, however, the voltage across the capacitor is added to the transformer voltage; in this period, therefore, a considerably higher "inverse voltage" is applied to the tube.

The rectified output voltage is not constant but has superimposed on it an alternating voltage having a frequency equal to that of the mains. This "ripple" voltage may be undesirable in the load circuit and should therefore be suppressed as far as possible. It is obvious that the ripple voltage will be smaller if the reservoir capacitor is chosen larger or if the load current is smaller. Moreover a considerable reduction of the ripple voltage can be obtained if a smoothing filter is added consisting of a choke (inductance) and a smoothing capacitor (fig. 2-4). Assuming a sinusoidal ripple voltage, the amount of smoothing, i.e. the ratio of the ripple voltage amplitudes following and preceding the filter will be

$$\frac{V_2}{V_1} = \frac{1}{\omega^2 LC_2 - 1} \tag{2.1}$$

Two single-anode rectifying tubes can be connected in such a way that the

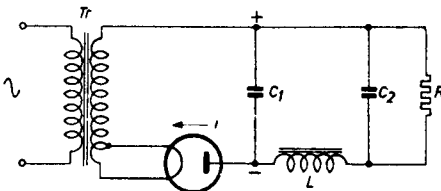


Fig. 2-4. Single-phase half-wave rectifier with smoothing filter.

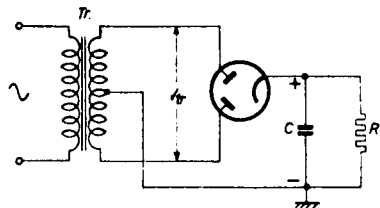


Fig. 2-5. Two-phase half-wave rectifier circuit.

reservoir capacitor will be charged by both half cycles. In that case it is often convenient to use tubes containing two anodes and a common cathode. Fig. 2-5 shows the basic circuit of a two-phase half-wave rectifier *) incorporating

*) This circuit is often called a "full-wave rectifier" in literature, but to avoid confusion, preference is given to the term "two-phase half-wave rectifier", the rectifier being fed by an alternating voltage having two opposite phases to which half-wave rectification is applied. The term "full-wave rectification" will be reserved for bridge circuits, which are discussed later.

such a two-anode tube; the single-phase mains voltage is converted into two voltages which are in antiphase.

Fig. 2-6 indicates the voltage waveform across the capacitor in this circuit. Obviously the ripple voltage now has twice the frequency of the mains, and therefore, as can be shown from equation (2.1), a given filter will give four times

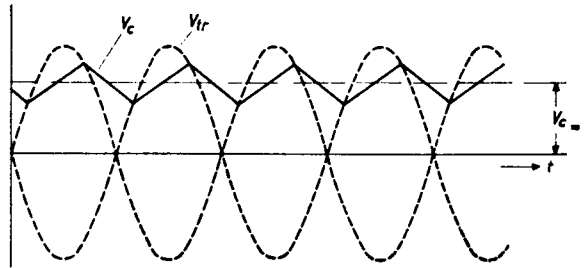


Fig. 2-6. Output voltage of a two-phase half-wave rectifier.

the smoothing in a two-phase circuit compared with a single-phase circuit. Moreover, the ripple voltage percentage is smaller than in a single-phase rectifier because the discharge times of the capacitor are much shorter. Also the peak current drawn from the cathode is smaller because the current flow is distributed over the whole cycle. For these reasons two-phase rectifying circuits are generally used when high vacuum tubes are employed. A commercial type of two-anode rectifying tube for two-phase rectification is illustrated in fig. 2-7.

In contrast to high vacuum rectifiers, there are the gas-filled rectifying tubes

which are widely used in industrial electronics. Like vacuum tubes, they have a directly or indirectly heated oxide-coated cathode and one or more anodes, but the envelope is filled with one of the inert gases argon, xenon or helium at low pressure, or with mercury vapour or with a mixture of both.

If a positive potential of only a few volts is applied to the anode of such a tube only a very small current will flow. If the anode voltage is increased beyond a certain value termed the "ignition voltage", V_{ign} , the tube suddenly fires and the anode current rises to a value which may be considerably higher than that in a vacuum tube of comparable size. This effect is caused by the ionisation of the gas atoms; at the ignition voltage the electrons achieve such a high speed that, when colliding with the gas atoms, they eject other electrons from the atoms, leaving the gas in the form of positive ions. The newly produced electrons are attracted to the anode, thereby increasing the anode current, and the positive ions move towards the cathode, where they neutralize the "space charge" or cloud of electrons which exists in the region of the cathode and limits

the number of electrons which can leave the cathode. Due to this neutralizing action the anode current is further increased.

When this condition is reached the value of the anode current is determined almost entirely by the values of the anode voltage and the external resistance in the anode circuit. The voltage drop across the tube itself, i.e. the so-called

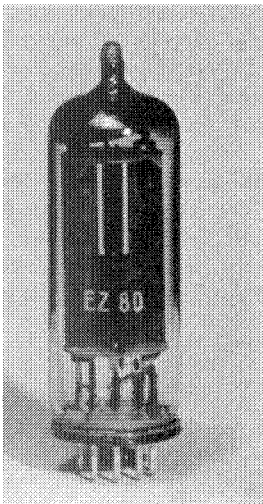


Fig. 2-7. Two-anode rectifying tube type EZ 80.

“arc voltage”, V_{arc} , arises mainly in the vicinity of the cathode, and is practically independent of the value of the anode current. Its value depends upon various factors, but mainly on the gas used for the filling and its pressure. The gas pressure in the case of a mercury-vapour tube is greatly dependent upon the temperature, so in these tubes the arc voltage is influenced by the working temperature. The arc voltage ranges from 8 to 32 volts according to the type of gas-filling; mercury-vapour filled tubes have an arc voltage of approximately 16 volts. It will thus be seen that the losses generated in the tube, i.e. the product of arc voltage and average anode current are comparatively small, and the higher the anode voltage the lower will be the percentage loss.

If the effective anode voltage becomes smaller than the arc voltage, the discharge is extinguished and the flow of anode current ceases. This occurs, for example, if the external anode circuit is interrupted.

Rectifying circuits using gas-filled tubes do not greatly differ from those for high-vacuum tubes except that the reservoir capacitor should be omitted. This is necessary because owing to the very small internal resistance of the gas-filled tube, such high peak currents would flow while the capacitor is charged that the cathode would be seriously damaged. The smoothing filter circuit will therefore commence with a choke.

In the remainder of this chapter some factors which are of importance in the design of rectifiers employing gas-filled tubes will be dealt with. The simplest circuit is the single-phase, half-wave rectifier shown in fig. 2-8, from which the multi-phase and full-wave rectifying circuits illustrated in fig. 2-16 can be comparatively easily derived.

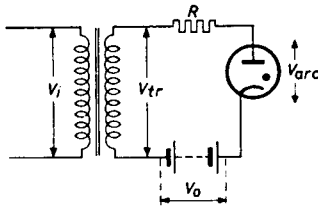


Fig. 2-8. Simple single-phase half-wave rectifier circuit with gas-filled tube and counter e.m.f.

It can be imagined that a counter-electromotive force V_o is inserted in the circuit, and is independent of the value of the current flowing (this is the case for instance when batteries are charged by the rectifier). The load is symbolized by an ohmic resistance R which also includes the resistance of the transformer winding. If V_{ign} is the ignition voltage, it is obvious that ignition will occur only if

$$\sqrt{2} V_{tr} > V_{ign} + V_o,$$

i.e. if

$$\frac{\sqrt{2} V_{tr}}{V_{ign} + V_o} = k > 1. \tag{2.2}$$

It is reasonable to choose V_{tr} so that the value of k is between 1.15 and 1.2 to ensure that ignition of the tube will occur with certainty even where the mains voltage is subject to fluctuation. The tube having ignited, the voltage across the tube decreases to the arc voltage V_{arc} . The instantaneous value of the current flowing then is

$$i = \frac{\sqrt{2} V_{tr} \cdot \sin \varphi - V_{arc} - V_o}{R}. \tag{2.3}$$

Making use of the concepts of firing angle φ_1 and extinguishing angle φ_2 of the tube (see fig. 2-9), the mean anode current per phase I_o is obtained by integrating the current from φ_1 to φ_2 and dividing by the whole cycle 2π :

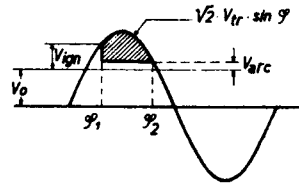


Fig. 2-9. Voltage curve of a rectifier according to fig. 2-8.

$$I_o = \frac{1}{2\pi} \cdot \int_{\varphi_1}^{\varphi_2} i \cdot d\varphi = \frac{1}{2\pi R} \cdot \int_{\varphi_1}^{\varphi_2} (\sqrt{2} V_{tr} \cdot \sin \varphi - V_{arc} - V_o) \cdot d\varphi. \quad (2.4)$$

If a denotes the voltage ratio $(V_{arc} + V_o) / \sqrt{2} V_{tr}$ (a obviously always being less than unity), equation (2.4) becomes:

$$I_o = \frac{\sqrt{2} V_{tr}}{2\pi R} \cdot \int_{\varphi_1}^{\varphi_2} (\sin \varphi - a) \cdot d\varphi. \quad (2.5)$$

The following approximations are admissible (see fig. 2-9):

$$\left. \begin{aligned} \varphi_1 &= \sin^{-1} a, \\ \varphi_2 &= \pi - \sin^{-1} a. \end{aligned} \right\} \quad (2.6)$$

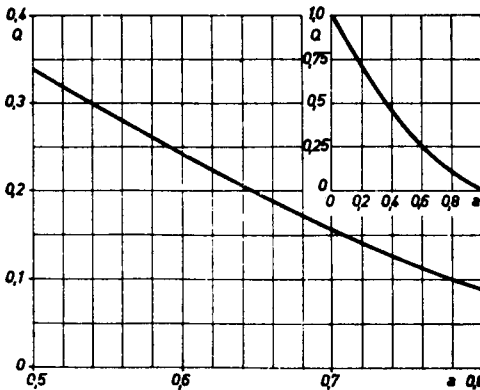
If the integration is carried out, it will be found that the mean current per phase is:

$$I_o = \frac{\sqrt{2} V_{tr}}{\pi R} (\sqrt{1 - a^2} - a \cdot \cos^{-1} a), \quad (2.7)$$

or

$$I_o = \frac{\sqrt{2} V_{tr}}{\pi R} \cdot Q. \quad (2.8)$$

In fig. 2-10 Q is plotted as a function of a . The maximum permissible mean current of the tube is specified in the published data; it can be shown from equ. (2.8) that the total resistance of the rectifying circuit must be



$$R \geq \frac{\sqrt{2} V_{tr}}{\pi I_c} \cdot Q. \quad (2.9)$$

When designing a rectifier, therefore, resistors R_a should be inserted in the anode leads in order to limit the current. The values of these resistors can be determined from equation (2.9) after deducting the transformer and load resistances.

Fig. 2-10. Graph showing Q as a function of a .

The peak value of the current

$$i = \frac{\sqrt{2} V_{tr} (\sin \varphi - a)}{R}$$

obviously occurs at $\varphi = 90^\circ$, and is

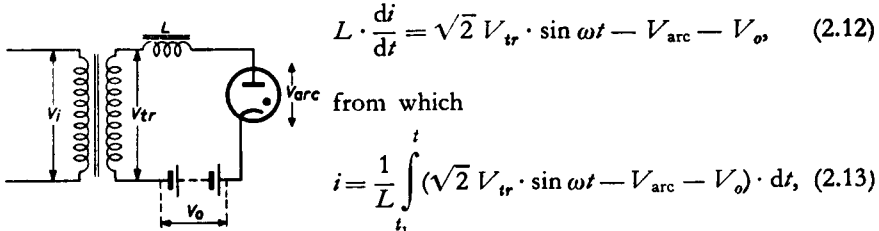
$$\hat{i} = \frac{\sqrt{2} V_{tr} (1 - a)}{R}, \tag{2.10}$$

so that

$$\frac{\hat{i}}{I_o} = \frac{(1 - a) \pi}{Q}. \tag{2.11}$$

When using a rectifier tube of a given type it is necessary to check that the maximum permissible peak anode current as calculated from equations (2.10) and (2.11) is not exceeded under working conditions.

In many cases the load of a rectifier is inductive. If the ohmic resistance of the rectifier circuit is then neglected the basic circuit will be as fig. 2-11 and the following equation applies:



$$L \cdot \frac{di}{dt} = \sqrt{2} V_{tr} \cdot \sin \omega t - V_{arc} - V_o, \tag{2.12}$$

from which

$$i = \frac{1}{L} \int_{t_1}^t (\sqrt{2} V_{tr} \cdot \sin \omega t - V_{arc} - V_o) \cdot dt, \tag{2.13}$$

or

$$i = \frac{1}{\omega L} \int_{\omega t_1}^{\omega t} (\sqrt{2} V_{tr} \cdot \sin \omega t - V_{arc} - V_o) \cdot d\omega t. \tag{2.14}$$

From equation (2.12) it follows that the maximum of current, which occurs at $di/dt = 0$, is reached at a point at which:

$$\sqrt{2} V_{tr} \cdot \sin \omega t = V_{arc} + V_o.$$

This means that there is a phase difference between voltage and current as is shown in fig. 2-12. Moreover it follows from equation (2.14) that $i = 0$ if the integral becomes zero, i.e. if the two shaded areas above and below the horizontal line $V_{arc} + V_o$ in fig. 2-12 are equal. Curve *b* shows the voltage V_L appearing across the inductance L , and curve *c* shows the anode current i .

It is to be seen that the current is maintained beyond the instant at which the transformer voltage V_{tr} has passed the zero line and has become negative. This may be strange at first sight because, as has already been stated, a gas-filled tube ceases to conduct when the anode voltage falls below the value of the arc voltage. However, the inductance L produces a voltage tending to maintain the current flow, and this voltage is sufficient to keep the anode of the tube positive with respect to the cathode by an amount equal to the arc

voltage. Thus the current continues to flow until the energy stored in the inductance has discharged, when the voltage collapses and the current ceases. This, as already stated, occurs when the shaded areas in fig. 2-12 are equal. However, if an additional ohmic resistance R is present in the circuit (this will always be so in practice) the current flow ceases somewhat earlier. The difference in the two shaded areas is then determined by the voltage drop across the resistance R caused by the current flow.

Equation (2.14) can be written

$$i = \frac{\sqrt{2} V_{tr}}{\omega L} \cdot \int_{\omega t_1}^{\omega t} (\sin \omega t - a) \cdot d\omega t, \quad (2.15)$$

whence

$$i = \frac{\sqrt{2} V_{tr}}{\omega L} \left\{ \cos \omega t_1 - \cos \omega t - a(\omega t - \omega t_1) \right\}. \quad (2.16)$$

Considering $\omega t_1 = \sin^{-1} a$, and introducing the conducting time

$$\tau = \omega t - \omega t_1:$$

$$i = \frac{\sqrt{2} V_{tr}}{\omega L} \left\{ \sqrt{1 - a^2} (1 - \cos \tau) + a \cdot \sin \tau - a\tau \right\}. \quad (2.17)$$

It has already been stated that the conducting time of the tube is considerably extended by the inductance. This can easily be shown mathematically if the

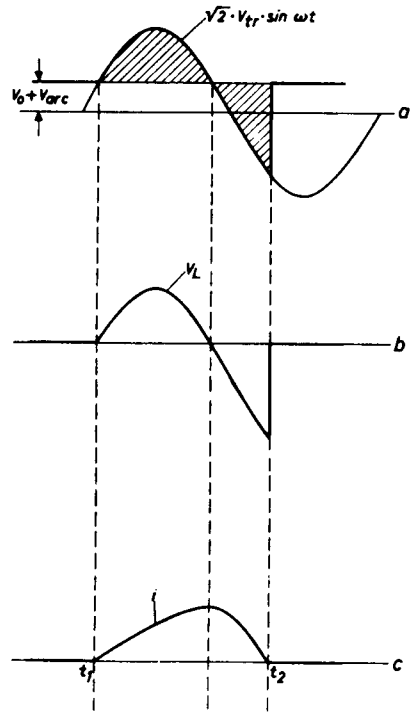


Fig. 2-12. Voltage and current curves of a rectifier according to fig. 2-11.

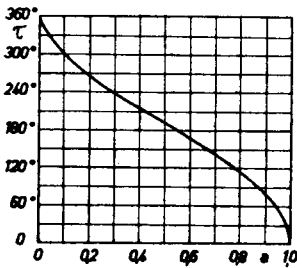


Fig. 2-13. Conducting time τ as a function of a .

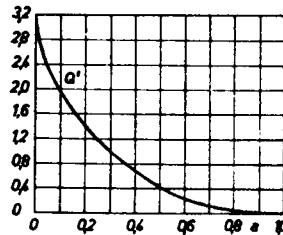


Fig. 2-14. Graph showing Q' as a function of a .

angle at which the current reaches its peak value is determined. This can be done by putting equation (2.12) equal to zero:

$$\omega t_{(i=\max)} = \pi - \sin^{-1}a. \tag{2.18}$$

The current reaches its peak value at a point where it would have already become zero if the load had been purely resistive. The actual conducting time τ can be calculated by putting equation (2.17) equal to zero; it is a function of a (see fig. 2-13).

The peak value of the current, i , can be determined easily from equation (2.16) and is:

$$i = \frac{2\sqrt{2} V_{tr}}{\omega L} (\sqrt{1 - a^2} - a \cdot \cos^{-1}a) = \frac{2\sqrt{2} V_{tr}}{\omega L} \cdot Q. \tag{2.19}$$

Furthermore, the mean anode current I_o can be calculated from equation (2.14); i.e.:

$$I_o = \frac{\sqrt{2} V_{tr}}{\pi\omega L} \cdot Q', \tag{2.20}$$

Q' being another function of a which is plotted in fig. 2-14.

A matter of great importance is the average rectified direct voltage V_o which can be obtained from an m -phase rectifier if the r.m.s. transformer voltage per phase is V_{tr} . Because the m phases are equivalent, each phase contributes to the output voltage during the time interval $2\pi/m$, as is shown in fig. 2-15. On this assumption it may be written:

$$V_o = \frac{m}{2\pi} \cdot \int_{\frac{\pi}{2} - \frac{\pi}{m}}^{\frac{\pi}{2} + \frac{\pi}{m}} \sqrt{2} V_{tr} \cdot \sin \varphi \cdot d\varphi; \tag{2.21}$$

integrating across the interval $2\pi/m$ gives:

$$V_o = \sqrt{2} \cdot \frac{m}{\pi} \cdot \sin \frac{\pi}{m} \cdot V_{tr} = M V_{tr}. \tag{2.22}$$

The term

$$M = \sqrt{2} \frac{m}{\pi} \cdot \sin \frac{\pi}{m} \tag{2.23}$$

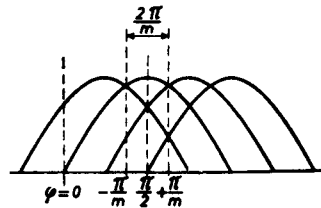
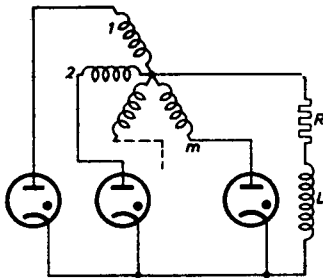


Fig. 2-15. Voltage curves of an m -phase rectifier.

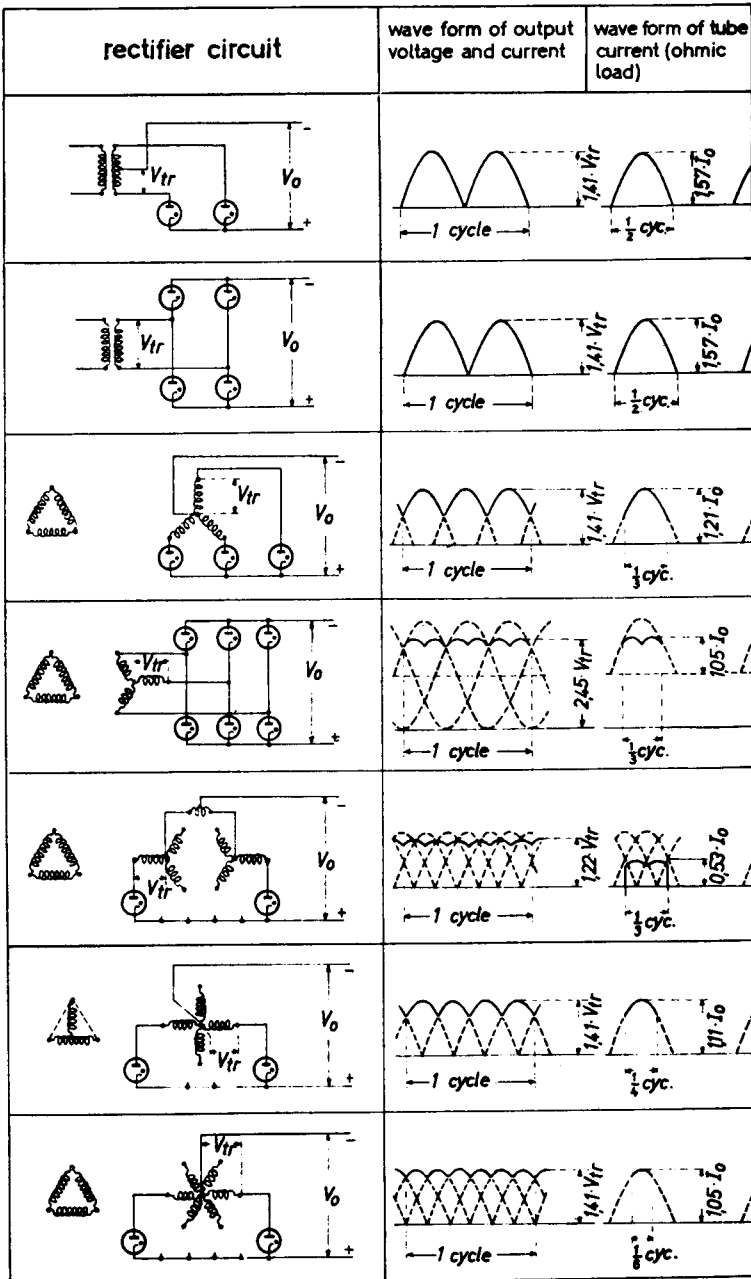


Fig. 2-16. Some typical rectifier circuits.

is called the phase factor. In the following table the values of M are indicated for several values of m :

m	2	3	4	6	12	18	...	∞
M	0.900	1.170	1.273	1.350	1.398	1.407		$\sqrt{2}$
$\frac{1}{M}$	1.111	0.855	0.786	0.741	0.715	0.711		0.707

In equation (2.22), however, the voltage losses arising in the transformer windings and across the rectifying tubes are neglected.

The most common rectifying circuits are shown in fig. 2-16. The following table gives values for the output voltages and currents in terms of the tube voltages and currents which may be used in approximate calculations.

V_{tr} is the transformer voltage per phase (r.m.s. value) and I_a the mean current per anode. The table indicates the theoretical values, transformer reactances being neglected; in practice the voltage losses in the transformer and across the tubes should be taken into account. Moreover, a certain voltage loss arises during the commutation period, the time required for the current to be taken over from one anode to the next. Generally, the over-all voltage losses caused by these factors at full load are approximately 10 to 15 per cent of the output voltage.

Circuit No.	Rectifying circuit	Mean output voltage V_o	Mean output current I_o	Inverse voltage (peak) per tube V_{inv}
1	two-phase half-wave	$0.318 V_{inv}$ $0.900 V_{tr}$	$2 I_a$	$2.828 V_{tr}$ $3.142 V_o$
2	two-phase full-wave (bridge)	$0.636 V_{inv}$ $0.900 V_{tr}$	$2 I_a$	$1.414 V_{tr}$ $1.571 V_o$
3	three-phase half-wave	$0.478 V_{inv}$ $1.170 V_{tr}$	$3 I_a$	$2.450 V_{tr}$ $2.094 V_o$
4	three-phase full-wave (bridge)	$0.956 V_{inv}$ $2.340 V_{tr}$	$3 I_o$	$2.450 V_{tr}$ $1.047 V_o$
5	three-phase half-wave (double Y)	$0.478 V_{inv}$ $1.170 V_{tr}$	$6 I_a$	$2.450 V_{tr}$ $2.094 V_o$
6	four-phase half-wave	$0.450 V_{inv}$ $1.273 V_{tr}$	$4 I_a$	$2.828 V_{tr}$ $2.221 V_o$
7	six-phase half-wave	$0.478 V_{inv}$ $1.350 V_{tr}$	$6 I_a$	$2.828 V_{tr}$ $2.094 V_o$

For even values of m the maximum value of the inverse voltage per tube can be calculated from the formula:

$$V_{inv} = 2\sqrt{2} V_{tr} = 2,828 V_{tr}. \quad (2.24)$$

For three-phase circuits:

$$V_{\text{inv}} = \sqrt{3} \cdot \sqrt{2} V_{tr} = 2,450 V_{tr}. \quad (2.25)$$

The inverse voltage as a function of the mean rectified output voltage V_o for even values of m is given according to equation (2.22) by:

$$V_{\text{inv}} = \frac{2}{\frac{m}{\pi} \cdot \sin \frac{\pi}{m}} \cdot V_o, \quad (2.26)$$

and for three-phase circuits by:

$$V_{\text{inv}} = \frac{\sqrt{3}}{\frac{3}{\pi} \cdot \sin \frac{\pi}{3}} \cdot V_o = \frac{2}{3}\pi V_o. \quad (2.27)$$

For full-wave (bridge) circuits the inverse voltage per tube is only half of these values, so that the values given by equations (2.26) and (2.27) should be divided by 2.

The life to be expected from gas-filled tubes is generally very high. Two circumstances may influence life: bombardment of the cathode by positive ions and the gas "clean-up". The components of the tube assembly tend to absorb gas atoms during life. For instance in rectifying circuits the anode attains a high negative potential during the inverse period, so that gas ions may reach the anode with a high speed and partially penetrate it. Due to the slowly decreasing gas pressure in the tube the electrical data ultimately change so greatly that the tube must be replaced. However, gas clean-up is of importance only in inert gas-filled tubes because mercury-vapour filled tubes contain excess mercury in liquid form which maintains the pressure during the life of the tube. This is one of the reasons why mercury-vapour filled tubes are usually preferred for use in industrial equipment.

Bombardment of the cathode by positive ions occurs when the tube is conducting and increases as the anode current and thus the gas ionisation increases. As the cathode can be ultimately destroyed by continuous heavy ion bombardment, the maximum per-

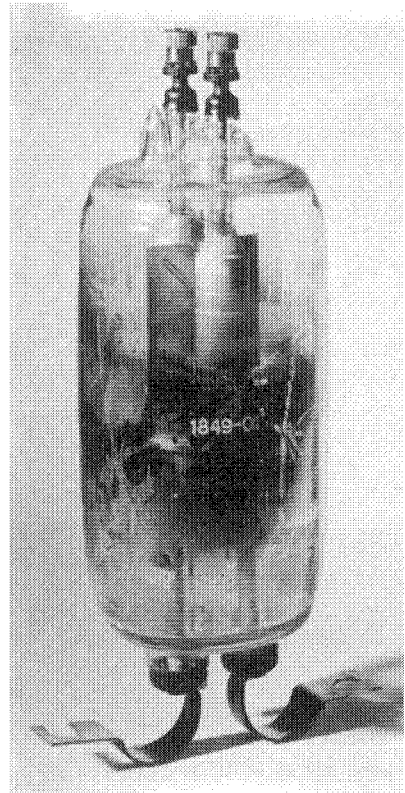


Fig. 2-17. Long-life rectifying tube, Philips Type 1849.

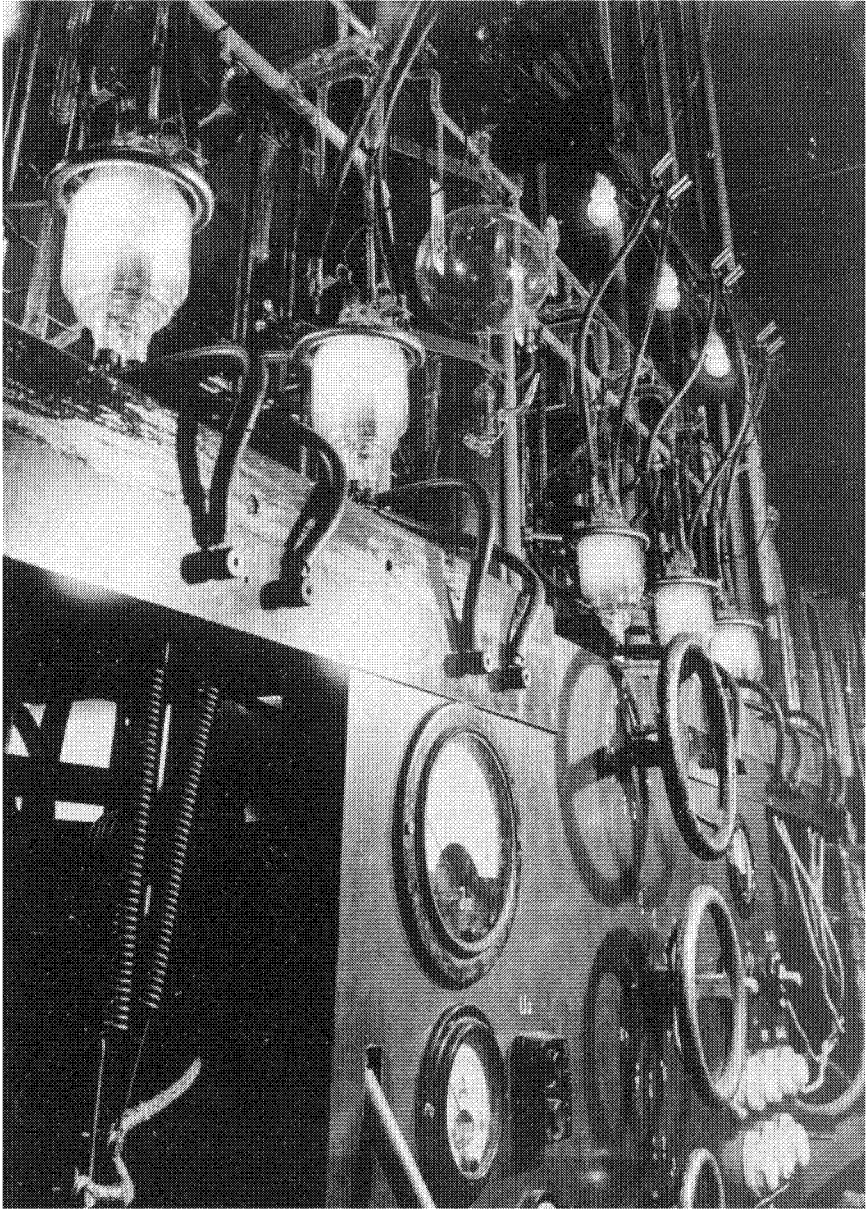


Fig. 2-18. Gas-filled rectifying tubes in course of manufacture.

missible value of the mean anode current and the peak anode current specified by the tube manufacturers should never be exceeded. By special constructional measures inside the tube, particularly by good screening of the cathode, ionic bombardment under normal working conditions can be kept very low. For instance, the two-anode rectifying tube shown in fig. 2-17, which is designed for a mean anode current of 25 amperes, attains a useful life of 20 000 to 30 000 hours.

If mercury-vapour filled tubes have been in stock for a considerable time, or have been exposed to heavy mechanical shocks (for instance during transportation etc.), it may happen that some of the liquid mercury inside the tube has splashed on the anode or other parts of the electrode system. If the tubes were immediately put into service in this condition there would be a risk of arc-back, i.e. a current flow in reverse direction. To avoid this, it is recommended that the cathodes of the tubes should be run at working temperature for a short length of time without applying anode voltage before the tubes are put into operation, thus allowing all mercury to evaporate from the anode and condense in the lower, cooler part of the tube. Recommendations on these lines will usually be given by tube manufacturers; they should be observed carefully in order to ensure a long useful tube life.

Fig. 2-18 shows gas-filled rectifying tubes in course of manufacture.

3. THYRATRONS

A thyratron is a gas-filled discharge tube containing a heated cathode, one or more grids and an anode. The influence of the grid gives the thyratron characteristics very different from those of the gas-filled rectifying tubes described in the preceding chapter, and makes it one of the most important components of industrial electronic apparatus. If a sufficiently great negative voltage is applied to the grid of a thyratron, a positive voltage being applied via a resistor to the anode (fig. 3-1), practically no anode current flows and the tube remains cut off, even though the anode voltage may be several hundred volts or more. If the negative grid bias is now progressively reduced, a point is reached at which the tube fires instantaneously and a large anode current flows, the value of which is dependent on the anode resistance and the applied anode voltage. The thyratron then behaves just like a gas-filled rectifying tube without grid.

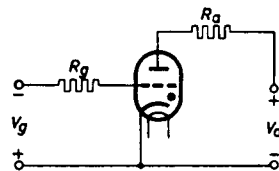


Fig. 3-1. Basic thyratron circuit.

For each value of positive anode voltage there is a corresponding value of negative grid voltage at which firing of the tube occurs. This connection between anode voltage and grid ignition voltage is usually plotted in a graph, termed the critical control characteristic of the tube. There are tubes with negative, positive or transition control characteristics, according to the values of grid voltage required for igniting.

Fig. 3-2 shows the three possibilities for mercury-vapour filled tubes. Under conditions corresponding to the region to the left side of the characteristics the tubes remain non-conducting; on the right-hand side ignition will occur.

It must be understood, however, that a tube, once ignited, cannot be cut

off by increasing the negative grid voltage as in the case of high vacuum tubes where the anode current may be continuously controlled by the negative grid voltage. In the thyratron a sheath of positive gas ions screens the grid so that it can no longer exercise control. The anode current therefore continues to

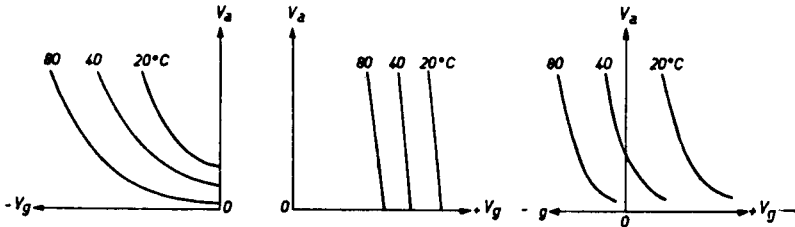


Fig. 3-2. Control characteristics of mercury-vapour filled thyratrons.

flow until the voltage between anode and cathode is reduced below the arc voltage value *).

If it is desired that the tube should remain non-conducting after the discharge has been interrupted, the anode voltage should not be allowed to reach its full value until after a certain short time interval has elapsed. This interval is the "de-ionisation time", i.e. the time needed for the grid to regain control of the tube; it is in the order of microseconds and commences at the moment when the anode current ceases. As the de-ionisation time depends on the intensity of ionisation, and this is again a function of the value of the anode current; it is specified for a given maximum mean anode current and a condensed mercury temperature of 40 °C. It will be shortened if the mean anode current is reduced, or the mercury temperature is decreased.

The maximum mean anode current is equivalent to the direct current which may continuously flow through the tube. As the current flow in thyratrons is usually discontinuous and often takes the form of impulses, the maximum mean anode current is stated as well as the maximum peak anode current in the published data. The latter is that value of anode current which may be permitted during a short period without overheating the anode or destroying the cathode by excessive ion bombardment. Usually two values are stated; one for an anode current with a frequency above 25 cycles per second, and the other for a frequency below 25 cycles per second. Because the duration of a current impulse is greater at lower frequencies, more heat is generated and also the cathode is bombarded by a greater number of ions. The maximum permissible peak anode current is therefore smaller.

The maximum mean anode current is a decisive factor in determining the load which may be applied in the anode circuit of a thyratron. In case of discontinuous load the mean current should be calculated from the averaging time stated in the published data. Assuming that the technical data of a certain

*) This is not correct in case of the so-called gas triodes which can be used for instance for generation of saw-tooth oscillations. These tubes can be rendered non-conducting by increasing the negative grid current, so that the ion cloud is absorbed by the grid. This can be achieved by increasing the negative grid bias, or by reducing the resistance inserted in the grid circuit.

thyatron tube specifies a maximum mean current of 3 amps, a peak anode current of 20 amps and an averaging time of 10 seconds, this means that during one averaging time a tube current of 20 amperes for 1,5 seconds, or a current of 10 amperes for 3 seconds, or 3 amperes for 10 seconds is permissible. The product $I_a \cdot T_{av} = 30 \text{ A} \cdot \text{sec}$ should therefore not be exceeded. Generally the following condition must be satisfied:

$$\int_0^{t+T_{av}} i \cdot dt \leq I_a \cdot T_{av} \quad (3.1)$$

with the additional condition that the anode current i never exceeds the maximum permissible peak value.

As stated before, tubes with different gas fillings behave in different ways. For example, the control characteristics of mercury-vapour filled thyratrons are dependent on temperature; with increasing temperature the characteristics are shifted toward the region of increasing negative grid voltage (fig. 3-2). The de-ionisation time is rather high (approx. 1000 μs), so these tubes can be used only in circuits working at low frequencies (up to approximately 500 c/s).

The tubes should be mounted vertically, socket downwards, in order that the liquid mercury shall flow to the bottom of the tube. The cathode should be switched on for several minutes before switching on the anode circuit in order to evaporate sufficient mercury.

The advantage of mercury-vapour filled tubes, however, is the reduced risk of arc-back, and long life. This type of thyatron is generally preferred in industrial equipment handling large powers.

If the thyatron is filled with inert gas (argon, helium, neon or xenon) the control characteristics are practically independent of temperature, as the gas pressure changes but little over the range of temperatures encountered in practice. As the life of inert gas thyratrons is considerably shortened by gas clean-up, the manufacturer endeavours to put a sufficient amount of gas into the tube. This may be done by increasing either the gas pressure or the dimensions of the bulb. For practical reasons the first method is chosen.

The maximum permissible inverse voltage decreases, however, as the gas pressure is increased, and for inert gas-filled thyratrons with normal dimensions and good life expectancy it is usually not greater than approximately 1300 V. On the other hand, mercury-vapour filled thyratrons can easily be made suitable for inverse voltages up to 30 kV. Inert gas-filled thyratrons have very short de-ionisation times (in the order of some microseconds), so they are suitable for circuits working at fairly high frequencies (up to 150 kc/s). These tubes may be operated in any position.

In addition to single-grid thyratrons (triodes) there are gas-filled tubes having two grids (tetrodes). The advantages of tetrodes are: low control-grid current, small capacitances between anode and control grid, and the possibility of altering the control characteristic by varying the screen-grid voltage.

The introduction of the additional screen grid considerably reduces the current flowing to the control grid when the tube is non-conducting. This may be as low as a few microamperes, which may be of importance in cases where a high resistance is inserted in the grid circuit. If the control grid current is large a considerable voltage drop will appear across this resistance, shifting

the ignition point and thus changing the operating conditions of the tube. A tetrode will therefore be preferred in those circumstances.

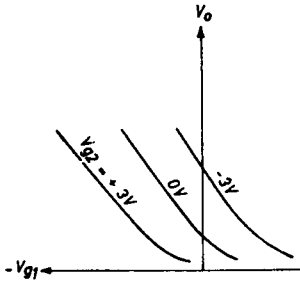


Fig. 3-3. Control characteristics of a tetrode thyatron.

Due to the small capacitance between anode and control grid, the grid circuit of a tetrode is almost independent of the anode circuit, so the risk of unintentional firing is small. For instance, in circuits where thyratrons are ignited by impulses applied to the control grid via a transformer, and the thyatron is a triode with a high grid-to-anode capacitance C_{ag} , it could happen that an impulse occurring unintentionally in the anode circuit is transferred via C_{ag} to the grid, thus firing the tube. In such a circuit a tetrode would give a better performance.

The third advantage of a tetrode is the possibility of shifting the control characteristic over a certain range. This can be done by adjusting the screen grid voltage V_{g2} , as is shown in fig. 3-3. It is of value in special circuits where it may be necessary to match two tubes having slightly different characteristics. Furthermore, small alterations of tube characteristics occurring during life can be compensated by screen-grid voltage adjustment.

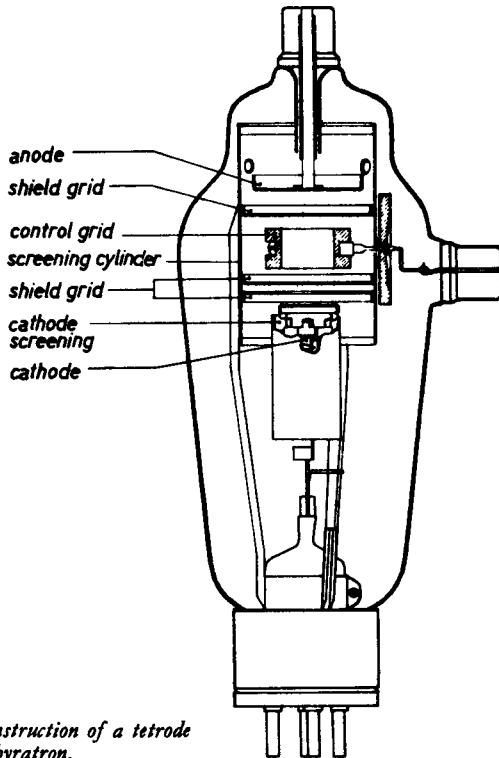


Fig. 3-4. Construction of a tetrode thyatron.

Finally, in special cases the screen grid can be used as an additional control electrode for igniting the tube.

In the published data of thyratrons the maximum permissible anode voltage is always stated. This value must not be exceeded during operation, otherwise the tube may ignite even though a high negative grid bias is applied. The inverse voltage limitations (anode negative with respect to the cathode) must be observed. If the inverse voltage exceeds the permitted value, current may flow through the tube in the reverse direction (arc-back). The occurrence of arc-back depends, among other factors, on the density of gas ionisation, and this again depends on the frequency of the anode current, so the maximum permissible inverse voltage is lower when operating at high frequency. Furthermore, because the inverse voltage is influenced by the gas pressure, the permissible inverse voltage of tubes with mercury vapour filling is to a certain extent dependent upon the temperature, being considerably reduced as the temperature rises.

Fig. 3-4 shows the internal construction of a tetrode thyratron. The bulb consists of thick glass to withstand the rough treatment which may be encountered in industrial service. The anode is made of metal or graphite and is specially shaped to provide good heat dissipation and thus to avoid excessive temperature rise and the consequent risk of secondary emission. The cathode is oxide coated, and is capable of giving high emission. Directly or indirectly heated cathodes may be used, their construction differing, however, from those used in high-

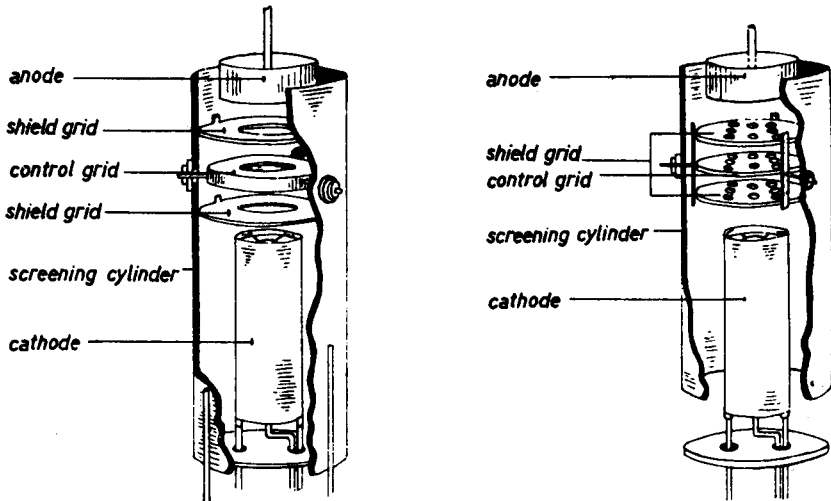


Fig. 3-5. Tube assemblies with negative (left) and positive (right) control characteristic.

vacuum tubes. As the electron paths in gas-filled tubes need not be straight lines, the cathodes may be for instance helical, thus increasing the thermal efficiency of the cathode by reducing radiation losses.

The warming-up time ranges from some seconds to several minutes according to the kind and size of the cathode. It should be noted that a longer warming-up period is required in the case of mercury-vapour filled tubes, to ensure that a

sufficiently high gas pressure has been reached before load is applied to the anode circuit of the tube.

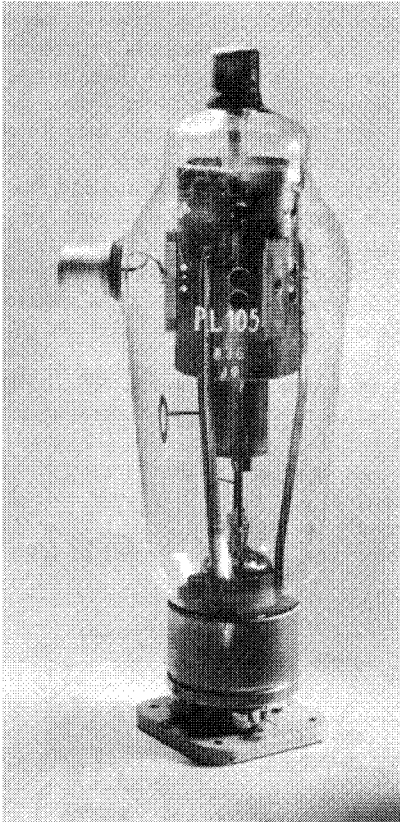


Fig. 3-6. Philips PL 105 tetrode thyatron.

The control grid generally takes the form of a ring or screen and consists of graphite or metal. Thyratrons with negative control characteristics have an annular control grid with a rather wide aperture. Tubes with positive control characteristics usually contain a system of perforated metal screens (see fig. 3-5). It is obvious that on account of the screening thus provided, firing occurs only at positive values of the control-grid voltage.

In tetrode thyratrons, the additional grid, screening the control grid from anode and cathode, also consists of perforated metal screens. The construction of tetrode assemblies is illustrated in figures 3-4 and 3-5. Fig. 3-6 shows a tetrode thyatron (Philips PL 105) designed for an average anode current of 6.4 amperes.

In most practical applications, thyratrons will be used as rectifying elements and fed with an alternating voltage. The anode voltage is therefore sinusoidal, as is shown in fig. 3-7. Obviously the instant of firing can then be determined by the value of the negative bias applied to the control grid. In the example shown, the thyatron will ignite at $V_g = 0$ V (d) only with a small delay (d'), but this delay

will increase as the negative grid voltage is increased (c , b).

If for every value of the sinusoidal anode voltage the corresponding ignition value of the grid voltage is plotted as a function of time, the "igniting characteristic" of the thyatron is obtained (fig. 3-8, broken line). By means of this igniting characteristic the retarded firing of the tube as a consequence of

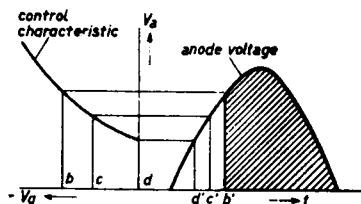


Fig. 3-7. Firing delay by negative grid bias.

varying the negative grid voltage can be well illustrated. The straight line corresponding to the grid voltage $V_{g(1)}$ intersects the igniting characteristic at the point A , so firing takes place and anode current will flow from this point to the end of the half cycle. If the grid voltage is increased to $V_{g(2)}$ the control characteristic is intersected at the point B , and it can be seen that the conducting time is then shorter. Since the hatched area is a measure for the power passed by the thyatron, it is obvious that, for instance, the output power of a rectifier equipped with such tubes can be controlled by varying the grid voltage.

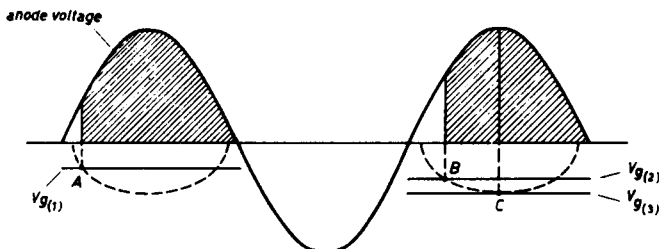


Fig. 3-8. Firing delay up to 90 degrees by negative direct grid voltages.

However, it can be seen from fig. 3-8 that control is possible only until point C is reached, i.e. up to a "firing angle" of 90 degrees, if only direct voltages are applied to the control grid. To extend the firing range up to 180 degrees, an alternating voltage is applied to the grid, the phase of this voltage with respect to that of the anode voltage depending on the desired firing angle (fig. 3-9).

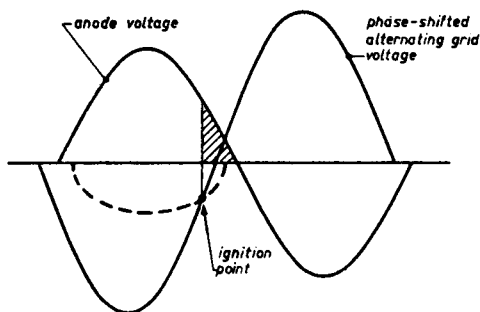


Fig. 3-9. Horizontal control by phase-shifted alternating grid voltage up to 180 degrees.

This method of controlling the ignition point is called horizontal control, because the alternating grid voltage is shifted horizontally on the time axis.

In many cases grid control by a sinusoidal voltage is not quite sufficient to obtain accurately timed firing on account of the inevitable tolerances on the igniting characteristic due, for instance, to its dependency on temperature etc. In these circumstances ignition can be initiated by impulses which are superimposed on a negative grid bias. The phase of these impulses is shifted with respect to the anode voltage by the firing angle α (fig. 3-10). The impulses are often generated by a special peaking transformer schematically shown in fig. 3-11.

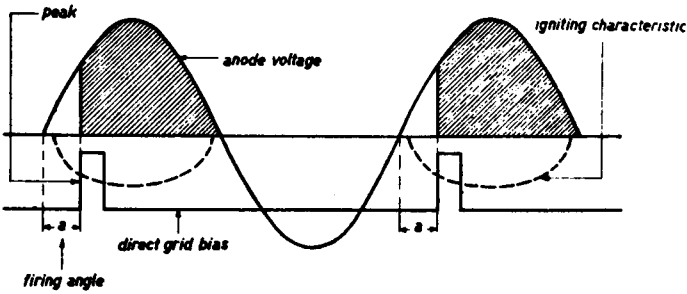


Fig. 3-10. Horizontal control by phase-shifted impulses.

This transformer has a primary winding p and one or more secondary windings s . The transformer core is of a material having low magnetic saturation and small cross section, so that it is saturated at low magnetic field strengths. Any further increase of the magnetic flux will be diverted to the by-pass and the air gap g . The magnetic flux B_s induced in the core of the secondary winding has the shape illustrated in fig. 3-11, and produces the peaked voltage V_s in the secondary winding.

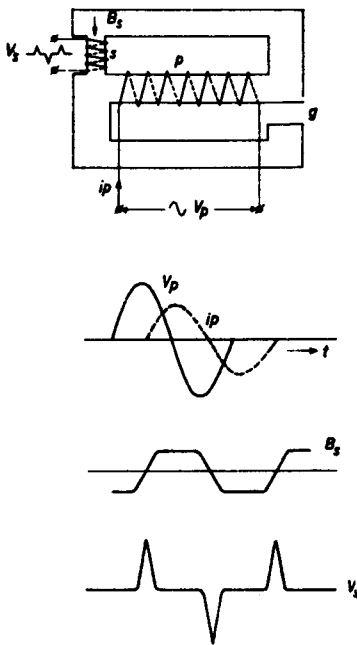


Fig. 3-11. Schematic representation of a peaking transformer.

Obviously the primary current lags with respect to the supply voltage by a certain angle, because of the inductance of the primary winding; similarly the secondary voltage peaks are shifted in phase by the same angle. However, it is often desired that the peaks appear at the beginning of the positive half cycles of the supply voltage. This can be ensured by inserting a resistor

R in series with the primary winding. If this resistor is large in comparison with the reactance of the winding, the voltage V_p appearing across the winding will lead by nearly 90 degrees with respect to the supply voltage V_a , as is shown by the vector diagram of fig. 3-12. Thus the angle of lag of the peaks can be compensated.

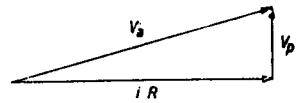


Fig. 3-12. Vector diagram showing phase conditions of peaking transformer primary when a series resistor R is inserted.

If this resistor is large in comparison with the reactance of the winding, the voltage V_p appearing across the winding will lead by nearly 90 degrees with respect to the supply voltage V_a , as is shown by the vector diagram of fig. 3-12. Thus the angle of lag of the peaks can be compensated.

Fig. 3-13 shows a typical peaking transformer together with a small rectifier

unit which serves as the voltage source for biasing the thyratrons.

Another method of controlling the firing angle is the so-called vertical control

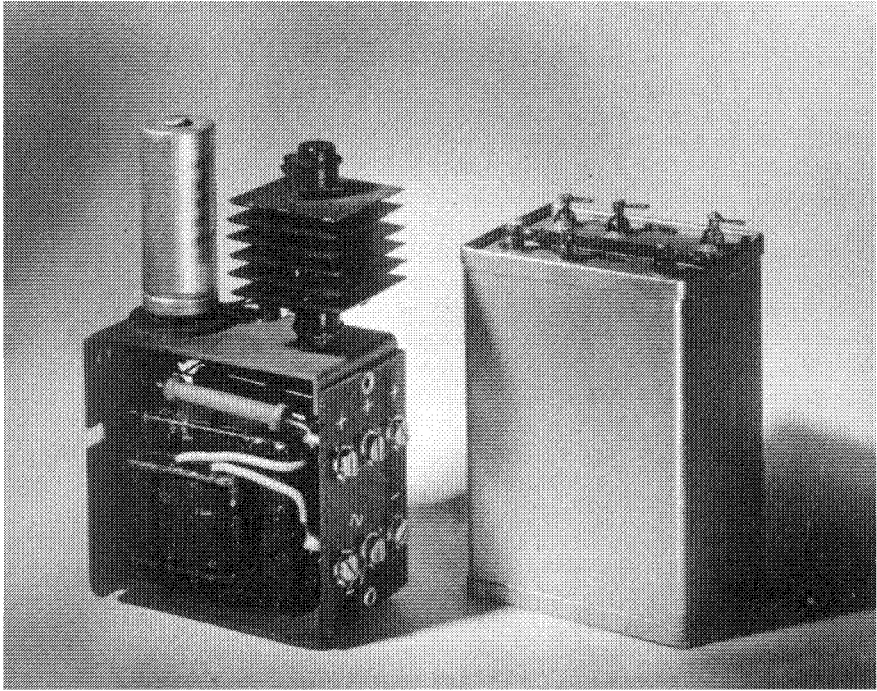


Fig. 3-13. Philips peaking transformer type 84590 (right) and rectifier unit type 1289 (left) for grid control of thyratrons.

which is often used in industrial apparatus because it requires few components (fig. 3-14). An alternating voltage lagging about 90 degrees behind the anode voltage is applied to the grid, superimposed on a direct voltage which can be varied between negative and positive values. As can be seen, the firing angle

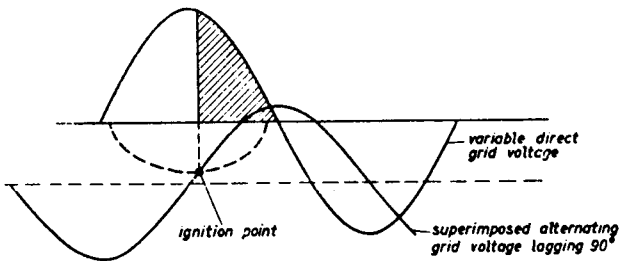


Fig. 3-14. Vertical control of thyratrons.

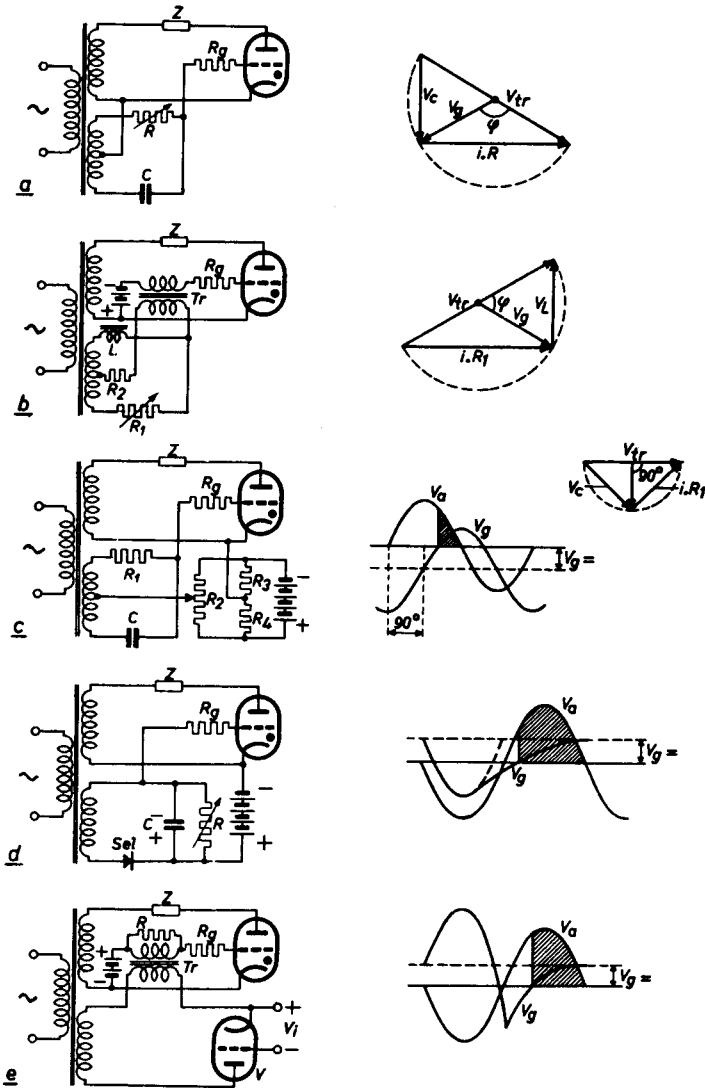


Fig. 3-15. Basic circuits for grid control of thyratrons.

can be varied between approximately 0 and 180 degrees, but it must be noted that the amplitude of the alternating grid voltage must not be too small, or it may not intersect the ignition characteristic, particularly at large firing angles.

A disadvantage of the vertical control is that the cathodes of simultaneously controlled thyratrons must have the same potential as the control is exercised by a direct voltage. This presents no difficulties in the case of multi-phase rectifiers, but if two tubes connected "back-to-back" are to be controlled, as is

the case in electronic welding contactors, light-dimming equipment and the like, vertical control cannot be used. In these cases horizontal control has to be adopted or one of the special circuits described later.

A number of typical basic circuits for retarded firing of thyratrons are reproduced in fig. 3-15. Circuit (a) shows a simple horizontal control by a phase-shifted alternating voltage. The anode circuit of the thyatron includes an impedance Z representing the load. The grid voltage is taken from a phase-shifting network consisting of the transformer secondary winding, a capacitor C and a variable resistor R . As shown by the corresponding vector diagram, the voltage across the capacitor lags about 90 degrees behind the voltage appearing across the ohmic resistance. As the sum of both voltages must always be equal to the transformer voltage V_{tr} , the top of the vector symbolizing the output voltage V_g moves on a semi-circle whose diameter is the vector of the transformer voltage V_{tr} . The magnitude of the output voltage V_g remains constant and is equal to the half of the transformer voltage, but its phase is shifted by an angle φ which may be varied over the range of approximately 0 to 180 degrees by altering the value of R . As will be shown later, if the load is inductive (which is generally the case), a much smaller range of firing angle is needed to control the tube current from zero up to its full value. The curves of the grid and anode voltages present in a circuit according to fig. 3-15a have been already shown in fig. 3-9.

Circuit (b) shows the horizontal control by phase-shifted peaks illustrated in fig. 3-10. Here again a phase-shifting network is employed. It consists of an ohmic resistor R_1 and an inductance L . The corresponding vector diagram is shown at the side and needs no further explanation. The output voltage of the phase-shifting network is applied to the primary winding of a peaking transformer Tr via a resistor R_2 which serves for phase correction. The grid circuit of the thyatron consists of the secondary winding producing the peaks and a d.c. source supplying the negative grid bias.

In circuit (c) is shown vertical control using a variable direct grid voltage and superimposed alternating voltage which lags by about 90 degrees. The latter is generated by a network consisting of C and R_1 , in which $R_1 = 1/\omega C$. The direct grid voltage, which may vary between positive and negative values, is taken from the slider of the potentiometer R_2 which forms a bridge circuit together with the equal resistors R_3 and R_4 . With this circuit also a firing angle range up to 180 degrees is obtainable. For the sake of simplicity in the illustration of the voltage curves the igniting characteristic of the thyatron is assumed to coincide with the zero axis.

In circuit (d) a device is shown which is neither a vertical nor a horizontal control. The tube is given a positive grid bias by a d.c. source. During the negative half cycles of the alternating anode voltage the capacitor C will be charged via a selenium rectifier Se' with the polarity indicated. In the following (positive) half cycle the capacitor will discharge through the variable resistor R . The form of the voltage present at the tube grid is shown at the side, and as is seen the voltage curve intersects the zero axis at a definite instant as determined by the value of the discharge resistor R , whereupon the tube becomes conductive.

Circuit (e) illustrates a grid control which may be used successfully when several thyratrons whose cathodes have not the same potential are to be controlled, for instance, in welding control devices. The amplifying tube V is

connected in such a way that it passes current during the negative half cycle of the anode voltage, so that during this period a current impulse flows through the primary winding of transformer *Tr*. As a result, a voltage is induced in the secondary winding and has the form shown at the side. It will be seen that the

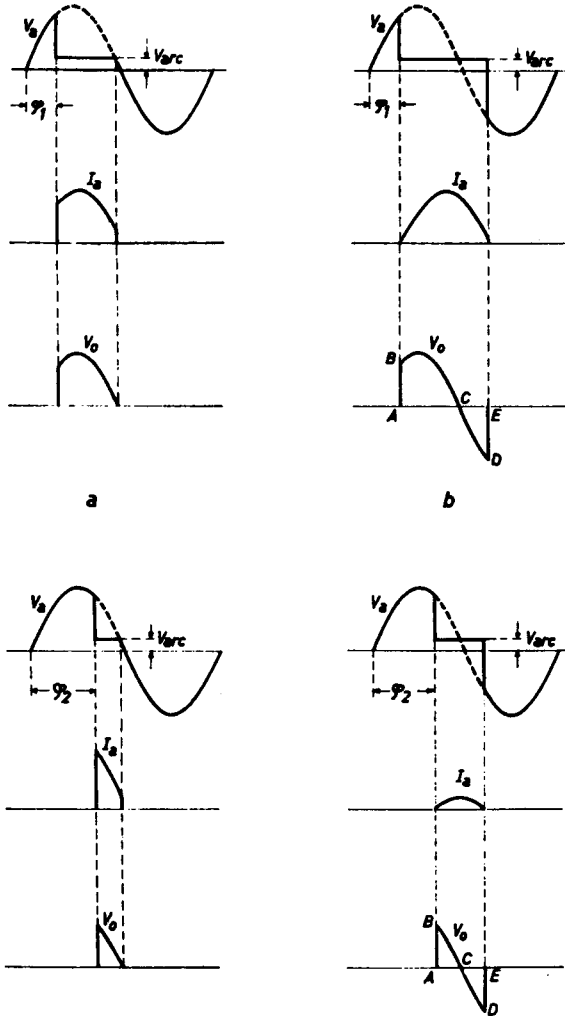


Fig. 3-16. Voltage and current curves at different firing angles (a) with resistive load, (b) with mixed resistive and inductive load.

thyatron is blocked at the beginning of the following positive half cycle of anode voltage. But the grid voltage again passes through the zero axis because of the positive bias applied to the grid circuit, and the thyatron fires. The firing angle may be varied by adjusting the negative control voltage V_c applied to the grid of the amplifying tube V . If the current through this tube is reduced

by increasing voltage V_i , a smaller current impulse flows through the primary winding of Tr , and the voltage appearing across the secondary winding swings negative for a shorter time, thus passing the zero axis at an earlier moment and reducing the firing angle of the thyatron.

All circuits shown in fig. 3-15 contain a protective resistor R_g which is inserted in the grid circuit of the thyatron. It serves to limit the grid current and will usually have a value between 10 k Ω and 50 k Ω .

The methods of grid control described above are widely used in electronic equipment. There are numerous variants, of course, but they are all derived from one or other of these basic circuits.

The way in which the anode voltage and current are affected by change of firing angle must now be considered. In fig. 3-16 the voltage and current curves are depicted for several values of the firing angle φ . At (a) pure ohmic load is assumed whilst (b) shows the curves when the load is inductive. In the first case current and voltage are in phase, so the current half wave will be "cut" if the firing angle increases. Moreover it is seen that the whole firing angle range from 180 to 0 degrees is needed for current control from zero up to its full value. In case of an inductive load the current wave lags with respect to the voltage curve, so the full current wave may pass the thyatron even if the firing point is retarded by a certain angle φ_1 . On the other hand the average current approaches zero for firing angles φ_2 which are smaller than 180 degrees, so it is evident that the firing angle range needed for full regulation is considerably smaller than in the case of a purely resistive load. When the load is entirely inductive the current flow ceases at the moment when the area CDE is equal to the area ABC . However, if there is a resistive component of load (which will always be so in practice) the current flow ceases earlier. The difference between ABC and CDE is then equal to the voltage drop resulting from the resistive load component.

As mentioned before, it is possible to construct rectifiers with thyratrons in which the output power may be easily adjusted by grid control in any way desired. The basic circuits used in this application are the same as described in the preceding chapter. Practically all that was previously said concerning the behaviour of gas-filled rectifying tubes is equally valid for thyratrons because as soon as they become conductive they behave just like gas-filled diodes. The various control circuits, however, will be described in detail in the second part of this book.

In fig. 3-17 the block diagram of a controlled three-phase rectifier is shown. Assuming firstly that the load is purely resistive, an output voltage is obtained which, plotted as a function of the firing angle φ_0 (now measured from the intersection of two successive anode voltage waves) is of the form depicted in fig. 3-18 for three different values of φ_0 . On the left of this figure the output voltage is shown for the case of unretarded firing. Each tube passes current during the time interval $\tau = 2\pi/m$, m being 3 in the case considered. In the

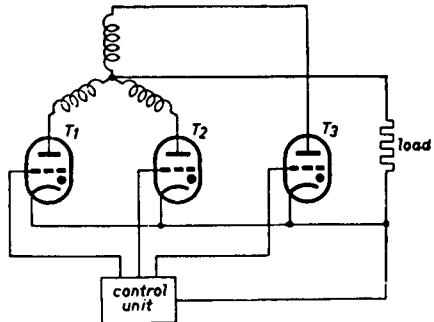


Fig. 3-17. Block diagram of a controlled three-phase rectifier.

centre of the diagram the voltage curve is shown for a firing angle $\varphi_0 = \bar{\varphi}$, where the output current is not yet interrupted, and the conducting time τ per

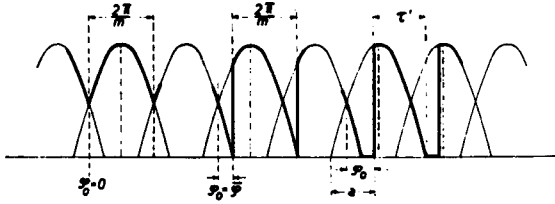


Fig. 3-18. Output voltage of a multi-phase rectifier at different firing angles (resistive load assumed).

anode is still not reduced. On the right the output voltage is plotted for a larger firing angle. As is seen, the conducting time τ' has become shorter, and the output current is no longer continuous, but is interrupted at regular intervals.

The general expression for the output voltage of an m -phase controlled rectifier with ohmic load, in terms of the firing angle φ_0 , is:

$$\begin{aligned}
 V_o' &= \frac{m}{2\pi} \cdot \int_{\frac{\pi}{2} - \frac{\pi}{m} + \varphi_0}^{\frac{\pi}{2} + \frac{\pi}{m} + \varphi_0} \sqrt{2} V_{tr} \cdot \sin \varphi \cdot d\varphi = \\
 &= \sqrt{2} \cdot \frac{m}{\pi} \cdot \sin \frac{\pi}{m} \cdot V_{tr} \cdot \cos \varphi_0 = M V_{tr} \cdot \cos \varphi_0. \quad (3.2)
 \end{aligned}$$

This voltage differs from the output voltage of an uncontrolled rectifier only in respect of the factor $\cos \varphi_0$. However, the equation (3.2) is valid only when the conducting time τ is not reduced, i.e. $\varphi_0 \leq \bar{\varphi}$, where $\bar{\varphi}$ obviously is given by

$$\bar{\varphi} = \frac{\pi}{2} - \frac{\pi}{m}. \quad (3.3)$$

If the firing angle is $\varphi_0 > \bar{\varphi}$, the reduced conducting time per anode is

$$\tau' = \frac{\pi}{2} + \frac{\pi}{m} - \varphi_0, \quad (3.4)$$

and the output voltage of an m -phase rectifier becomes:

$$\begin{aligned}
 V_o' &= \frac{m}{2\pi} \int_{\frac{\pi}{2} - \frac{\pi}{m} + \varphi_0}^{\pi} \sqrt{2} V_{tr} \cdot \sin \varphi \cdot d\varphi = \sqrt{2} \frac{m}{2\pi} \cdot \left[1 - \sin \left(\varphi_0 - \frac{\pi}{m} \right) \right] \cdot V_{tr}. \quad (3.5) \\
 &\frac{\pi}{2} - \frac{\pi}{m} + \varphi_0
 \end{aligned}$$

As can easily be seen from fig. 2-15, in a two-phase rectifier the anode conducting time will always be reduced if $\varphi_0 > 0$. This conducting time is $\tau' = \pi - \varphi_0$. The output voltage of a controlled two-phase rectifier is according to equation (3.5):

$$V_o' = \frac{\sqrt{2}}{\pi} \cdot (1 + \cos \varphi_0) \cdot V_{tr}. \quad (3.6)$$

If the firing angle φ_0 is not measured from the intersection of two successive anode voltage waves, but is referred to the point at which the anode wave crosses the zero axis (see fig. 3-10), then:

$$a = \varphi_0 + \frac{\pi}{2} - \frac{\pi}{m}. \quad (3.7)$$

as can be seen from fig. 3-18.

Things are quite different, however, if the load of the rectifier is mainly inductive. In this case the current from each anode continues until the following anode becomes conductive. This is shown in fig. 3-19 for different firing angles,

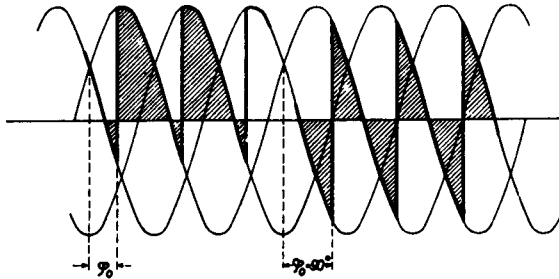


Fig. 3-19. Output voltage of a multi-phase rectifier at different firing angles (inductive load assumed).

$m = 3$ assumed. As the hatched areas below the zero axis must be subtracted from those above the zero axis, the output voltage is lower than in the case of a resistive load. For a firing angle $\varphi_0 = 90^\circ$ the resulting output voltage is zero for inductive load, although in the case of a resistive load an output voltage will still appear, and will be equal to:

$$V_o' = \frac{3\sqrt{2}}{2\pi} \cdot \int_{120^\circ}^{180^\circ} \sin \varphi \cdot d\varphi \cdot V_{tr} = 0,34 V_{tr}. \quad (3.8)$$

In the case of two-phase rectification and an inductive load the mean output voltage is:

$$V_o' = \frac{\sqrt{2} \cdot V_{tr}}{\pi} \int_{\varphi_c}^{\varphi_0 + \pi} \sin \varphi \, d\varphi, \quad (3.9)$$

since in this case each tube conducts during 180° . Integration gives:

$$V_o' = \frac{\sqrt{2} \cdot V_{tr}}{\pi} \cdot \left\{ \cos \varphi_0 - \cos (\varphi_0 + \pi) \right\} = \frac{2\sqrt{2}}{\pi} \cdot V_{tr} \cdot \cos \varphi_0. \quad (3.10)$$

It is thus seen again that for $\varphi_0 = 90^\circ$ the output voltage becomes zero, whereas in the case of a resistive load, an output voltage of

$$V_o' = \frac{\sqrt{2} \cdot V_{tr}}{\pi} = \frac{V_o}{2}. \quad (3.11)$$

will still be present according to equ. (3.6).

Parallel connection of thyratrons in order to obtain higher powers is usually not practicable unless special precautions are taken. The reason is that one tube will always fire first and, as the arc voltage is lower than the ignition voltage of the other tubes, the first tube to fire will take over the whole load. To avoid this, resistors R_a can be inserted into the anode leads (fig. 3-20). When the first

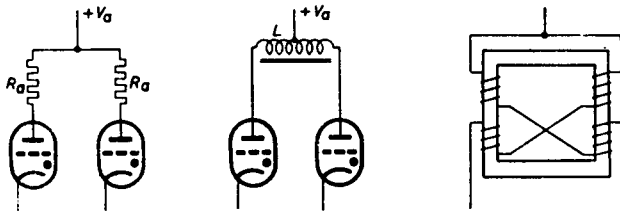


Fig. 3-20. Parallel connection of two thyratrons.

tube fires, a voltage drop appears across the resistor in its anode circuit; this voltage may be large enough to equal the difference between ignition voltage and arc voltage, thus permitting the second tube to fire also. A still better arrangement is the use of a balance choke. When the first tube fires, a voltage is induced in the other half of the winding by the sudden current rise, and this ensures immediate ignition of the second tube. Moreover, the power lost in the balance choke is much smaller than that in anode resistors. Fig. 3-20 shows schematically the construction of such a choke; the two halves of the winding are cross-connected for better performance; the iron core has no air gap.

The burning life of thyatron tubes is approximately the same as that of gas-filled rectifying tubes, as far as the effects of ion bombardment of the cathode and gradual reduction of gas pressure are concerned. The cathode is either of the shielded type or a special cathode is used in which the emitting oxide coating is located on the inner surfaces of a multiple subdivided hollow cylinder.

There is, however, a further risk which must be guarded against. If a portion of the emissive cathode coating should evaporate and be precipitated on the grid, the grid will emit electrons (grid emission). It will then lose its ability to control the tube. By special treatment of the grid surface and other constructional features, however, the danger of grid emission can be practically eliminated. Moreover this effect is less likely to occur in tetrodes, because in these tubes the control grid is well shielded from the cathode by the screen grid (see fig. 3-4).

As already mentioned, the control of thyratrons often is affected by phase-shifting networks containing both resistance and inductance or a capacitance. In fig. 3-21 is given an abac for determining the phase angle φ of such networks. A more exact calculation, however, can be made by applying the formula

$$\varphi = 2 \tan^{-1} (\omega RC) \quad (3.12)$$

for RC networks, and

$$\varphi = 2 \tan^{-1} \left(\frac{\omega L}{R} \right) \tag{3.13}$$

for *RL* networks.

Attention is drawn to the fact that in all circuits equipped with thyratrons provision must be made to ensure that the tubes are given sufficient time to heat up. The anode voltage may therefore not be applied before the filament current has been switched on for a certain time as specified in the data published by the tube manufacturer.

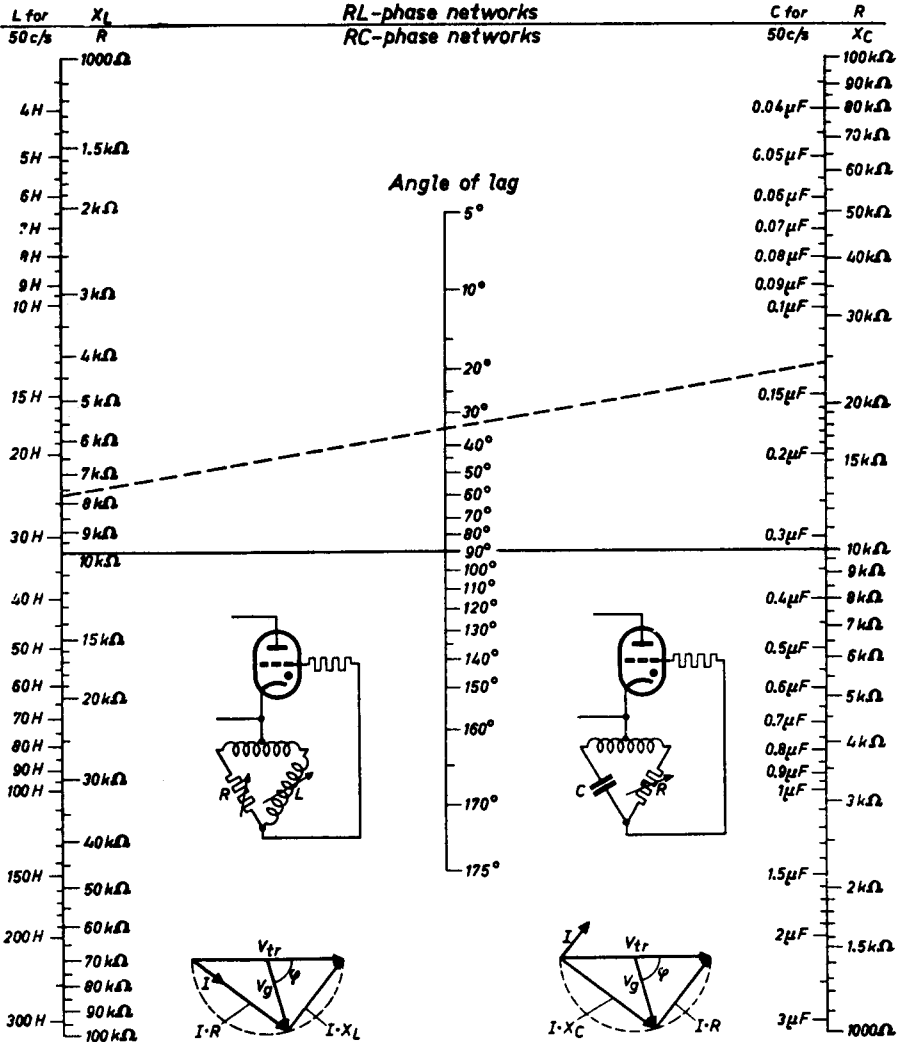


Fig. 3-21. Tab'e and abac for determining the phase angle of *RL* and *RC* networks.

4. SENDITRONS

Another type of tube often employed in industrial applications as a quick-acting electronic switch or relay is the senditron. It is able to carry instantaneously very heavy currents — up to several thousand amperes. Compared with mechanical contactors, contactor devices equipped with senditrons are much more reliable, more precise in operation, and often less costly. Unlike that of the thyatron, the cathode of the senditron consists of liquid mercury, as in the familiar mercury-pool rectifier valve. There is, however, an important difference between the method of starting the senditron and of starting the mercury-pool rectifier.

In starting the latter, the connection between the mercury pool and an igniting electrode has to be interrupted — usually by mechanical means, for instance by tilting the tube — when the arc so produced initiates the main discharge between cathode and anode. During operation a small hot spot is produced on the cathode, and moves at high speed over the surface of the mercury. Since this cathode spot disappears, and the discharge ceases immediately the discharge current falls below a certain value, one or more auxiliary anodes are usually provided in order to maintain the cathode current should the main discharge current become too small. Repetition of the rather complicated starting process, which would otherwise be necessary, is thus avoided. However, this device has the disadvantage that it involves additional current consumption by the auxiliary anodes. There is also the risk of arc-back as the result of permanent ionisation.

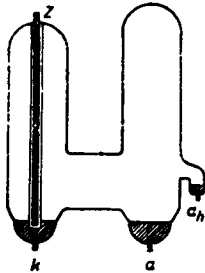


Fig. 4-1. Diagrammatic view of a senditron.

In the senditron, however, it is not necessary to maintain the cathode hot spot. The firing procedure is simplified in such a way that the tube can be fired practically instantaneously at an accurately determined time, and the firing cycle may be repeated at very small intervals of time up to a specified upper limit of frequency.

The senditron contains a capacitive igniter, the operation of which can be explained by reference to fig. 4-1. An igniting electrode Z extends downwards into the mercury pool k , but is insulated from the mercury by a thin layer of dielectric material. If a sufficiently high voltage is applied between the igniting electrode and the cathode, the electric field strength will be sufficient to produce a small amount of electron emission ("field emission" *).

A high voltage momentarily applied to an auxiliary electrode a_h accelerates these electrons, which attain sufficiently high speeds to ionize the mercury vapour atoms by collision, thus starting the main discharge. As in all gas-filled discharge tubes, the discharge will be interrupted when the anode voltage drops below the value of the arc voltage.

*) Until recently it was thought that the field strength between the igniting electrode and the mercury cathode was not sufficient to produce field emission. Later investigation, however, suggests that the field causes the mercury surface to deform; small peaks of mercury arise and cause local concentration of the field up to the intensities necessary for field emission. The theoretically calculated time needed for deformation agrees well with the small time delays, in the order of 10^{-8} seconds, observed by actual measurement.

Not only the cathode k , but also the anode a and auxiliary electrode a_a consist of liquid mercury. If other materials were used, the anode material would slowly erode during the life of the tube, and might contaminate the mercury surface of the cathode, causing irregular firing.

The advantage of senditrons over thyratrons with heated cathodes is that the mercury pool cathode is able to deliver currents up to many thousands of amperes. In some applications a high but only momentary current impulse is desired, the duration of which is small compared with the subsequent pause, so that the mean current, averaged over the complete cycle, is much smaller than the peak value. Since it is the product of the mean current and the arc voltage which constitutes the effective power dissipated in the tube, senditrons of small dimensions can switch very large current pulses of short duration.

A senditron, Philips type PL 5, of a type normally used in electronic equipment is illustrated in fig. 4-2.

The great advantage of the capacitive ignitor used in senditrons is the very small energy required for starting the main discharge (approximately 6 milliwatt-seconds). Senditrons are therefore particularly suitable for controlling small welding machines when the ignitron tubes, described in the next chapter, would be too expensive.

One of the most important applications of the senditron is in resistance welding, and particularly in spot-welding control systems. To ensure optimum quality of the weld, accurate control of welding time is necessary. This may be effected

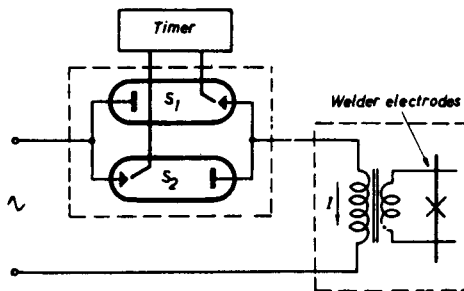


Fig. 4-3. Basic circuit of an electronic welding contactor using two inverse-parallel connected senditrons.

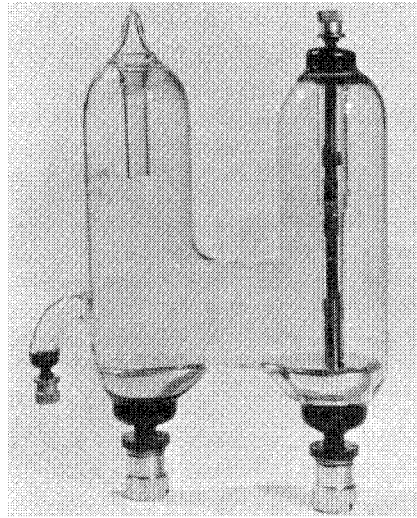


Fig. 4-2. Philips PL 5 senditron.

by the arrangement shown in fig. 4-3. Two senditrons S_1 and S_2 , connected in inverse-parallel (back-to-back) are included in the primary circuit of the welding transformer, and function as an inertialess contactor. One tube carries the current during the positive half-cycles and the other during the negative half-cycles (see fig. 4-4). Firing is controlled by a special timer, the operation of which is explained in Chapter 12.

A welding control device using two senditrons is particularly suitable for spot welding of thin material in mass production, for instance in the

toy industry and in the manufacture of precision apparatus, measuring instruments, watches, musical instruments and so forth.

A further application is the control of electronic flash-lamps for observing or recording the motion of high-speed rotating or moving objects such as propellers, motors wheels etc. The basic circuit of such a stroboscope is shown in fig. 4-5. Its action is as follows.

Assuming the flash-tube L to be extinguished, the capacitor C is charged

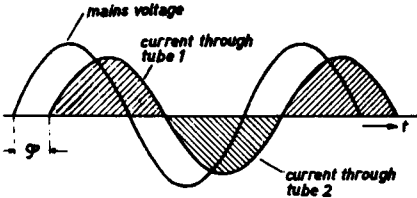


Fig. 4-4. Waveforms of currents flowing through two inverse-parallel connected senditrons.

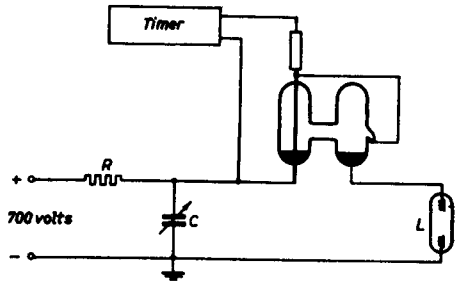


Fig. 4-5. Basic circuit of a stroboscope.

by the d.c. source via resistor R . To initiate the flash the senditron is made conductive, allowing the capacitor to discharge, so that a heavy current surge passes through the senditron and the flash-lamp L . By adjusting the value of the capacitor, the duration of the flash may be varied between 5 and 10 microseconds. The minimum interval between consecutive impulses will depend upon the time constant of the RC circuit, i.e. upon the values of R and C , the lower limit corresponding to the highest permissible switching frequency of the senditron, which in turn will depend upon the de-ionisation time of the gas filling.

The timer, once set, automatically controls the flash frequency by continuously switching the senditron. It will usually be synchronised at a standard frequency. The basic circuit of a timer suitable for stroboscope control is shown in fig. 4-6. It is a relaxation oscillator, the operating principle being as follows.

When the H.T. supply is switched on, the capacitor C charges up to the full

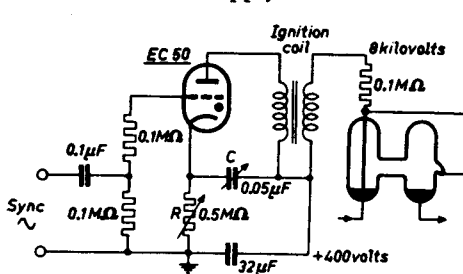


Fig. 4-6. Timer circuit suitable for the stroboscope shown in fig. 4-7.

potential of 400 volts. During the charging period the charging current causes a voltage drop across R , so that the grid of the thyatron EC 50 becomes negative with respect to the cathode. When C is almost fully charged the charging current becomes practically zero and the negative bias almost disappears, permitting the thyatron to fire. The thyatron in its conductive condition virtually short-circuits the capacitor C which discharges very

rapidly through the tube and the primary of the ignition coil, thus inducing a very steep-fronted voltage pulse at the transformer secondary.

the second half cycle, thus causing alternate firing of the tubes. When Tb_1 is ignited, capacitor C_3 discharges rapidly through Tb_1 and the primary of transformer Tr_2 , and the steep-fronted current surge through this winding produces a high voltage peak of about 8 to 10 kV at the secondary, thus igniting senditron S_1 .

Due to the inductance of the primary winding of Tr_2 , the discharge current will be maintained long enough for C_3 to charge momentarily with reverse polarity, so that for a short time a negative voltage is applied to the anode of Tb_1 , causing extinction of the tube.

In a corresponding manner, thyatron Tb_2 and senditron S_2 are fired in the following half-cycle, and the process will repeat as long as switch S is closed. On opening the switch the generation of impulses ceases and S_1, S_2 cease to fire. By varying R_1 the phase angle, and hence the mean anode current flowing through S_1, S_2 and the primary winding of the welding transformer can be adjusted (see fig. 4-8).

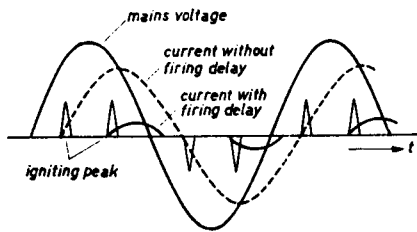


Fig. 4-8. Voltage and current curves with and without retarded firing.

The value of the alternating current flowing through two inverse-parallel connected senditrons can be derived from the average current per tube stated in the published data. Obviously the mean anode current, I_a , is equal to the integral over one half cycle of the

alternating current divided by its base π , thus:

$$I_a = \frac{1}{\pi} \cdot \int_0^{\pi} I_p \cdot \sin \varphi \cdot d\varphi = \frac{2}{\pi} \cdot I_p, \tag{4.1}$$

where I_p is the crest value of the alternating current. Since each tube carries current only during alternate half-cycles, twice the value of the mean anode current may be assumed, and the r.m.s. value of the resulting alternating current flowing through the inverse-parallel connection is:

$$I_{rms} = \frac{\pi}{\sqrt{2}} \cdot I_a. \tag{4.2}$$

For example, the alternating current through two type PL 5 senditrons, which may carry an average current of 70 amperes each during 0.05 sec, will be $70\pi/\sqrt{2} = 154 A_{rms}$ during this time.

5. IGNITRONS AND EXCITRONS

Ignitrons, which are often used in industrial equipment for controlling or rectifying very large alternating currents, are discharge tubes consisting essentially of a mercury pool cathode, an anode, and an ignition electrode which extends down into the mercury. A sectional drawing of an ignitron is reproduced in fig. 5-1. The envelope is of stainless steel and has double walls to permit the circulation of cooling water. The anode and cathode leads are of large cross-section since they have to carry very heavy currents. The mercury pool

cathode is at the same potential as the steel jacket, and the anode lead is insulated by a glass-metal seal.

The ignition electrode or ignitor, which is a rod of refractory material (boron carbide), and is shaped like the pointed end of a lead pencil, dips into the mercury pool but is not wetted by the mercury (see fig. 5-2). If a positive potential of about 150 volts is applied to the ignition electrode, a momentary current of approximately 10 to 40 amperes will flow, and will cause a tiny arc at the junction between the ignitor and the mercury. This arc immediately initiates the main

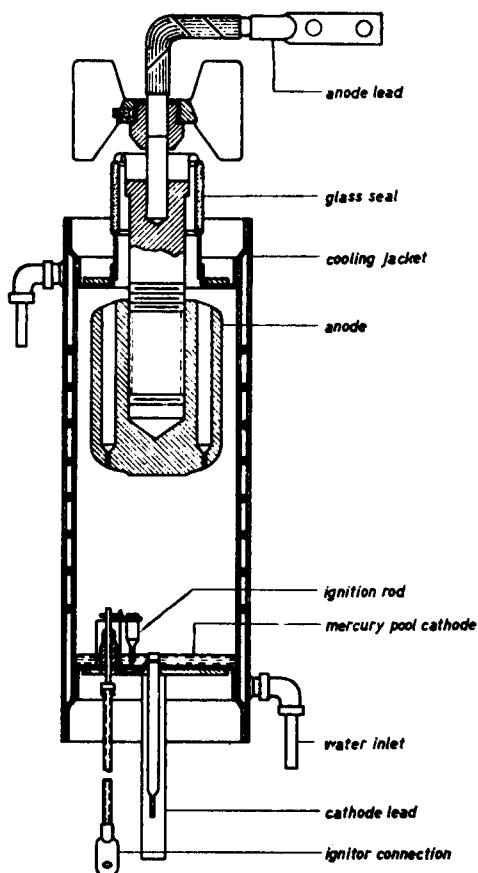


Fig. 5-1. Sectional diagram of an ignitron tube.

discharge between the mercury cathode and the positive anode, the familiar small cathode hot spot appearing on the surface of the mercury. Ignition takes place by virtue of the very strong electric field which arises at the junction of the ignitor rod and the mercury pool, thus drawing electrons from the mercury.

The main discharge ceases and the cathode spot disappears if the anode voltage falls below the "arc voltage" value or if the anode current falls below a certain critical value. The arc voltage value is in the order of 12 to 18 volts as in the case of mercury vapour thyratrons, so the tube losses are

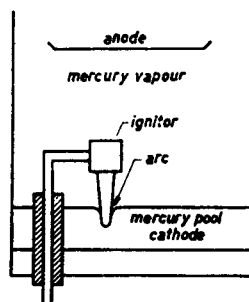


Fig. 5-2. Detail of the firing arrangement in an ignitron tube.

small compared with the output power, and the overall efficiency, even at low output voltages, is high. Two general types of ignitron tubes are available — those designed for high peak anode currents at comparatively low voltages in the order of 250 to 600 volts, and those designed to deliver high average currents at high anode voltages up to 2400 volts. The former are mainly intended for the control of welding equipment and the latter for the rectification of heavy currents.

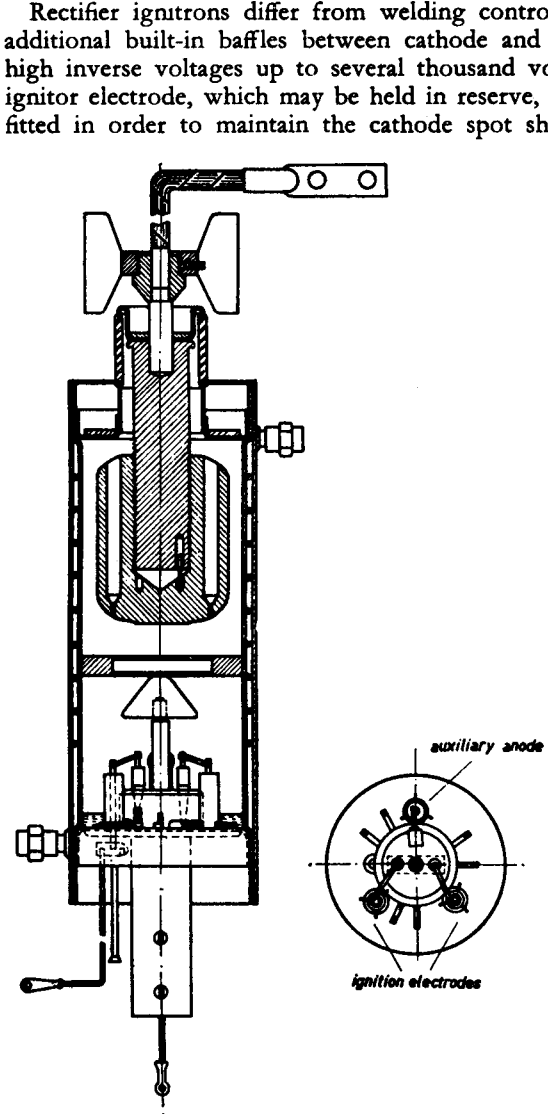


Fig. 5-3. Construction of a rectifier ignitron.

of approximately 3.5 kW have to be dissipated, and because the heat capacity of the electrodes of ignitrons — as of all electron tubes — is comparatively small, temperature equilibrium is reached very quickly. Moreover, high momentary current peaks cause a rapid rise of vapour pressure in the tube, approximately 0.75 grammes of liquid mercury being vaporised for an electrical quantity of 100 ampere · seconds. Care must therefore be taken that adequate water cooling is provided to keep the gas pressure low enough to avoid arc-back or loss of control. Roughly speaking, the mercury vapour pressure is doubled for every 10 °C of

fall temporarily below a value of about 10 amperes, as may happen, for example, in rectifier circuits if the connected load produces a counter electromotive force.

Fig. 5-3 is a sectional drawing of a rectifier ignitron. Rectifier ignitrons can, of course, be used also in welding equipment if connected in inverse-parallel. Similarly, a welding control ignitron may be used for rectification if due precautions are taken.

The limiting operating conditions for ignitron tubes are determined by the maximum permissible values of the anode voltage and of the instantaneous anode current. Another important factor is the temperature and the rate of flow of cooling water, for these not only determine the amount of power which can be dissipated but also the mercury vapour pressure within the tube. For example, the Philips Ignitron Type PL 5555 may carry an average anode current of 200 amperes. Since the arc voltage is about 17 volts, power losses

temperature rise, so that the maximum permissible pressure is reached very rapidly.

The lower temperature limit is determined by the freezing point of the cooling water and by such a low vapour pressure that adequate generation of ions cannot occur. The ignitor, however, will operate successfully even in a pool of frozen mercury.

The published data of an ignitron include the values of current and voltage required for ignition and the maximum permissible positive and negative potentials at the ignitor. The positive voltage is usually equal to the anode voltage; the negative voltage must not be more than a few volts, since an inverse current would cause a cathode spot to form on the rod which would become damaged after a very short time. For this reason, a rectifying element, which passes current in one direction only, is always inserted in series with the ignitor.

An important field of application for ignitrons is the control of resistance welding equipment. The electrical portion of this equipment consists of a transformer, and means for closing and interrupting the primary circuit in order to control the duration of the welding time. The devices previously used were chiefly switches and mechanical contactors, but in many cases these have proved unable to withstand the heavy duty demanded. For example, even the currents flowing in the primary circuit are fairly large, and trouble quickly arises due to burned contacts. Again, mechanical contactors possess considerable inertia, and this affects the accuracy of timing, and thus the quality of the weld. Ignitrons, however, do not suffer from either of these defects.

The basic circuit for welding control is shown in fig. 5-4. Here, I_1 and I_2 are the two ignitrons, which are connected back-to-back in the primary circuit of the welding transformer; R_1 and R_2 are resistors, Sel_1 and Sel_2 are selenium rectifiers, and S_1 is an automatic safety switch which interrupts the circuit if the flow of cooling water is insufficient.

Resistor R_3 across the primary of the welding transformer has two functions. The discharge in an ignitron may be considered as a number of small discharges in parallel, each carrying a current of about 10 amperes. When the current decreases, towards the end of a half-cycle, the individual discharges are extinguished in succession, until finally the last one breaks down, the current thus suddenly interrupted amounting to approximately 10 amperes. A voltage surge thus arises at the transformer primary, and might cause damage but for the fact that the primary is by-passed by R_3 . Furthermore, R_3 , by providing additional load, assists in maintaining the discharge if the current in the primary winding drops below the critical value.

The operation of the circuit reproduced in fig. 5-4 may best be followed by assuming that terminal B of the mains is positive at the instant switch S_2 is closed. Current then flows from B through the load to the cathode of ignitron I_2 , and then via the selenium rectifier Sel_2 , through S_2 , S_1 and resistor R_1 to the

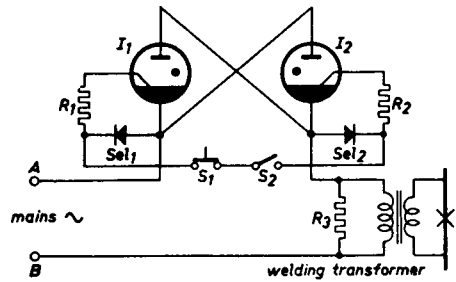


Fig. 5-4. Basic circuit of an ignitron welding contactor. Resistors R_1 , R_2 are often replaced by selenium rectifiers in order to keep the inverse current through the ignitrons low.

ignitor of I_1 . The current impulse now flowing through the ignitor is sufficiently great to produce a cathode spot, thus initiating the main discharge, the anode of I_1 being positive.

When the current flow through I_1 ceases, a positive potential appears at the anode of I_2 due to the inductance of the transformer winding, so that I_2 is ready to ignite as soon as the direction of the current reverses. In this way tubes I_1 and I_2 conduct alternately until the circuit is interrupted by S_2 . Instead of S_2 a special mechanical or electronic timer may be employed to provide exact control of the welding time. Suitable timers will be fully described in a later chapter.

The selection of suitable ignitrons for a given welding equipment is determined

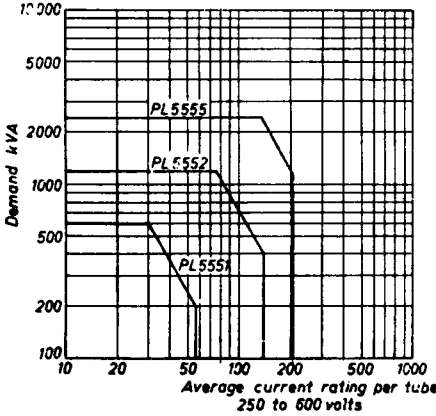


Fig. 5-5. Demand kVA ratings of ignitrons for welding equipment.

by the maximum kVA demand. The output power of each type of ignitron can be ascertained from the diagrams published by the tube manufacturers. In these diagrams the kVA output obtainable from two tubes connected back-to-back is plotted as a function of the maximum average current per tube (see fig. 5-5). These diagrams are usually valid for mains supply voltages between 250 and 600 V_{rms} . For lower supply voltages the values at 250 volts apply. The diagram for the PL 5555 is valid for 2400 V_{rms} only.

value of the current through a pair of inverse-parallel connected tubes is:

If the supply voltage is V and the kVA demand P , the r.m.s.

$$I_d = \frac{1000 P}{V}, \tag{5.1}$$

and the average value of the current per tube is:

$$I_m = \frac{\sqrt{2}}{\pi} \cdot I_d = \frac{1000 \sqrt{2} P}{\pi V}. \tag{5.2}$$

The corresponding averaging time, T_{av} , can be calculated from the average current per tube I_m , corresponding to the maximum kVA demand at the given supply voltage, which is assumed to flow during 0.5 seconds, and the corresponding maximum average anode current per tube, I_a , which can be taken from the diagram. Thus:

$$T_{av} I_a = \frac{1}{2} I_{max} = \frac{\sqrt{2}}{2\pi} \cdot I_{dmax}, \tag{5.3}$$

and

$$T_{av} = 0.225 \cdot \frac{I_{dmax}}{I_a}. \tag{5.4}$$

The time, t , in each averaging time, during which a current of the average value per tube, I_m , may flow is therefore:

$$t = \frac{I_a T_{av}}{I_m} \quad (5.5)$$

The ratio of t to T_{av} is the "duty factor", D . In the form of a percentage,

$$D = \frac{t}{T_{av}} \cdot 100 = \frac{I_a}{I_m} \cdot 100\% \quad (5.6)$$

As an example, the kVA demand of a welding machine may be $P = 1200$ kVA at a supply voltage of 500 V_{rms}. Two ignitrons, Philips Type PL 5552 are used, and are capable of controlling a maximum load of 1200 kVA. From equation (5.1) the r.m.s. value of the demand current is:

$$I_d = \frac{1\,200\,000}{500} = 2400 \text{ amperes.}$$

From fig. 5-5 it is seen that the maximum kVA demand corresponds to a maximum average anode current per tube of 75.6 amperes. So the averaging time will be:

$$T_{av} = 0.225 \cdot \frac{2400}{75.6} = 7.1 \text{ seconds.}$$

The average value of the demand current per tube is:

$$I_m = \frac{1.41 \times 2400}{3.14} = 1080 \text{ amperes,}$$

the maximum conducting time per averaging period is:

$$t = \frac{75.6 \times 7.1}{1080} = 0.5 \text{ second,}$$

and the corresponding duty factor is:

$$D = \frac{0.5 \times 100}{7.1} = 7\%.$$

This means that two PL 5552 tubes may carry an alternating current of 2400 amperes at a supply voltage of 500 volts with a duty factor of 7%. The same information can be obtained from the corresponding characteristic shown in fig. 5-6 in which the current through two inverse-parallel connected tubes is

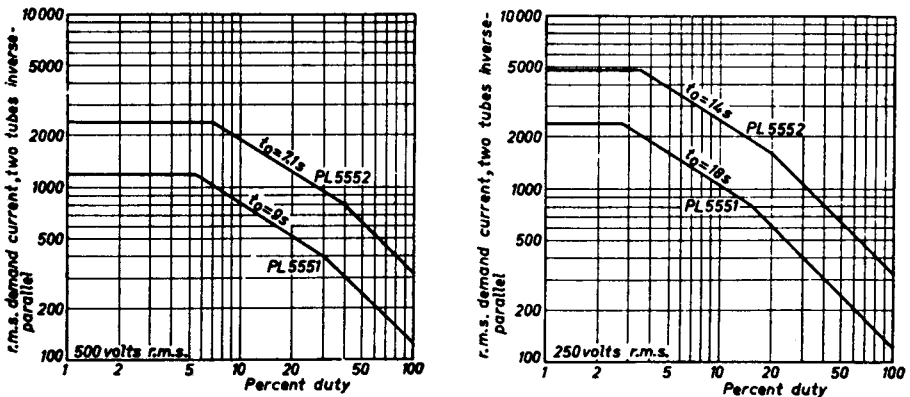


Fig. 5-6. Demand current ratings in A_{rms} as functions of the percent duty for two tubes connected in inverse parallel, for different types of ignitron and supply voltages.

plotted as a function of the duty factor, for different supply voltages and types of tube.

In some cases it may be desirable to control the power, and consequently the amount of heat applied to the work, by delaying the instant at which the ignitrons fire in each half cycle. Adjustable delay can be applied by controlling each ignitron by a separate thyatron as shown in fig. 5-7. The thyratrons T_1 and T_2 are connected in series with the ignition electrodes of I_1 and I_2 respectively. Assuming terminal B to be positive, T_1 will be fired by a positive peak applied

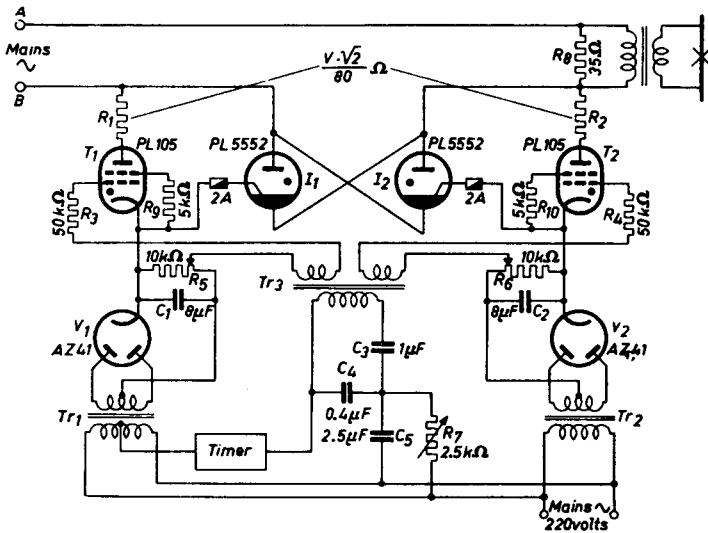


Fig. 5-7. Ignitron welding contactor with heat control.

to its control grid, and a current surge of approximately 40 amperes will flow through T_1 and the ignition electrode of I_1 , firing the ignitron. In the same way I_2 will be fired via T_2 in the following (negative) half cycle. The igniting peaks applied to the grids of T_1 and T_2 are taken from the peaking transformer Tr_3 , and are in antiphase. The primary of Tr_3 is fed from a phase-shifting network, the chief components of which are the primary of Tr_1 , the capacitor C_5 and the variable resistor R_7 . The operation of this type of phase-shifting network has already been explained in Chapter 3. The thyratrons are also given an adjustable negative grid bias supplied by two rectifiers comprising the tubes V_1 and V_2 and their associated circuits. By means of R_7 , the phase delay of the peaks applied to the thyatron grids can be adjusted in order to control the mean current flowing through the ignitrons and the primary of the welding transformer, in a similar way to that previously shown in fig. 4-8.

Fig. 5-8 again shows the voltage and current curves appearing in such a circuit, with and without phase delay. For the sake of simplicity the theoretical case in which $\cos \phi = 0$ has been assumed.

Another important application of ignitrons is in industrial rectifiers, where continuous control of the output power can be obtained by phase-delayed ignition. Such rectifiers are employed, for example, for power supply to rolling mills, conveyor belt installations, electric locomotives and so forth. The basic

circuits for these applications are similar to those already described in Chapters 2 and 3. The special firing circuits, however, will be described in a later chapter. An ignitron, Philips Type PL 5555, designed for rectifying currents up to 200 amperes, is illustrated in fig. 5-9.

The life of ignitrons may be expected to be even greater than that of normal gas-filled tubes. When ultimate failure occurs it may be due to air leakage or to ignitor failure. Air leakage may be recognised by the tube passing current under certain circumstances without the ignitor being operated.

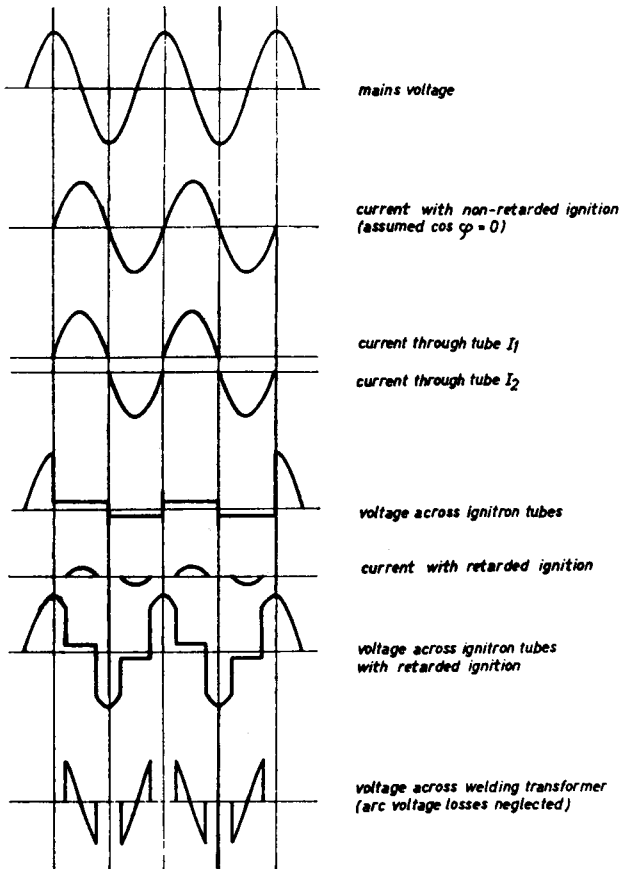


Fig. 5-8. Voltage and current curves of a device according to fig. 5-7 with and without phase delay ($\cos \varphi = 0$ assumed).

The vacuum may be tested by applying a potential of from 5 to 10 kV between cathode and anode, a series resistor being included to limit the current to from 5 to 10 milliamperes. If the tube passes current for longer than one minute, gas "clean-up" may be attempted by operating the tube carefully under reduced load. If the tube again passes current when the high tension test is repeated, the tube should be replaced. Another simple test for air leakage can be made with a low frequency spark coil, connected to the anode lead. If a continuous

reddish glow is seen in the interior of the tube, it has become "gassy" and should be replaced. Air leakage often causes arc-back, which is usually accompanied by severe flashing or showers of red-hot sparks visible through the anode seal.

Ignitrons with burned igniting electrodes, caused for example by reverse current, are inclined to misfire. This failure can be confirmed by measuring the resistance between the ignitor and the mercury cathode, for instance with a "Philoscope". The cold resistance in a good tube is usually between 10 and 150 ohms, but may be as low as 5 ohms or as high as 500 ohms. It may also

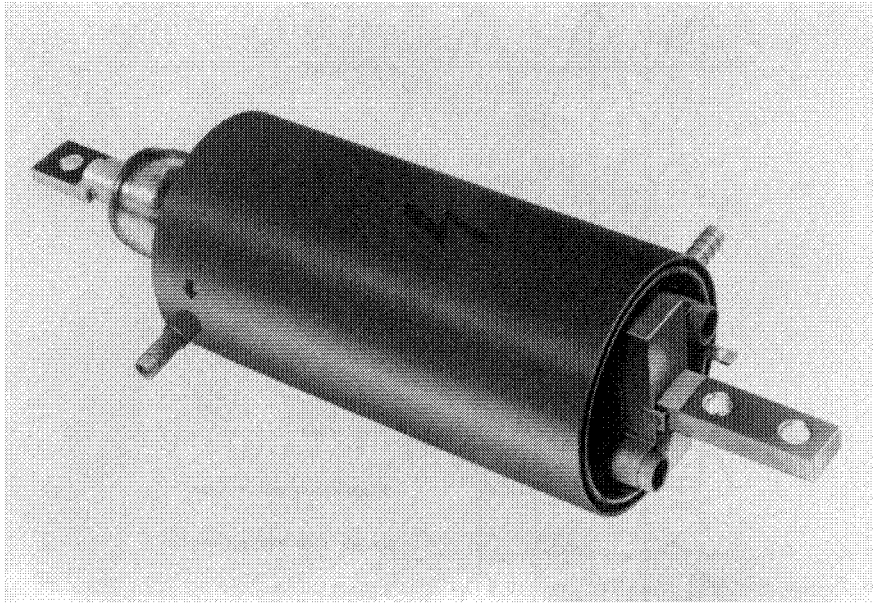


Fig. 5-9. 200 ampere rectifier ignitron.

happen that the igniting rod has become "wetted" by the mercury. This may occur if the tube has carried excessive current, when cathode spots may form on the steel wall of the envelope. Metal may then be vaporised and will ultimately enter the mercury pool, later to be re-evaporated by the arc appearing in each cycle between the ignitor and the mercury pool. The metal will then be deposited on the igniting rod and will form an amalgam with the mercury. A simple method of testing for this condition is to measure the resistance between ignitor and cathode, tipping the tube slightly to withdraw the ignitor slowly from the mercury pool. If the ignitor is in good condition a gradual increase of resistance should take place, but if the ignitor is "wet" the resistance will remain constant for a time, and then suddenly jump to a higher value.

When a tube approaches the end of its useful life it becomes "hard-starting", that is to say a longer and higher current impulse through the igniting electrode is required to start the main discharge. In the firing circuit described above, this defect may be recognised by intense flashing of the thyratrons during each firing operation. Careful supervision of the performance of the equipment will indicate when it becomes necessary to replace a hard-starting tube.

Another tube occasionally used in industrial electronic equipment is the Excitron. This is also a gas discharge tube with mercury-pool cathode, ignition electrode and an auxiliary anode. Unlike the ignitron, however, the ignitor of the excitron is operated only once when the equipment is switched on, the cathode spot thus generated being maintained continuously by the auxiliary anode which is fed from a d.c. source.

Furthermore, the excitron has both an anode and a control grid, so that, the auxiliary discharge once started, it behaves in a similar way to a thyatron. Since the envelope is usually made of stainless steel, the mercury pool cathode is surrounded by a ring of insulating material to prevent the continuously maintained cathode spot approaching the side walls. In addition, special shields and baffles are provided between cathode and anode to render the tube capable of withstanding a high inverse voltage.

Excitrons are chiefly employed in rectifiers handling large currents. In many cases, however, ignitrons or multi-anode rectifying valves are preferred, so that the range of application of the excitron in the industrial field is not extensive.

Fig. 5-10 is a sectional diagram of an excitron. The electrical data of this class of tubes are very similar to those of ignitrons and thyratrons, and a fuller description is therefore

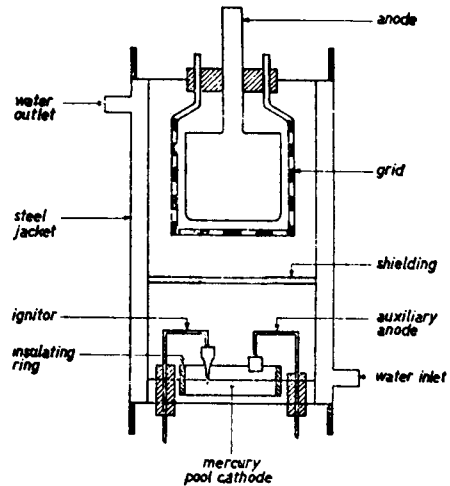


Fig. 5-10. Sectional diagram of an excitron tube.

not necessary.

6. VOLTAGE STABILISING TUBES

Voltage stabilising tubes are gas-filled discharge tubes which can be used for maintaining a voltage at a constant value or for smoothing voltage fluctuations caused by variations of the load or of the mains voltage. They may also be employed as voltage limiters and for generating oscillations of saw-tooth waveform. It will thus be seen that these tubes have a wide field of application in electrotechnics, in radio and television, and in industrial electronics.

The voltage reference tube is a special type of stabilizing tube. It has the advantage of keeping the voltage extremely constant within narrow current limits (a few millivolts only) even after prolonged use, which makes it particularly suitable for providing a reference voltage, for example in stabilizing equipment, thus rendering the use of standard cells superfluous. As a rule voltage reference tubes are designed for currents up to a few milliamperes only.

The special properties of voltage stabilising tubes arise from the shape of the curve relating the voltage across a cold cathode glow discharge and the current flowing through it. This curve is of the form illustrated in fig. 6-1.

If a direct voltage of a certain critical value (known as the "ignition voltage",

V_{ign} is applied between the cathode and anode of the tube, the electrons emitted by the cathode (negative electrode) attain sufficient speed to ionise any gas atoms with which they may collide. The positive ions thus produced are attracted to the cathode, which they strike with sufficient force to liberate further electrons. Eventually a continuous discharge is set up. If, now, the current through the tube increases, the amount of ionisation also increases to a point at which the discharge can be maintained by primary electrons travelling at much lower speeds, that is to say the voltage across the tube necessary to maintain the discharge is reduced to a value known as the "operating voltage".

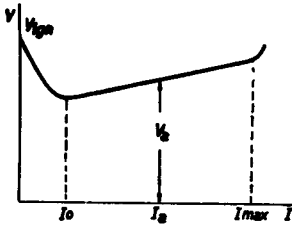


Fig. 6-1. Voltage-current diagram of a cold cathode glow discharge.

Further increase of current through the tube involves only a small increase of the voltage across the tube, this is mainly the voltage drop occurring between the cathode and a positive ion space charge.

It is upon this fact — that over a given range of current values the voltage across the tube is substantially constant — that the operation of the voltage stabiliser is based, the actual value at which the voltage is stabilised being the average value (V_a) of the operating voltage. It should be stated, however, that if the current is increased beyond a certain value I_{max} , the voltage rises again and the nature of the glow discharge changes to an arc discharge. Under this condition, the heat generated by the rapidly increasing current damages the tube, and it is therefore essential that a protective resistor of suitable rating be connected in series with the stabiliser tube unless the voltage source has a sufficiently high internal resistance to limit the tube current to a safe value.

It will be obvious that the flatter the curve between the points I_0 and I_{max} in fig. 6-1, the more accurate will be the stabilising effect. If this part of the curve is assumed to be linear, it may be described by the equation:

$$V = V_a + \Delta I \cdot R \tag{6.1}$$

R is the so-called a.c. resistance and is:

$$R = \frac{\Delta V}{\Delta I} \tag{6.2}$$

Obviously a small value of R is desirable for good stabilisation.

Fig. 6-2 is the basic circuit for stabilising a direct voltage which is subject to fluctuations. It is, of course, essential that the supply voltage V_b is greater than the ignition voltage of the stabiliser tube. If V_b changes by an amount $\pm \Delta V_b$, there will be a corresponding change in the current through R_1 ; but since the voltage V_a across the tube is practically constant and independent of the current through the tube, the whole of the variation in V_b will appear across R_1 . The change in the current through R_1 is therefore:

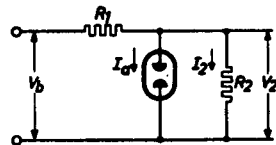


Fig. 6-2. Basic voltage stabilising circuit.

$$\Delta I = \frac{\Delta V_b}{R_1} \tag{6.3}$$

Because the voltage across the load R_2 , and thus the current I_2 flowing through

it, are practically constant, the change ΔI in R_1 is almost entirely carried by the tube, so that the nett voltage change across the load is, according to equation (6.1):

$$\Delta V_2 = \Delta I \cdot R = \frac{R}{R_1} \cdot \Delta V_b. \quad (6.4)$$

In order to ensure good stabilisation, R_1 should therefore be as large as possible. However, an upper limit is imposed for R_1 by the minimum current I_0 required to maintain the discharge (see fig. 6-1). The following condition must therefore be fulfilled:

$$V_b - \Delta V_b - V_a \geq R_1 (I_0 + I_2), \quad (6.5)$$

or

$$R_1 \leq \frac{V_b - \Delta V_b - V_a}{I_0 + I_2}. \quad (6.6)$$

A further limitation is imposed by the necessity of ensuring that ignition of the tube can occur. When the tube is non-conducting, the supply voltage V_b appears across the resistors R_1 and R_2 in series, and care must be taken that the part of this voltage appearing across R_2 is greater than the ignition voltage V_{ign} . The following condition must therefore be satisfied:

$$\frac{V_b - \Delta V_b}{R_1 + R_2} \cdot R_2 > V_{\text{ign}}, \quad (6.7)$$

or, with

$$R_2 = \frac{V_a}{I_2},$$

$$R_1 < \frac{(V_b - \Delta V_b - V_{\text{ign}}) V_a}{V_{\text{ign}} I_2}. \quad (6.8)$$

Experience shows that a lower limiting value of R_1 results from condition (6.8) than from condition (6.6). This means that, due to tube ignition requirements, the stabilising effect obtained is less than that which would be possible if R_1 could be chosen solely by consideration of the minimum tube current requirement.

Finally R_1 must be chosen with due regard to the maximum permissible current I_{max} through the tube. If the supply voltage changes from $V_b - \Delta V_b$ to $V_b + \Delta V_b$, the current through the tube increases, but must not be allowed to exceed the value I_{max} . Since the voltage across R_2 , and thus the value of I_2 are practically constant, the following condition must be satisfied:

$$R_1 \geq \frac{V_b + \Delta V_b - V_a}{I_{\text{max}} + I_2}. \quad (6.9)$$

This gives the lower limit for the series resistor R_1 .

The following calculation for a practical case will serve by way of example. Let it be assumed that a stabilised direct voltage of 85 volts at a load current

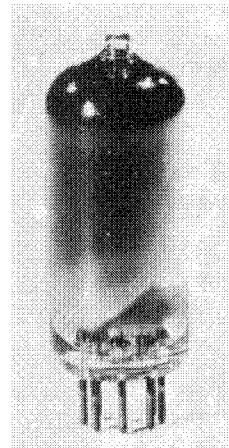


Fig. 6-3. Philips 85A2 voltage reference tube.

of 2 milliamperes is required, and that the nominal mains voltage of 200 volts is subject to fluctuations of $\pm 15\%$ — a percentage fluctuation which is quite typical of present-day conditions. Further, let it be assumed that a voltage reference tube Type 85 A2 is selected *). This tube, which is illustrated in fig. 6-3 has the following characteristics:

Operating voltage (V_a)	85 V
Ignition voltage (V_{ign})	125 V
A.C. resistance (R) (average value)	300 Ω
Maximum tube current (I_{max})	10 mA
Minimum tube current (I_0)	1 mA

According to condition (6.6):

$$R_1 \leq \frac{200 - 30 - 85}{1 + 2} = 28 \text{ k}\Omega,$$

However, according to condition (6.8):

$$R_1 < \frac{(200 - 30 - 125) \times 85}{125 \times 2} = 15.3 \text{ k}\Omega,$$

and according to condition (6.9):

$$R_1 \geq \frac{200 + 30 - 85}{10 + 2} = 12.1 \text{ k}\Omega.$$

A value of 15 k Ω for R_1 will therefore be chosen, and the fluctuation of the output voltage, ΔV_2 will be, according to equation (6.4).

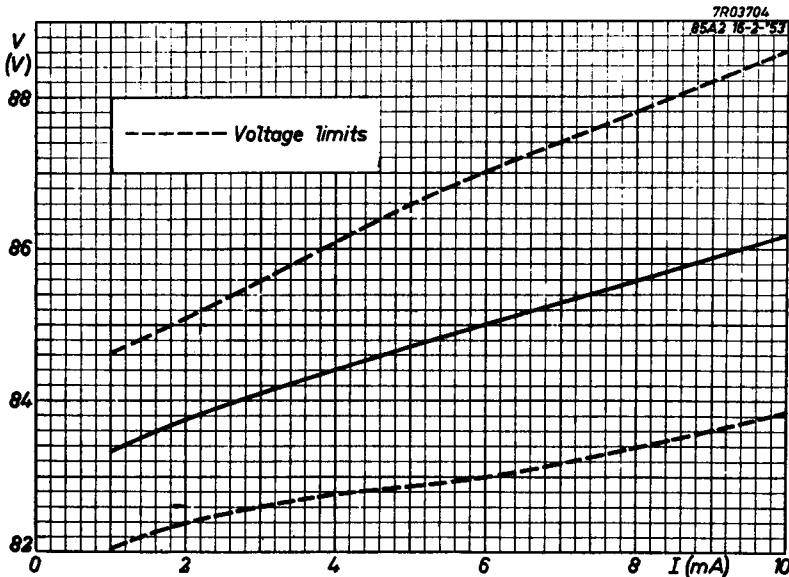


Fig. 6-4. Voltage-current characteristic of the 85 A2 stabilising tube.

*) Instead of the voltage reference tube 85 A2 a voltage stabiliser such as the 90 C1 can be used if the requirements imposed on the voltage stability are not too stringent.

$$\Delta V_2 = \frac{300 \times 30}{15\,000} = 0.6 \text{ V.}$$

It is thus seen that a variation of about $\pm 15\%$ in the input voltage results in a change of only about $\pm 0.7\%$ in the output voltage across the 85 A2 tube.

If a higher stabilised voltage than that according to the operating voltage of a given tube is required, two or more stabilising tubes may be connected in series. The overall characteristic of such a combination can be calculated by adding the voltages across the several tubes for the appropriate current, while the overall a.c. resistance is, of course, the sum of the a.c. resistances of the individual tubes.

The voltage required for igniting such a chain of tubes need not, however, be the sum of the ignition voltages of all the tubes. If, as shown in fig. 6-5, one of the tubes is by-passed by a resistor R_3 of high value, the remaining two tubes will ignite first, after which a voltage equal to twice V_a appears across them, and the remaining part of the input voltage suffices to ignite the by-passed tube.

When it is desired to stabilise the voltage between very close limits, the multiple stabiliser circuit shown in fig. 6-6 may be employed. Here, tubes *I* and *II* produce a certain degree of stabilisation so that a substantially constant current flows through tube *III* which, in turn, finally stabilises the voltage appearing at the slider of the potentiometer. The stabiliser Type 85 A2 is particularly suitable for this appli-

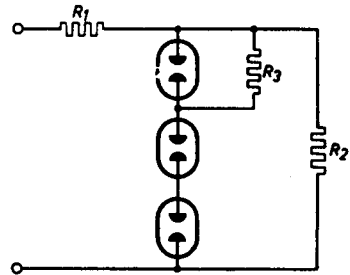


Fig. 6-5. Series connection of stabilising tubes

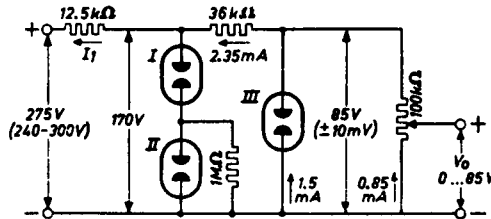


Fig. 6-6. Multiple stabilising circuit.

cation, and using three such tubes in the circuit shown, the fluctuations in V_0 are only $\pm 0.005\%$ for a $\pm 10\%$ variation of the input voltage.

It should be mentioned that it is not permissible to connect stabiliser tubes in parallel in order to operate at a higher current. In such circumstances tubes having a larger current range must be used.

Voltage stabilising tubes are occasionally used for generating saw-tooth oscillations. The basic circuit is shown in fig. 6-7. The tube is connected to a direct voltage source via the series resistor R_1 , and is also shunted by the capacitor C . The capacitor charges via R_1 until the voltage across C is equal to

the ignition voltage of the tube, when the tube ignites and the capacitor discharges; the initial value of the discharge current corresponding to point 2 on the tube characteristic is also reproduced in fig. 6-7. The voltage across the capacitor there-

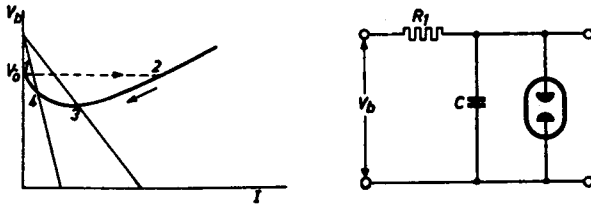


Fig. 6-7. Saw-tooth oscillator circuit.

fore decreases, and so does the discharge current. The value of the voltage follows the curve from point 2 towards the point 3. At the same time, however, current also flows from the supply source to the capacitor through R_1 , thus charging it once more.

If the discharge current through the tube is equal to the charging current through R_1 , a condition of equilibrium is set up so that the voltage across the capacitor becomes constant, and no oscillations will be generated. This will be the case if the line corresponding to the equation:

$$v = V_b - IR_1 \tag{6.10}$$

intersects the curve at the point 3. If, however, the value of R_1 is sufficiently high, the intersection will be at some such point as 4, the charging current via R_1 will be less than the discharge current through the tube, and the voltage across C will continue to fall until the tube is extinguished. The capacitor will then charge up again to the value of the ignition voltage, and the cycle repeats. To make the circuit operative, therefore, the value of R_1 must be so chosen that the intersection of the line and the tube characteristic curve falls on the left-hand side of the lowest point of the curve.

In electronic regulators, where some variable physical quantity, such as a pressure, temperature, velocity etc., is represented by a voltage, a very constant reference voltage is required for the purpose of comparison. Such a constant

voltage can be obtained by using voltage reference tubes, such as Philips Type 85 A2. This tube has a cathode of pure molybdenum, and in order to avoid damage to the cathode surface and contamination of the gas filling, the inner surface of the glass envelope is coated with a layer of molybdenum. This coating serves as a screen between the bulb and the gas discharge, ensuring high stability of the operating voltage throughout life.

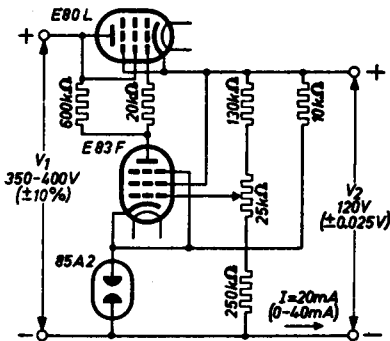


Fig. 6-8. Voltage stabilising circuit using an 85A2 reference tube.

Fig. 6-8 is the basic circuit for this application, the 85 A2 being used as a voltage reference tube. Its operation is

as follows: Increase of the input voltage V_1 causes a corresponding increase of the output voltage V_2 , so that the control grid of the E 83 F pentode, which is fed from a potentiometer across V_2 becomes more positive. The anode current of the E 83 F therefore rises, and the potential at its anode and thus at the control grid of the power pentode E 80 L decreases. The voltage drop across the E 80 L therefore increases, thus tending to stabilise the value of the output voltage V_2 .

The effectiveness of the stabilising process depends upon the constancy of the potential at the cathode of the E 83 F, and it is the function of the 85 A2 to ensure the constancy of this reference voltage.

The maximum current output of this circuit is determined by the maximum permissible anode current of the E 80 L. Higher outputs may be obtained by using a power tube of higher anode current rating.

When using voltage stabilising tubes it is important to remember that the electrodes have been polarised in the course of manufacture. The electrode marked "Anode" must therefore always be connected to the *positive* terminal of the voltage source.

7. PHOTOCELLS

During recent years the practical applications of the photoelectric cell, or photocell as it is more usually termed, have greatly multiplied. In addition to their familiar use in sound-film equipment, they are now employed in a wide variety of devices for automatic control, signalling and warning, particularly in connection with counting, and checking the uniform quality of finished products. They are also used in alarm and protective circuits in banks, museums, shops, offices and so forth.

Photocells are diode tubes consisting of an anode and a cathode sealed into a glass envelope which is either highly evacuated or gas-filled. Unlike that of the thermionic diode, the cathode of a photocell is photo-emissive, that is to say it emits electrons when irradiated by light. The cathode is usually semi-cylindrical in form, while the anode is usually a single wire situated approximately at the centre of the half-cylinder. The chemical structure of the photo-emissive cathode is rather complex, and usually consists of a layer of pure silver on which is deposited a thin layer of the oxide of one of the alkaline metals such as caesium, throughout which there is distributed a certain quantity of the free atoms of the metal.

When light falls upon such a cathode, the free atoms are ionised, electrons being liberated. These electrons then move towards the positive anode. The amount of light energy required to ionise a given atom depends upon the location of that atom with respect to its neighbouring atoms.

Owing to the random nature of the structure of a photo-emissive cathode, its sensitivity is not uniform over its whole surface. If, however, the light beam operating the cell is made divergent by means of a suitable lens, so that the whole of the cathode surface is irradiated, the cell will operate at a constant mean sensitivity. Errors due to the non-uniform distribution of the free atoms over the cathode surface cancel each other out, and furthermore, the risk of local overloading of the cathode will be avoided.

The spectral response of a photocell depends upon which alkali metal is

employed in the cathode. For example, caesium cathodes have a maximum sensitivity when irradiated with light having wavelengths between 7000 Å and 9000 Å, that is to say in the red and infra-red region of the spectrum as shown in fig. 7-1. Such cells are usually preferred in industrial devices since their spectral response corresponds to the more commonly used sources of

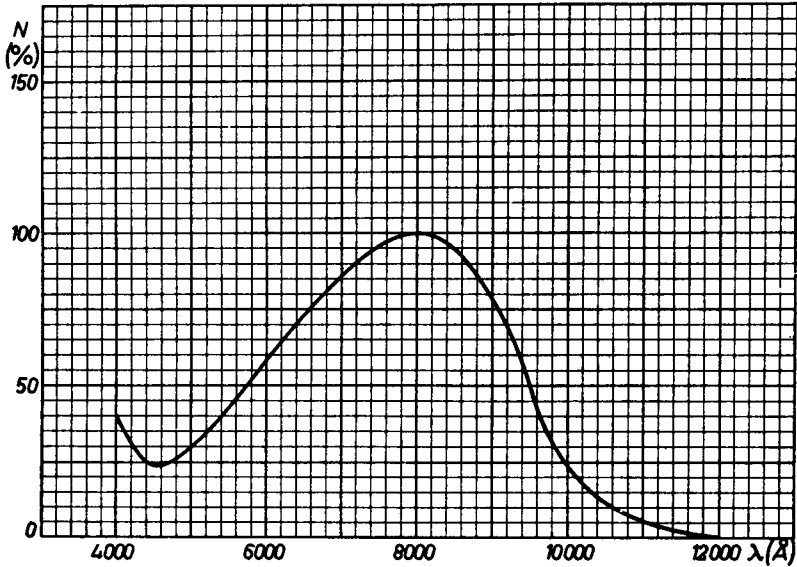


Fig. 7-1. Spectral response curve of a photocell with caesium cathode.

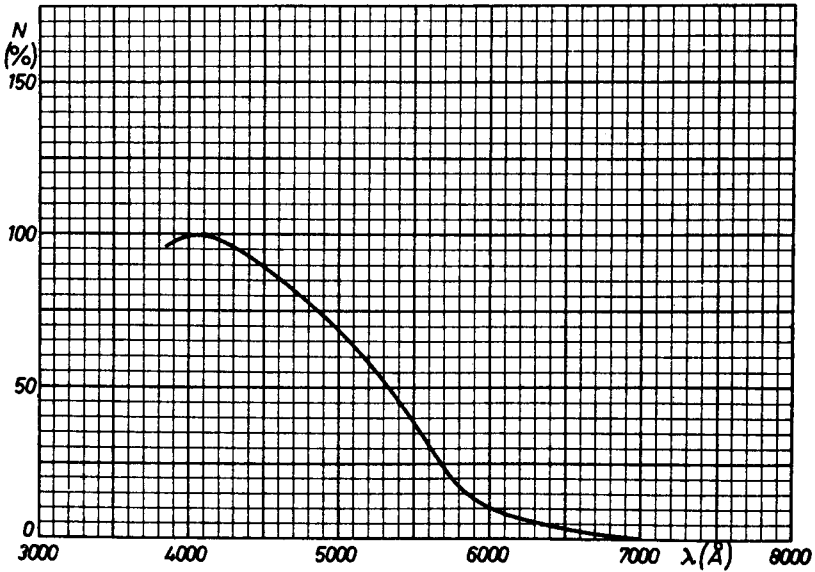


Fig. 7-2. Spectral response curve of a caesium-antimony cathode.

illumination. A very good blue-sensitive cathode is the caesium-antimony cathode. As seen from the spectral response curve reproduced in fig. 7-2, the greatest sensitivity of the caesium-antimony cathode is the range of wavelengths between 3800 Å and 4500 Å, that is to say for violet and blue light. Ultra-violet rays which have wavelengths below 3850 Å are absorbed by the glass envelope of the tube, so the curve does not continue below this value. If photocells sensitive to ultra-violet radiations are required the bulb must be provided with a quartz window through which these rays may pass and reach the cathode.

The maximum permissible cathode current for a particular type of tube is specified in the published data, and is the maximum permissible current when the entire cathode surface is uniformly illuminated. If, however, the operating conditions are such that only a part of the cathode surface is illuminated, the cathode current must be correspondingly reduced.

Even when a photo-cathode is not illuminated it will emit a small number of electrons, the emission depending upon the ambient temperature. The "work function" of the cathode surface is so low that even at room temperature some thermionic emission is observed. This "dark current" is, however, very small, being of the order of 10^{-8} amperes per square centimeter for a caesium cathode. For normal industrial applications, therefore, the dark current can usually be neglected since fairly high light intensities will be used, so that the ratio of photo emission to dark current will be large. Only in very sensitive apparatus where great accuracy is required will the dark current introduce appreciable errors, but this can be corrected by compensating circuitry.

In the basic circuit reproduced in fig. 7-3, a high-vacuum photocell is connected in series with a resistor R . If a sufficiently high positive potential (approximately 80 volts) is applied to the anode, all the electrons emitted by the cathode will pass to the anode. Further increase of the anode voltage does not result in any increase in the tube current — in other words the tube operates as a saturated diode.

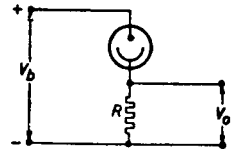


Fig. 7-3. Basic circuit of a high-vacuum photocell.

This is shown in the characteristic curve, fig. 7-4. If, however, the voltage applied to the anode is less than the saturation value, not all the electrons emitted by the cathode will reach the comparatively small anode, a proportion of them returning to the cathode, so that the value of the anode current is smaller than the saturation current, i.e. the sensitivity of the tube is reduced. To obtain full sensitivity, the photocell should be operated at an anode voltage of between 90 and 100 volts.

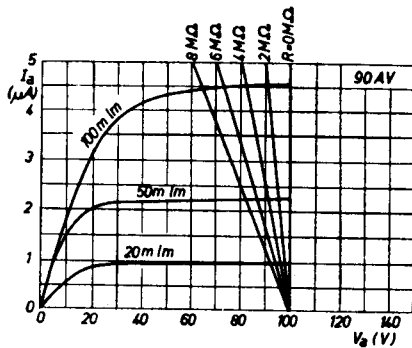


Fig. 7-4. Characteristics of a high-vacuum photocell with load lines.

The sensitivity of vacuum type photocells is somewhat low — in the order of 20 to 40 micro-amperes per lumen. If, however, the tube contains a small quantity of an inert gas, the sensitivity

will be considerably increased by "gas amplification", which may be explained as follows. When electrons pass from the cathode to the anode in a gas-filled tube, a proportion of them will collide with the gas atoms. If the velocity of the electrons is low, such collisions will not affect the anode current, but if the electrons have traversed a potential which exceeds the ionisation voltage of the gas, they will have attained sufficient momentum to ionise the gas atoms with which they collide. This means that at each collision a positive ion is produced which moves towards the cathode, and a free electron is released which, together with the original electron, moves towards the anode. These two electrons may now ionise further gas atoms, so that for one electron emitted from the cathode, several electrons ultimately reach the anode. Moreover, the positive ions reaching the cathode strike it with sufficient force to release further electrons, and these, in their turn, ionise further gas atoms. It will be clear that by this process, the original photo-electric current is considerably amplified, and that the gas amplification factor increases with the tube voltage since the velocity of the electrons is directly proportional to the voltage gradient traversed.

There is, however, an upper limit beyond which the anode voltage may not be increased. Excessive voltage will result in the setting up of a glow discharge across the tube; the anode current is then limited only by the value of the series resistor, and may reach a value at which the bombardment of the cathode by positive ions results in damage to the photocell. Care must therefore be taken that the maximum voltage specified by the tube manufacturer is never exceeded. Since the ignition voltage of gas-filled photocells ranges from about 100 volts to 150 volts according to type, the maximum permissible anode voltage is usually in the order of 75 to 100 volts. The gas amplification factor obtainable is in the region of 5 to 10.

A typical I_a-V_a characteristic of a gas-filled photocell is shown in fig. 7-5. Over the curved parts of the characteristic the voltage developed across the load resistor R will not be exactly proportional to the amount of light falling on the cathode. Gas-filled photocells are not, therefore, very suitable for applications in which very accurate measurements are required. They are, however,

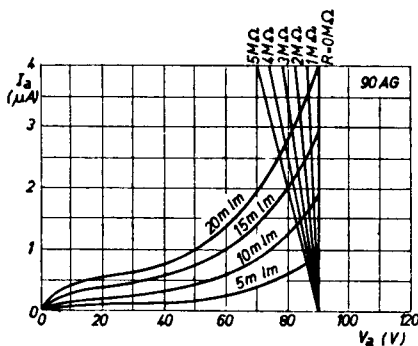


Fig. 7-5. Typical I_a-V_a characteristic of a gas-filled photocell with load lines.

very suitable for sound reproduction in sound-film equipment because, with load resistors of values normally used in this application, distortion due to the non-linearity of the characteristic is negligible.

Gas-filled photocells are also usually preferred for those industrial applications in which great sensitivity is more important than strictly linear response. The sensitivity of gas-filled cells normally ranges up to about 150 micro-amperes per lumen.

It is now necessary to consider the behaviour of a gas-filled photocell when the light falling upon the cathode is subject to very rapid fluctuations. It

is found that, at the higher frequencies, the gas amplification decreases. This is

because the positive ions resulting from the ionisation of the gas are so much heavier than the electrons that their velocity is relatively low. When, therefore, the light is suddenly interrupted, there are still some ions reaching the cathode and these release new electrons capable of ionising further gas atoms. The number of newly formed positive ions is, however, insufficient to maintain the anode current at its original value, and the current will therefore decrease.

Similarly, the anode current will not reach its final value immediately the illumination of the cathode is restored. This means that at the higher frequencies the response of the cell is unable to follow the light fluctuations, and at frequencies above 1000 c/s the amplification begins to fall off. However, this decrease of sensitivity is not serious for frequencies below 10 kc/s, so that over this frequency range the variation in amplification does not sensibly affect practical applications.

When using photocells it must be borne in mind that the photo-electric current is in the order of a few micro-amperes only. Good insulation of the tube socket

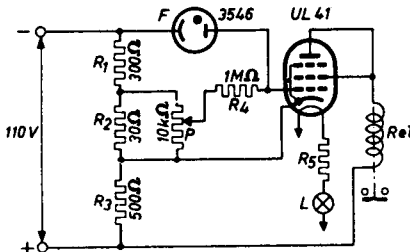


Fig. 7-6. Simple light-controlled relay circuit.

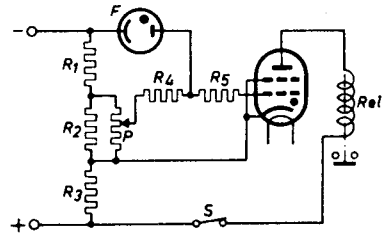


Fig. 7-7. Relay circuit using a thyatron.

and of the wiring is therefore essential in order to avoid leakage. Where a very large load resistor, or a high gain in the following amplifier is necessary, careful screening of the cell and of the wiring is recommended. In addition, care should be taken that the ambient temperature does not exceed 50°C , otherwise the cathode emission may decrease, and thus the sensitivity of the tube. The sensitivity will also decrease temporarily if the cell has been exposed to very intense illumination, for example direct sun-light, even if no anode voltage was applied. The possibility of this must be considered in installing photocell equipment.

A circuit for a simple light-controlled relay is given in fig. 7-6. It uses a gas-filled photocell, and an amplifying pentode Type UL 41. A voltage divider consisting of three resistors R_1 , R_2 and R_3 is connected across the 110 volt d.c. supply, R_3 being at the positive end of the chain. The voltage across R_3 is applied between the cathode and anode of the UL 41 via relay Rel , and the control grid of the pentode receives an adjustable negative bias via potentiometer P . The photocell F (Philips Type 3546) obtains its anode voltage from the voltage appearing across R_1 .

If the cathode of the photocell is illuminated, the photo-electric current flowing through R_4 produces a voltage drop of such a value that the UL 41 is almost cut off. If, now, the light falling upon the cell is interrupted, the photo-electric current ceases, so that the negative control grid voltage falls, and the anode

current of the UL 41 increases and energises the relay, closing the main control circuit. Obviously the relay will open again if light once more falls upon the cell.

In applications such as alarm devices, in which it is desired that the alarm circuit closes on interruption of the light beam and remains closed until the circuit is opened by hand, even if the light beam is restored after a short interval, a special type of relay may be used. But a simpler solution is to use a small thyatron in place of the power pentode. The circuit for this arrangement is shown in fig. 7-7, and differs but little from the previous circuit.

As explained in an earlier chapter, a thyatron, once ignited, remains conductive until extinguished by reducing the anode voltage to a value less than the arc voltage, or even to zero. Thus, in the circuit shown in fig. 7-7, the thyatron, having been ignited by the action of the photocell, will remain conductive so that relay *Rel* continues to be energised until switch *S* is opened, even if the cell is illuminated again.

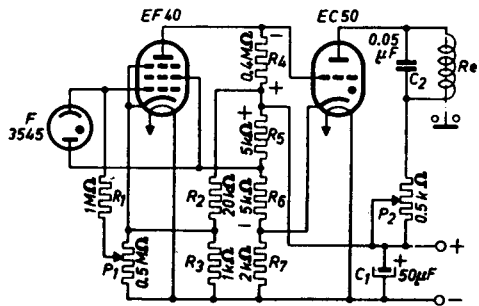


Fig. 7-8. Light-controlled relay circuit with pre-amplifier stage.

before. If the photocell *F* is illuminated, the grid potential becomes less negative, and the EF 40 passes current, thus producing a voltage drop across resistor *R*₄ with the polarity indicated. The grid of the gas triode EC 50 is thus driven negative, and this tube is cut off (see note on page 30). If now the light beam falling upon photocell *F* is interrupted, the grid potential of the EF 40 becomes more negative, the anode current of the EF 40 falls and the potential at the grid of the EC 50 is driven positive causing the tube to ignite and energise the relay.

In this circuit, of course, the gas triode will extinguish and become non-conducting, thus de-energising the relay as soon as the photocell is once more illuminated, since the positive ions will be neutralised by the increasing negative voltage on the grid of the EC 50. If it is desired that the relay should remain energised, the gas triode EC 50 should be replaced by a suitable thyatron such as Philips Type PL 2D 21.

A somewhat similar circuit, operated from an alternating current supply, is shown in fig. 7-9. Both the thyatron and the photocell are fed from the

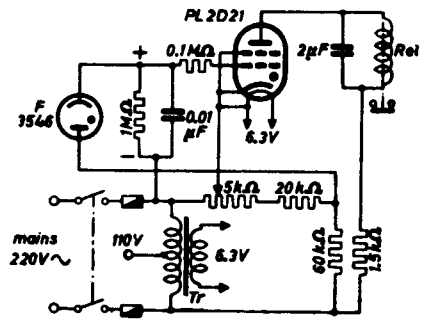


Fig. 7-9. Photoelectric relay circuit operated by a.c. supply.

a.c. supply, but since these tubes do not operate during negative half cycles the response may be delayed by a maximum of 1/100 second. The circuit is so designed that the thyatron ignites, energising relay *Rel*, when light falls upon the photocell. By a slight modification of the circuit, however, the reverse

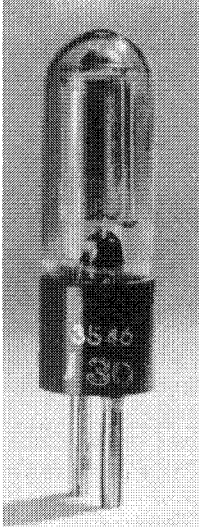


Fig. 7-10. (left) Gas-filled photocell with caesium cathode.

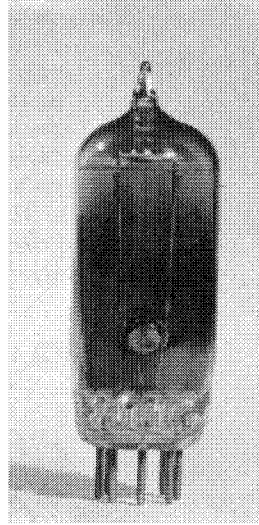


Fig. 7-11. (right) Modern high-vacuum photocell with caesium-antimony cathode.

operation can be obtained, that is to say the relay is energised when the light beam is interrupted.

Fig. 7-10 shows a typical gas-filled photocell with caesium cathode sensitive to red and infra-red light, and in fig. 7-11 is seen a modern high-vacuum cell with caesium-antimony cathode, sensitive to the blue region of the spectrum.

8. TRIGGER TUBES

The name "trigger tube" is sometimes applied to all those forms of gas discharge tube which may serve as relays for starting a process by means of a comparatively small signal. Thus thyratrons, senditrons and the like are sometimes called "trigger tubes". In this book, however, "trigger tube" is reserved for one particular class of tube, namely cold cathode thyratrons, which are particularly suitable for this type of application. An important advantage of the cold cathode thyratrons is that, as they require no cathode heater supply, they are ready for instantaneous operation at all times. They are therefore often used in signalling and controlling equipment which must always be ready for immediate service.

A cold cathode thyatron contains a specially prepared cathode, a main anode and an auxiliary anode or "starter", all sealed into a gas-filled bulb. If a positive potential of a certain value (the ignition voltage) is applied to the starter,

electrons are emitted from the cathode and ionise gas atoms by collision. The positive ions thus produced bombard the cathode and release further electrons, and if a positive potential is applied to the anode the main discharge between cathode and anode is set up.

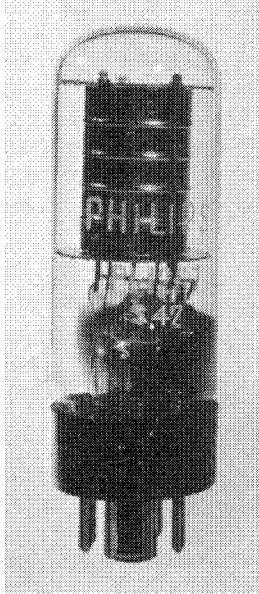


Fig. 8-1. Philips PL 1267 trigger tube.

The voltage drop between cathode and anode is then practically independent of the value of the anode current, so that the anode current is governed only by the applied anode voltage and the resistance in the external circuit. The circuit must therefore be so designed that the anode current can never exceed the maximum rated value specified in the tube data.

The main advantages of the trigger tube are as follows:

- (1) As already stated, the tube is always ready for instant operation, and requires no warming-up time.
- (2) It consumes no energy when not operating — a matter of particular importance if the equipment is fed from batteries.
- (3) There is no wear and tear on the tube unless during operation, and so a long working life may be expected.
- (4) The auxiliary voltage required to ignite the tube does not depend upon the value of the anode voltage.

A typical trigger tube, Philips Type PL 1267, is illustrated in fig. 8-1. This tube is rated for an average anode current of 25 milliamperes, so that a relay having a coil wound for this current may be connected in the anode circuit. However, as the useful life of the tube largely depends upon the average load current, it is advisable to operate the tube at a current somewhat less than the maximum rated value.

The behaviour of a trigger tube under any operating conditions can be ascertained from its ignition characteristic, which is always of the form indicated in fig. 8-2. At any combination of anode and starter voltage within the closed loop no discharge will occur if the boundary of the curve has not been crossed previously. If, however, the tube is subjected to conditions corresponding to a point upon the curve itself or outside the loop, a discharge will be initiated and will be maintained even if the conditions are changed to values corresponding to some point within the loop, the particular electrodes involved in the discharge and the direction of the discharge depending upon the values of the anode and starter voltages when the discharge is initiated.

Thus, section *a* of the loop corresponds to a discharge between starter and cathode (direction of *positive* current). It is seen that the ignition voltage is independent of the value of the anode voltage.

A discharge from anode to cathode occurs when conditions corresponding to a point on section *b* of the curve are applied. Section *c* corresponds to a discharge from anode to starter; section *d* to a discharge from cathode to starter;

section *e* to a discharge from cathode to anode; and section *f* to a discharge from starter to anode.

As, however, the potentials V_a and V_{ah} applied to the anode and the starter respectively are usually both positive, the tube will normally operate in quadrant *I* of the closed loop.

The basic circuit of a light-controlled relay which may be operated direct from the 110 volt a.c. mains is shown in fig. 8-3. This circuit is particularly suitable for warning or signalling equipment which must remain ready for operation over a period of several years without maintenance. A similar device may be used for such purposes as the automatic switching of street lighting, the lighting of business premises and so forth, where the apparatus has to be installed in

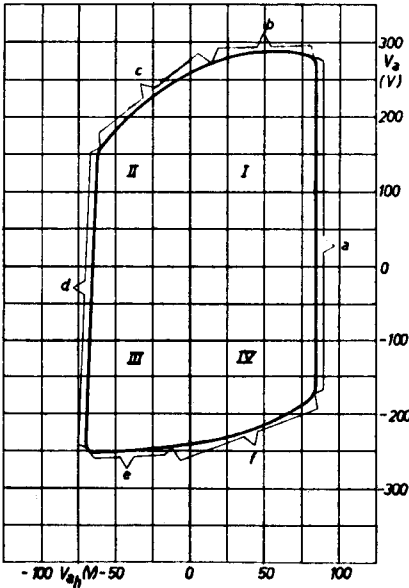


Fig. 8-2. Ignition characteristic of a trigger tube.

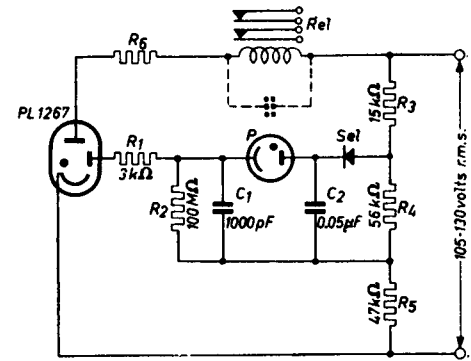


Fig. 8-3. Basic circuit of a light-controlled relay using a trigger tube.

inaccessible positions such as on the roof, chimneys etc.

The operation of this circuit is as follows: Capacitor C_2 is charged via selenium rectifier *Sel* to a voltage which is applied via resistor R_2 to the photocell *P* — which should preferably be of the gas-filled type. When the photocell is illuminated, the photo-electric current produces a voltage drop across R_2 , a positive potential appearing at the upper end of the resistor. This positive direct potential, added to the alternating voltage drop across R_5 , is applied to the starter of the trigger tube PL 1267 via R_1 , and causes the tube to fire during every positive half cycle. Resistor R_6 is a limiting resistor of sufficient value to prevent the anode current of the trigger tube exceeding the maximum permitted value. It is advisable to shunt the relay coil by a capacitor of from 1 to 2 μ F to prevent the relay fluttering by reason of the intermittent nature of the anode current.

A variant of this circuit, in which the trigger tube is ignited if the light continuously falling upon the photocell is interrupted, is shown in fig. 8-4.

By using a suitably designed potential divider or a transformer, these circuits may be operated from 220 volt a.c. mains. If a direct current supply is available

the circuits may also be used as alarm devices. In this case, however, once

ignited, the trigger tube remains conductive, and the circuit has to be opened by hand. If this is not convenient, a gas triode may be substituted for the trigger tube; the circuit will then be interrupted when the grid of the gas triode is made negative.

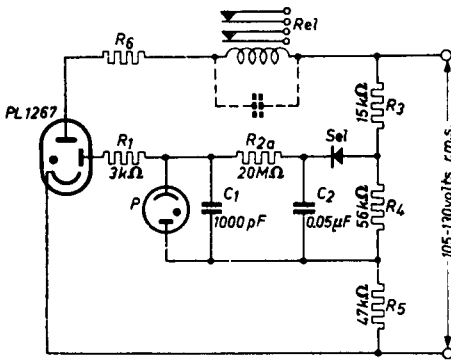


Fig. 8-4. Variant of the circuit shown in fig. 8-3.

A very interesting application of the trigger tubes is in "ring" counter circuits for counting impulses. These circuits may form part of such devices as electronic computers. The basic circuit is given in fig. 8-5. If switch *A* is closed, a positive potential which is, however, insufficient to cause ignition, is applied to the

first trigger tube *V*₁. If, now, a positive impulse is superimposed on the auxiliary electrode potential, *V*₁ ignites, and a positive potential due to the discharge

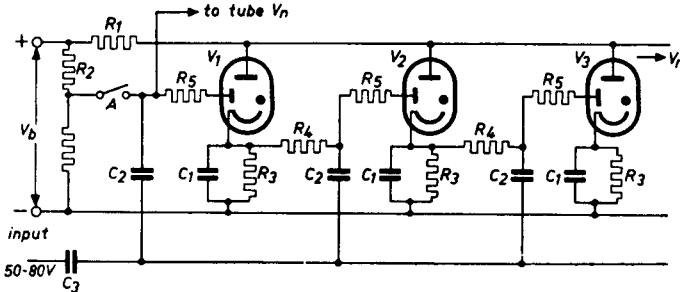


Fig. 8-5. Basic circuit of a counting circuit.

current flowing through *R*₃ is applied to the auxiliary anode of *V*₂, but again is of insufficient value to cause ignition. If, however, switch *A* is re-opened and a second positive pulse is applied to the circuit, *V*₂ will ignite, thus increasing the current through *R*₁ and the voltage drop across it. Because at that instant capacitor *C*₁ still holds its charge, the voltage across *V*₁ is reduced below the operating voltage and *V*₁ will be extinguished. Similarly, a third impulse will fire *V*₃, since this is now the only tube having a positive bias, and simultaneously *V*₂ will be extinguished. This process will continue until the *n*th pulse fires *V*_{*n*}.

In fig. 8-6 the *n*th stage of the counter circuit is shown, and it will be seen that the *n*th tube provides the positive bias for the first tube, *V*₁, so that the (*n* + 1)th pulse fires *V*₁ again, thus closing the ring. Another operation of switch *A* is therefore not necessary. The (*n* + 1)th pulse simultaneously fires tube *V*_{*n*+1}, which forms the input to the next counter circuit. This tube passes current only while capacitor *C*₄ discharges, and becomes non-conductive before *C*₄ can recharge via resistor *R*₆. In this way a positive voltage pulse appears across

resistor R_8 , causing ignition of the first tube in the second ring of the counter circuit. Further impulses applied to the circuit will cause the tubes of the first ring to fire again in succession until the $(2n + 1)^{\text{th}}$ impulse fires the second tube of the second ring counter circuit via V_{n1} .

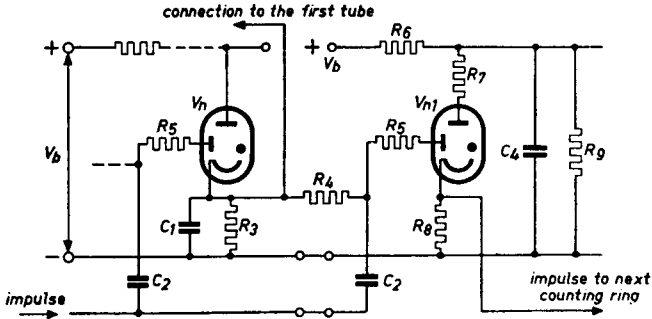


Fig. 8-6. Circuit of the n^{th} stage of the ring counter circuit.

The whole or a part of the resistors R_3 may consist of suitable relays, signal lamps or the like which can indicate the number of pulses counted.

If, in fig. 8-5, the resistor R_1 and the capacitors C_1 are omitted, the preceding tube will remain conducting after every impulse until, after n impulses, all the tubes in the first counter ring are ignited. In this arrangement, which also has practical applications, the anode voltage line has to be interrupted shortly after each complete cycle, thus extinguishing all the tubes. This can be done, for example, by a relay operated by tube V_{n1} .

9. CATHODE-RAY TUBES

Although the cathode-ray tube is seldom employed in electronic industrial control equipment, it is an essential component of a most important instrument — the cathode-ray oscilloscope — which is widely used for measurement and for checking and maintaining industrial electronic devices. It is for this reason that a description of the cathode-ray tube is included in this chapter.

Briefly described, a cathode-ray tube is a thermionic tube in which electrons, emitted from an indirectly-heated cathode located at one end of the tube, are focused into a convergent beam which, impinging on a luminescent screen at the other end of the tube, produces a light spot at the point of impact. By deflecting the electron beam simultaneously in the vertical and horizontal directions, the light spot can be moved to any point on the screen. If the horizontal deflection of the spot is made, say, a linear function of time, and the vertical deflection is made proportional to the instantaneous values of a variable voltage which it is desired to examine, then the light spot will trace on the screen a graph showing the voltage under examination as a function of time.

In the cathode-ray tube, free electrons are emitted from an indirectly-heated cathode, the actual emissive area of which is very small. Surrounding the cathode is a small cylindrical electrode to which is applied a negative biasing potential.

Its main function is to control the intensity of the electron beam. Since this function is analogous to that of the control grid of an amplifying tube, this electrode is also termed the "grid".

The electrode assembly, known collectively as the "gun", also includes two electrodes to which positive potentials are applied. These electrodes, which are usually hollow cylinders or discs with a central aperture, not only accelerate the electrons in the direction of the screen, but also focus the beam into a narrow pencil. This is achieved by the form of the rotation-symmetrical electric field produced by these electrodes. Since the effect of this field upon the electron stream is similar to that of an optical lens upon a beam of light, this part of the electrode system is commonly termed an "electron lens". Correct focusing of the beam on to the luminescent screen is obtained by adjusting the potentials applied to the focusing electrodes.

In order to obtain a sharp, bright trace upon the screen, the positive potential at the final anode must be high, so that the electrons attain a high velocity. The electrons forming the beam mutually repel each other, so that the beam tends to diverge. If, however, the velocity of the electrons is sufficiently great, the electrons will reach the screen before their mutual repulsion has resulted in serious defocusing.

Deflection of the electron beam can be effected in two ways — by means of electric fields applied between pairs of plates within the tube, or by magnetic fields produced by a coil system outside the tube. The second system has the advantage that deflection over a wide angle can be attained so that the overall length of the tube can be comparatively small. It is for this reason that the picture tubes used in television receivers are always arranged for magnetic deflection, since for this service large pictures are required. Furthermore, the focusing of the electron beam in these tubes is usually also achieved by a magnetic field. Magnetic deflection, however, becomes ineffective at high frequencies, and, moreover, requires rather complicated circuitry. Therefore the electron beam in cathode-ray tubes used for oscilloscopes is always electrostatically deflected. Since generally much smaller images are required in oscilloscopes, a deflection angle of about 20° suffices; and this can be easily obtained by electrostatic deflection, the length of the tube still being reasonably small.

Two pairs of deflecting plates are provided, mounted mutually at right angles. The first pair, nearer the cathode, serves for vertical deflection, and it is to these plates that the voltage to be measured is applied. The second pair of plates, nearer the screen, provides horizontal deflection. A saw-tooth voltage, generated by a special oscillator, is usually applied to these plates in order to obtain the time-base. A photograph of the electrode assembly of a typical oscilloscope tube, together with its symbol, is reproduced in fig. 9-1.

If, between the two plates of a pair, is applied a voltage V_d , the electron beam is deflected by the resulting electric field in the direction of the positive plate. The amount of deflection is given by the following formula:

$$a = \frac{1}{2} \cdot \frac{Ls}{V_d d} \cdot V_d \quad (9.1)$$

where

- a = distance of the deflected spot from the centre of the screen,
- s = length of plates in the direction of the beam,
- L = distance between the centre of the pair of plates and the screen,

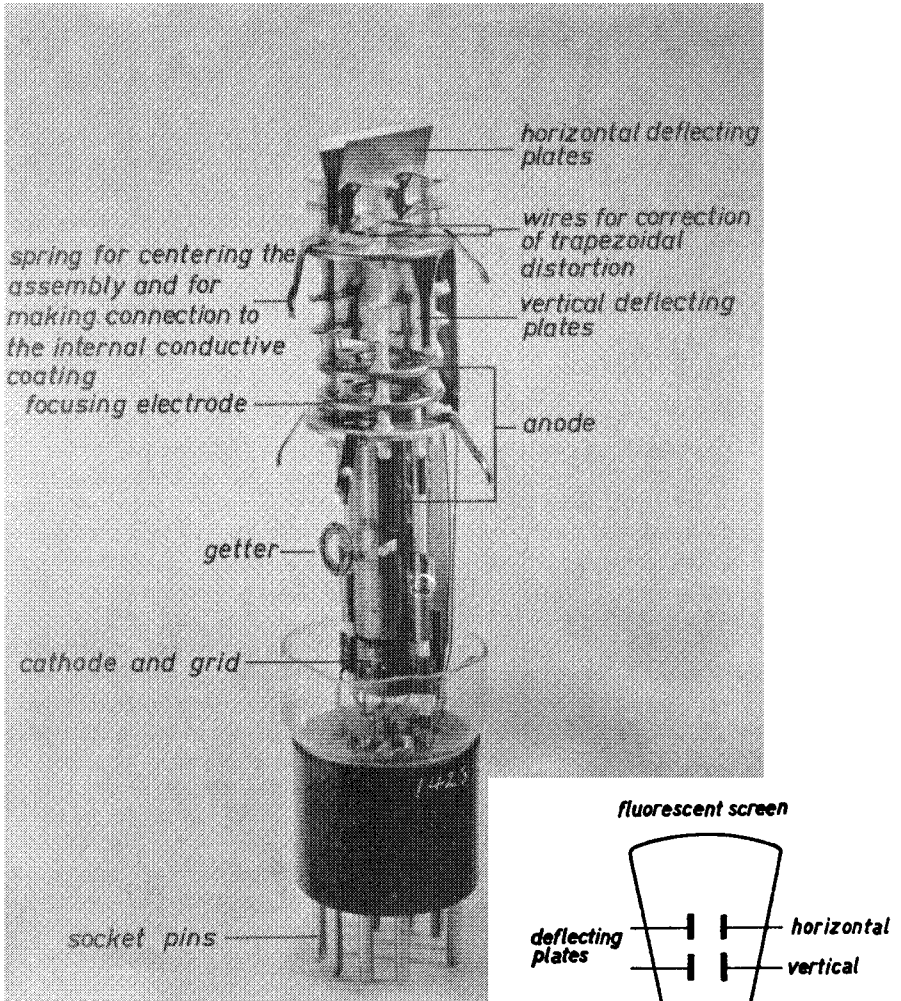


Fig. 9-1. Electrode assembly of a cathode-ray tube.

d = distance between the plates,
 V_a = anode voltage,
 V_d = voltage between plates.

From this formula it will be seen that the amount of deflection, i.e. the deflection sensitivity, decreases as the anode voltage is increased — in other words the beam becomes “stiff”. The conditions for good brightness and for high sensitivity are therefore conflicting. However, tubes of normal construction have adequate deflection sensitivity and brightness for most types of measurement. When very high scanning speeds are required, special tubes are used which incorporate an additional electrode to which a very high potential is applied. This electrode accelerates the electrons after they have left the deflecting plates with a comparatively low velocity.

The simplest way of measuring a voltage is to earth one deflecting plate of each pair, and also the anode, and the positive pole of the high tension unit, and to apply the voltage to be measured and the saw-tooth voltage supplying the time-base to the other two plates respectively. This circuit, however, being asymmetrical with respect to earth, results in certain inaccuracies in the trace unless special precautions are taken. The reason is that the potential at the mid-point between the deflection plates is not constant, but consists of a steady voltage upon which is superimposed an alternating voltage component equal to half the voltage to be measured. The electron velocity is determined by this potential, so that the effective anode voltage is dependent upon the voltage to be measured. Thus, the relation between the voltage to be measured, V_d , and the deflection, a , is not constant and the measurement is not accurate. The inaccuracy is, however, not very serious, and seldom exceeds about 4%.

A somewhat more disturbing consequence of the asymmetrical deflection voltage is trapezoidal distortion. This is due to the fact that the electron beam deflected by the first pair of plates is re-deflected towards the centre of the screen when the potential at the non-earthed plate of the second pair is positive, and towards the periphery of the screen when this plate is negative. This means that the sensitivity of the first pair of plates depends upon the voltage applied to the second pair and this results in trapezoidal distortion.

Both the above faults can be eliminated if the voltages are applied symmetrically to the deflecting plates, but in this case the circuits used for generating the time-base voltage and for amplifying the voltage to be measured are more complex. There are cathode-ray tubes, however, which are specially designed for asymmetric deflection. In these tubes the trapezoidal distortion is compensated by the special shape given to the second pair of plates. The alternating voltage to be measured is applied to the first pair symmetrically, and the time-base voltage is applied to the second pair asymmetrically. In these circumstances the only inaccuracy is that due to the fact that the deflection sensitivity of the second pair depends upon the time-base voltage and results in a slight non-linearity of the time base.

The circuit for asymmetrical operation is so much simpler, however, that where only qualitative examination of electrical magnitudes is required and no accurate measurement is needed, both the voltage to be examined and the time-base voltage are applied asymmetrically. This is quite satisfactory, for example, in cathode-ray oscilloscopes used for the checking and maintenance of industrial electronic equipment.

Cathode-ray tubes for oscilloscopes are made with screens giving green or blue short-persistence traces or with a long-persistence fluorescent screen.

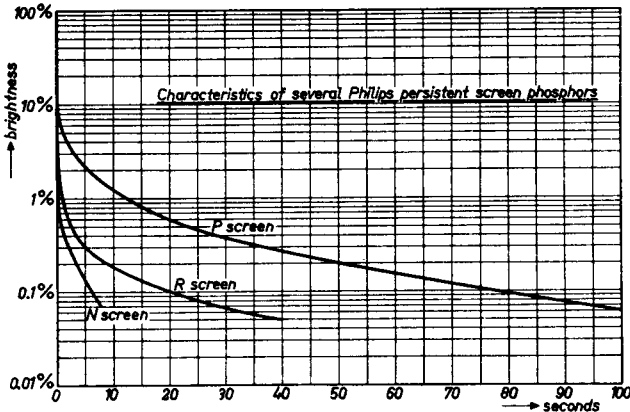


Fig. 9-2. Characteristics of typical persistent screen phosphors.

Generally tubes giving a green trace will be preferred, since this colour is very suitable for direct observation and for photographs taken on negative material. For observing single phenomena or slow-speed processes, a tube with persistent screen should be used, since the trace, once made, remains visible for an appreciable time.

There are several types of screen with different degrees of persistence. Fig. 9-2 shows the performance curves of the Philips type N, R and P screen materials. If pictures are to be recorded directly upon sensitised photographic paper, a blue fluorescent screen is recommended. The spectral responses of green, blue and persistent screens are illustrated in fig. 9-3.

The complete circuit of a simple cathode-ray oscilloscope suitable for industrial applications is given in fig. 9-4. A Philips DG 7-6 cathode-ray tube with a 7 cm diameter screen is used. The equipment includes a pre-amplifier stage for the voltage to be measured, the amplifying tube being e.g. a Type EF 42 high-slope pentode. The amplifier is designed to

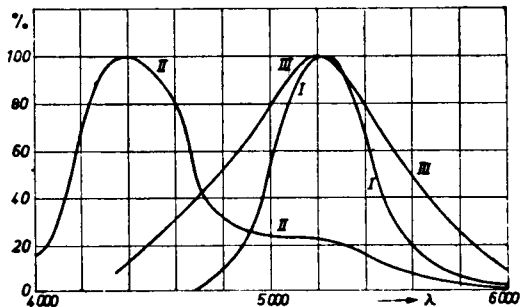


Fig. 9-3. Spectral response characteristics of green (I), blue (II) and persistent screen phosphors (III).

give almost uniform gain over a frequency range from 4 c/s to 130 kc/s. With such an equipment it is possible to amplify and to record with small distortion even voltages with a form greatly differing from a sine wave; e.g. square-topped waves, saw-tooth voltages etc. The input voltage is applied between terminal O and either terminal E_1 or terminal E_2 ; the sensitivity at E_1 being $0.14 V_{rms}$ per centimetre picture height and at E_2 , $2.8 V_{rms}$.

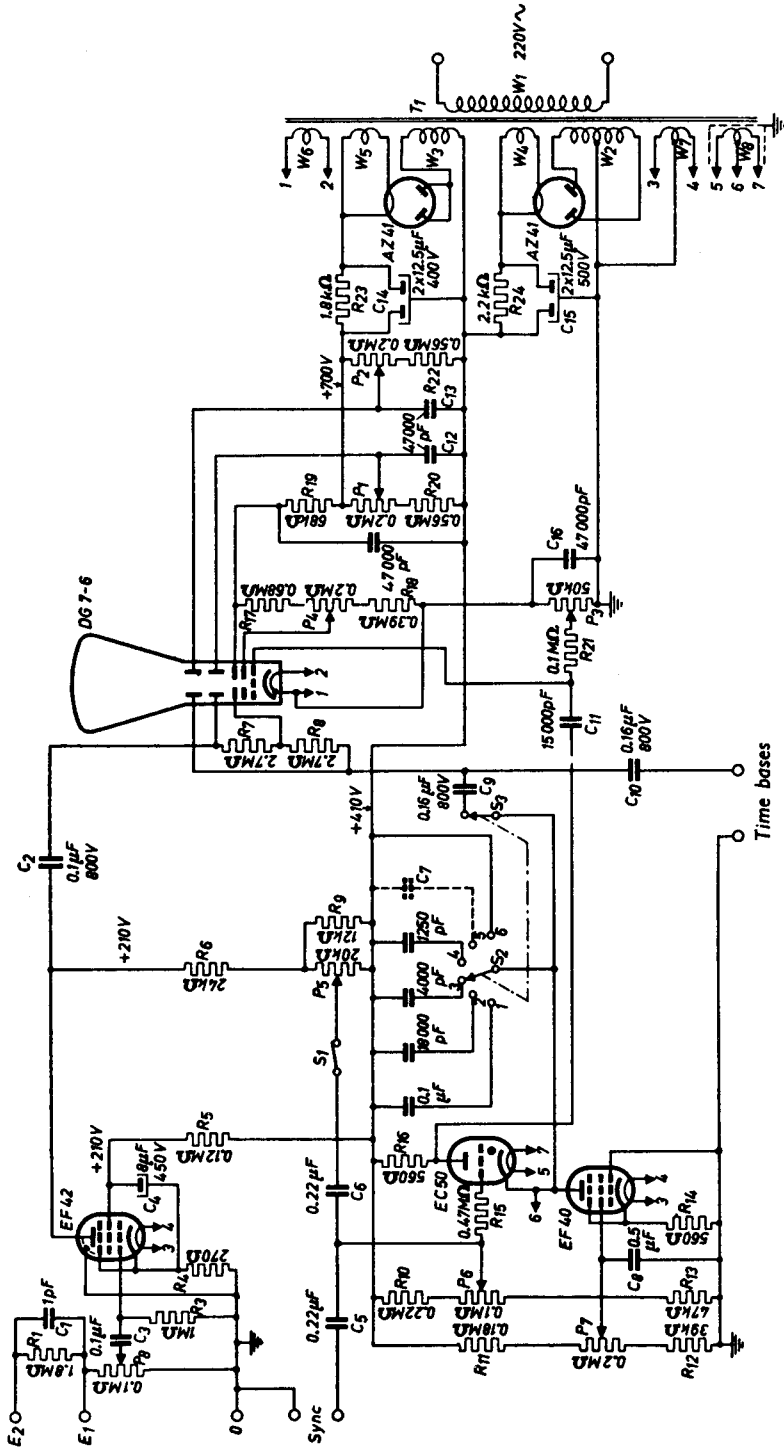


Fig. 9-4. Complete circuit of a simple cathode-ray oscilloscope.

Potentiometer P_8 serves for controlling the output of the amplifier. The amplified voltage is applied to the vertical deflection plates of the cathode-ray tube via capacitors C_2 and C_{12} . A part of this voltage is taken from potentiometer P_5 and is applied to the control grid of the gas-filled triode EC 50 via S_1 and C_6 , thus effecting synchronisation of the time-base oscillations. The time-base voltage is generated by an oscillator consisting of the gas-filled triode EC 50 and the pentode EF 40.

To explain the operation of this oscillator it should first be assumed that capacitor C_7 (4000 pF), which is connected to the circuit by switch S_2 (in position 3), is discharged. The cathode of the EC 50 is now at the same potential as the anode, and the control grid is at a negative potential with respect to the cathode via P_6 . Capacitor C_7 now charges slowly through the EF 40 tube and the voltage across the capacitor increases linearly. The potential of the cathode meanwhile falls at the same rate, and the grid becomes more positive until finally ignition of the gas-filled triode occurs and C_7 rapidly discharges through the EC 50 and resistor R_{16} . After the discharge the tube again becomes non-conducting, and the cycle repeats. The voltage developed across C_7 during the cycle is of saw-tooth waveform, as depicted in fig. 9-5, and this voltage is applied to the pair of plates for horizontal deflection, via the capacitors C_9 and C_{13} .

The frequency of the saw-tooth voltage can be approximately adjusted by connecting different capacitances C_7 by means of switch S_2 . Fine adjustment is obtained by varying the screen-grid voltage of the EF 40 tube by means of potentiometer P_7 . The capacitance switched in by position 5 of S_2 consists of the capacitance between the heater winding W_8 on the power transformer and the core, and also the stray capacitance of the circuit.

In position 6, the saw-tooth voltage generator is switched off, and switch S_3 is simultaneously opened, so that an external voltage for the time base can be applied at the terminals marked "time base".

As previously mentioned, synchronisation can be effected by applying a part of the voltage to be measured to the control grid of the EC 50 tube. Synchronisation can also be achieved by an alternating voltage from an external source, applied to the terminals marked "Sync". For this arrangement switch S_1 must be opened.

The amplitude of the saw-tooth voltage, and hence the width of the picture appearing on the luminescent screen, can be varied by adjusting the grid voltage of the gas triode EC 50 by means of potentiometer P_6 . The value of the grid voltage determines the voltage between cathode and anode at which ignition occurs. The range of adjustment of the time-base voltage which can be obtained in this way is from 170 to 300 volts, corresponding to variations in picture width between 3.5 and 6 cm.

During the discharge of capacitor C_7 a heavy current surge occurs through the gas triode. The anode resistor R_{16} limits the value of this current, so that it does not exceed the permitted maximum value of 750 mA. At the same instant

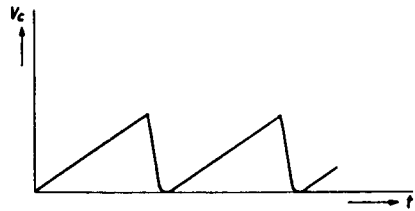


Fig. 9-5. Saw-tooth voltage across capacitor C_7 of fig. 9-4.

a voltage peak is developed across R_{16} , its polarity being such that the anode of the gas triode becomes negative with respect to the positive 410 volt supply.

This negative pulse is applied to the grid of the DG 7-6 cathode-ray tube via C_{11} , and suppresses the beam during the flyback. The grid of the cathode-ray tube also receives a permanent negative bias which is taken from the potentiometer P_3 . By varying this bias the brightness of the picture may be adjusted. The positive potential applied to the focusing electrode via potentiometer P_4 can also be varied, for adjusting the sharpness of the image.

The power supply is obtained from two rectifiers, equipped with rectifying

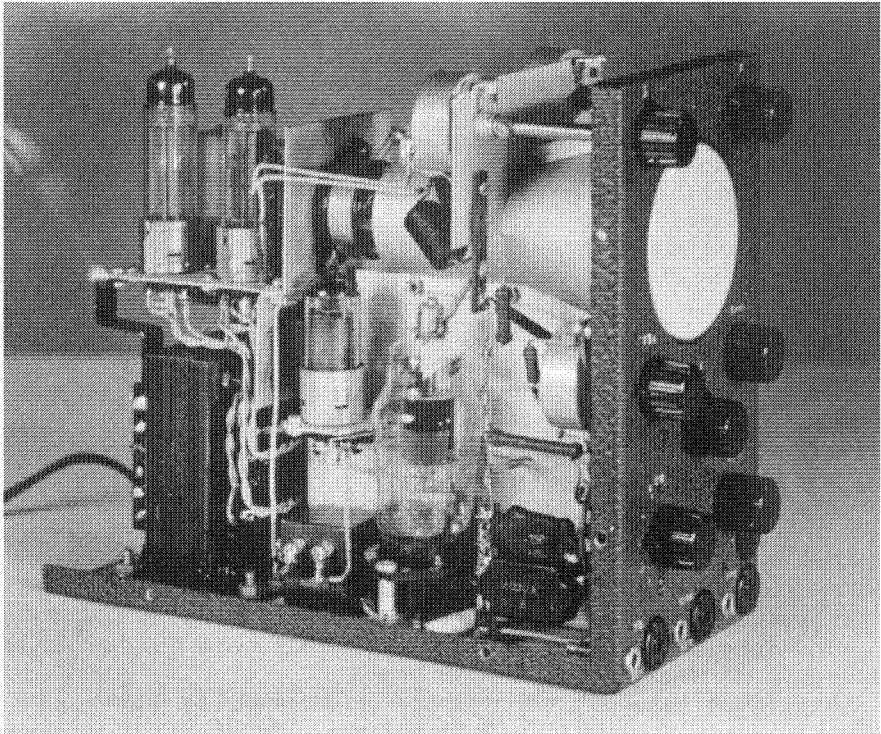


Fig. 9-6. Photograph of the cathode-ray oscilloscope according to fig. 9-4.

tubes AZ 41, one producing 410 volts, the other 300 volts. The 410 volt supply feeds the pre-amplifier stage and the time-base generator, and the two rectifiers connected in series produce a voltage of approximately 700 volts for the anode of the cathode-ray tube. The resistors R_{16} , R_{17} , P_4 , R_{18} and P_3 form a voltage divider which supplies the various auxiliary voltages and the anode voltage. Potentiometers P_1 and P_2 serve for vertical and horizontal adjustment of the position of the picture on the screen.

The complete equipment is illustrated in fig. 9-6. Both rectifying tubes, the gas triode and the charging tube are visible, but the pre-amplifier tube is on the other side of the chassis.

When operating a cathode-ray tube care should be taken that the light spot is never allowed to remain stationary or to move only slowly, unless the beam

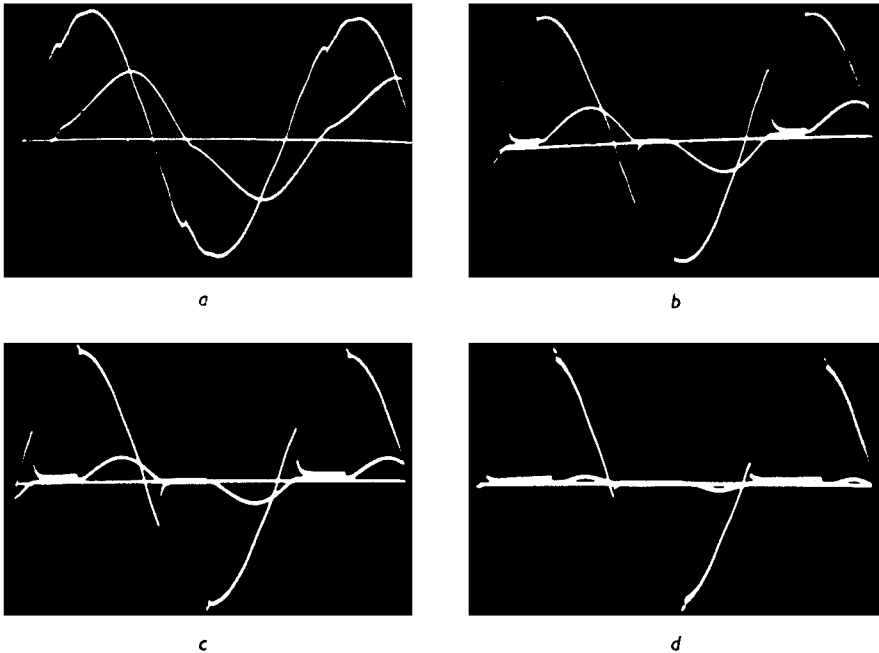


Fig. 9-7. Voltages and currents at an inductive load for two inverse-parallel connected thyratrons as functions of the firing angle (a) 1200 mA, (b) 600 mA, (c) 300 mA, (d) 75 mA (r.m.s. values).

current is first very greatly reduced. Otherwise the screen will be damaged. It is recommended that the time-base generator be kept operating if no voltage is applied to the plates for vertical deflection, so that the spot will not be stationary but will trace a horizontal line.

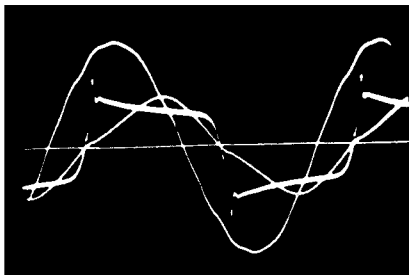


Fig. 9-8. Voltage and current of a fluorescent lamp connected in series with an inductance.

Ambient light falling on the screen reduces the picture contrast. If it is not convenient to shield the screen from stray light, it is recommended to use a filter of a colour similar to that of the light emitted by the screen. Some types of screen fluoresce when ultra-violet light falls upon them. If this occurs a suitable filter should be provided.

To avoid interference by magnetic fields it will generally be necessary to fit the cathode-ray tube with a screen of high-permeability metal, adapted to the shape of the tube. Such a screen can be seen in fig. 9-6.

A good example of the use of the cathode-ray oscilloscope is shown in fig. 9-7. Here, the voltages and currents for two inverse-parallel connected thyratrons operating with an inductive load, are plotted as functions of the firing angle. It may be of interest to compare these photographs with the drawings reproduced in fig. 5-9. Another example is given in fig. 9-8 showing the waveform of current and voltage of a fluorescent lamp connected in series with a choke, which may be compared with fig. 14-5.

PART II

Electronic Devices for Industrial Purposes

10. ELECTRONIC RELAYS

A problem which is constantly arising in one form or another is that of performing automatically some operation such as switching motors, lighting circuits and so forth on or off upon the receipt of a very feeble control signal. A most satisfactory solution is to use electronic relays, with which quite large loads can be switched by means of comparatively small signals such as those which are supplied, for example, by photocells. Conventional photo-electric relays have already been described in Chapters 7 and 8. Those now to be described incorporate special features to meet particular requirements in industrial applications.

Simple Photo-Electric Relay

The circuit shown in fig. 10-1, for example, is for a light-controlled relay for switching on such loads as room lighting, illuminated advertising signs or aircraft warning beacons at dusk and switching them off again at dawn. This very simple device is fed from the a.c. supply mains via the small transformer *Tr*, winding *a* of which feeds the photocell *F* and winding *b* the anode circuit of thyatron *T* (PL2D21).

The circuit is operative only during alternate half-cycles during which the right-hand ends of windings *a* and *b* are positive. So long as the photocell is illuminated, the current through the cell produces a voltage drop across resistor *R*₁ with polarity as shown in the diagram, thus applying a negative bias to the grid of the thyatron and preventing the tube from firing. In the evening, however, when the amount of light falling upon the photocell is greatly reduced and the photocell current is therefore small, the bias applied to the thyatron is correspondingly less. The thyatron therefore fires, causing current to flow through the heater of the indirectly-heated rectifying tube *V*. After a short delay the

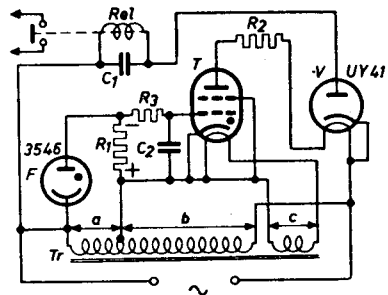


Fig. 10-1. Light-controlled relay circuit.

cathode of V attains the emitting temperature, and the tube passes current, thus energising relay Rel , which closes the load circuit. The short time delay corresponding to the warming-up time of rectifier tube V ensures that the circuit is operated only by the normal gradual alteration of the illumination. Temporary

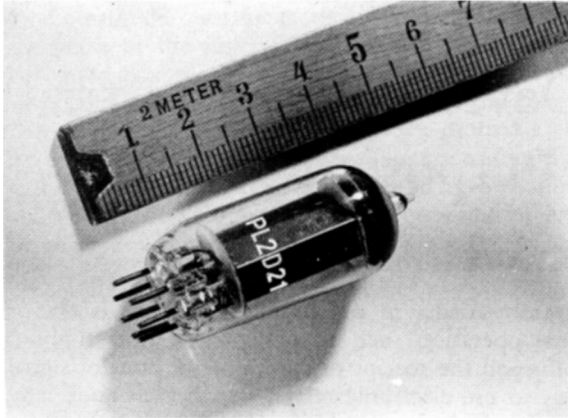


Fig. 10-2. PL2D21 tetrode thyatron.

fluctuations such as might be produced by aircraft during the daytime or by lightning at night cannot cause the relay to operate.

The thyatron used in this circuit is the miniature inert-gas-filled tetrode thyatron PL2D21 illustrated in fig. 10-2. It is particularly suitable for this class of application because its control characteristic is not temperature-dependent. The maximum average anode current rating of this tube is 100 milliamperes so that this tube can control the heater

current (100 milliamperes) of the UY 41 rectifier without using an additional relay. It will be observed that this thyatron is employed in several other circuits described in this book.

Selective Photo-Electric Relay

In some applications it is necessary to ensure that a photo-electric relay shall not be operated by stray light but only by light from one particular source. This can, of course, be arranged by using a colour filter so that the photocell receives only monochromatic light of a wavelength which does not occur in the spectrum of the stray light. This solution, however, often presents difficulties

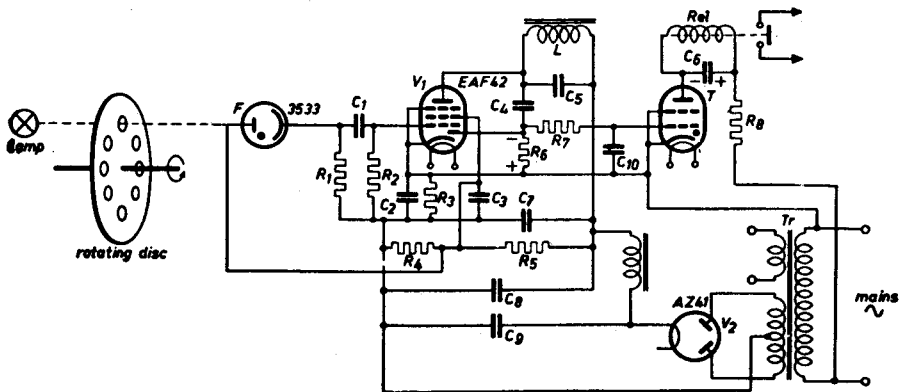


Fig. 10-3. Light-controlled relay using intermittent illumination.

in practical application, and a method based upon the use of intermittent light is frequently adopted.

A suitable circuit is given in fig. 10-3. An apertured disc is caused to rotate in front of the lamp so that the light beam is interrupted at a definite frequency. The current flowing through the photocell is therefore intermittent, and can be considered as consisting of a d.c. component and an a.c. component. The alternating component of the voltage appearing across the resistor R_1 is applied via capacitor C_1 to the control grid of the pentode amplifying tube V_1 , the direct component having no influence on this tube. The anode circuit of V_1 contains an oscillatory circuit consisting of the inductor L and capacitor C_5 , tuned to the frequency of the alternating voltage applied to the grid. Since this tuned circuit represents a very high resistance for this frequency, and since, as was explained in Chapter 1, the gain in a pentode amplifier is proportional to the resistance of the external load, this amplifier stage produces a large output voltage, which appears across the terminals of C_5 . This alternating voltage is applied via coupling capacitor C_4 to the diode incorporated in the amplifying tube V_1 and is there rectified so that a direct current flows in the diode load R_6 , producing a voltage drop with the polarity indicated in the circuit diagram. The negative end of R_6 is connected via R_7 to the control grid of thyatron T , type PL2D21, which thus receives a large negative bias and is prevented from firing. If the intermittent light beam falling upon the photocell is interrupted, the voltage across R_6 collapses, thyatron T ignites, and relay Rel is energised.

It will be seen that the operation of this circuit is controlled only by the intermittent light source. Stray light cannot cause the relay to operate, and even an intermittent beam of some other frequency will not affect it since the resistance of the tuned anode circuit (and hence the gain of the amplifier) is very small at other frequencies.

It will be seen that the relay coil Rel in the anode circuit of thyatron T is by-passed by a capacitor C_6 . This capacitor is charged during the positive half-cycles of the anode voltage with the polarity indicated in fig. 10-3, and discharges through the relay winding during the negative half-cycles. It thus maintains the current flow through the winding and prevents fluttering of the relay. The value of this capacitor must be carefully chosen, so that it is almost entirely discharged during the negative half-cycles. If it is too large, so that it does not discharge completely, the residual voltage across the capacitor will oppose the next positive half-cycle and may be sufficient to reduce the anode voltage of the thyatron to a value which does not allow the thyatron to fire. This condition will continue until the voltage across C_6 has dropped sufficiently for firing to occur. This irregular firing and extinguishing of the thyatron renders normal functioning of the relay impossible. As a general guide, the value of C_6 should not exceed from 1 to 2 μF . Alternatively, an a.c. relay can be used in place of a d.c. relay, and the by-pass capacitor can then be omitted; but this arrangement is not recommended because the d.c. component of the anode current would impose an excessive load on the a.c. relay.

Relay with Pre-Amplifier

Fig. 10-4 shows the circuit of a photo-electric relay with a pre-amplifier equipped with a modern high-vacuum photocell. When the latter is illuminated,

the negative control-grid voltage of the pentode E 80 F *) decreases, so that the output voltage of the bridge circuit which is formed by the tube V and

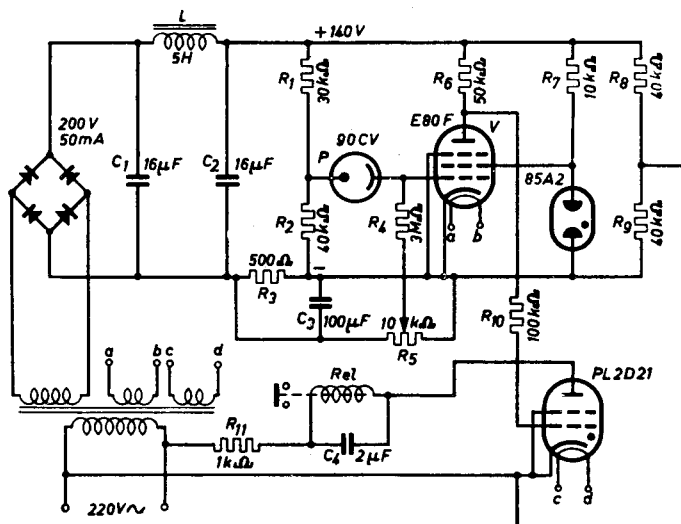


Fig. 10-4. Circuit of a photo-electric relay with pre-amplifier.

the resistors R_6 , R_8 and R_9 becomes negative, thus preventing the thyatron PL2D21 from igniting. However, as soon as the beam of light directed on the photocell is interrupted, the thyatron will ignite.

The screen-grid voltage of the pre-amplifying tube V is stabilised at 85 volts by the voltage reference tube 85 A2, which renders the operation of the circuit practically independent of mains voltage fluctuations.

Fig. 10-5 shows the construction of such a photo-electric relay.

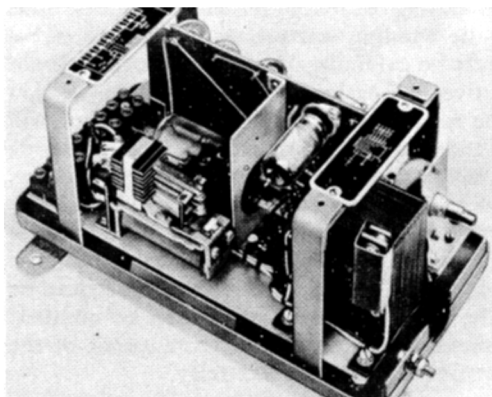


Fig. 10-5. Example of a photo-electric relay.

*) The pentode E 80 F is a special-quality long-life tube, similar to the double triode E 80 CC and the output pentode E 80 L, and is highly resistant to shock and vibration.

Photo-Electric Door Control

In factories and other industrial premises it is often necessary to arrange for the automatic opening and closing of doors to permit the transport of materials and so forth. Such a system should comply with the following conditions:

- (1) The doors should be locked when closed.
- (2) Opening of the doors should be initiated automatically by the approach of a person (or an object such as a truck) from either direction.
- (3) If several vehicles or groups of individuals are passing through, the door should remain open as long as necessary.
- (4) The operation of closing should also be initiated automatically.
- (5) The closing mechanism should be so designed that closing can be stopped easily by a single person in order to avoid accidents.
- (6) The complete system must be simple and reliable and be ready for immediate operation at any time; when it is switched off it should be possible to open or shut the door in the normal way.
- (7) The equipment should be inexpensive to install and operate.

A device which fulfils all the above conditions is illustrated in fig. 10-6 and is described below.

Two photocells F_1 and F_2 are connected in parallel and are mounted one on each side of the door so that the light beams from two lamps L_1 and L_2 directed on to F_1 and F_2 respectively pass across the passages to the doorway. If both photocells are illuminated their combined currents, flowing through resistor R_2 , produce a voltage drop which, applied to the grid of thyatron T , prevents it from firing. If one of the light beams is interrupted by an approaching person or object, the current through R_2 and thus the grid voltage of the thyatron PL2D21 is reduced, ignition occurs, and relay Rel is energised to close the circuit of the door-opening mechanism.

This mechanism is similar to that often used for the entrance doors of blocks of flats — a spring-loaded bolt pulls back the snap bolt by means of a short chain, thus unlocking the door. It differs from the normal construction, however, in that it incorporates an additional contact which is operated by a cam. When this contact is closed a pneumatic or motor-operated door-opening mechanism is set in motion and effects the actual opening of the door. Such devices are commonly employed in buses, trams and electric trains and need not be described in detail here. The door continues to open until a rod, hinged at the door-post, meets the bolt of the door-opener, pressing it back to its former position, at the same time opening the subsidiary contact and stopping the door-opening mechanism. The door closing mechanism is pneumatic and of conventional design.

It is essential to the satisfactory operation of the complete installation that it takes a shorter time for the door to open than for it to close. Obviously the door remains open so long as one of the light beams is interrupted, and a door which is in process of closing will immediately start to open again if one of the light beams is interrupted even for a short time.

If the main switch S is opened the automatic system is put out of action and the door can then be opened or closed in the normal way.

The circuit is shown twice in fig. 10-6, once in the open and once in the closed position.

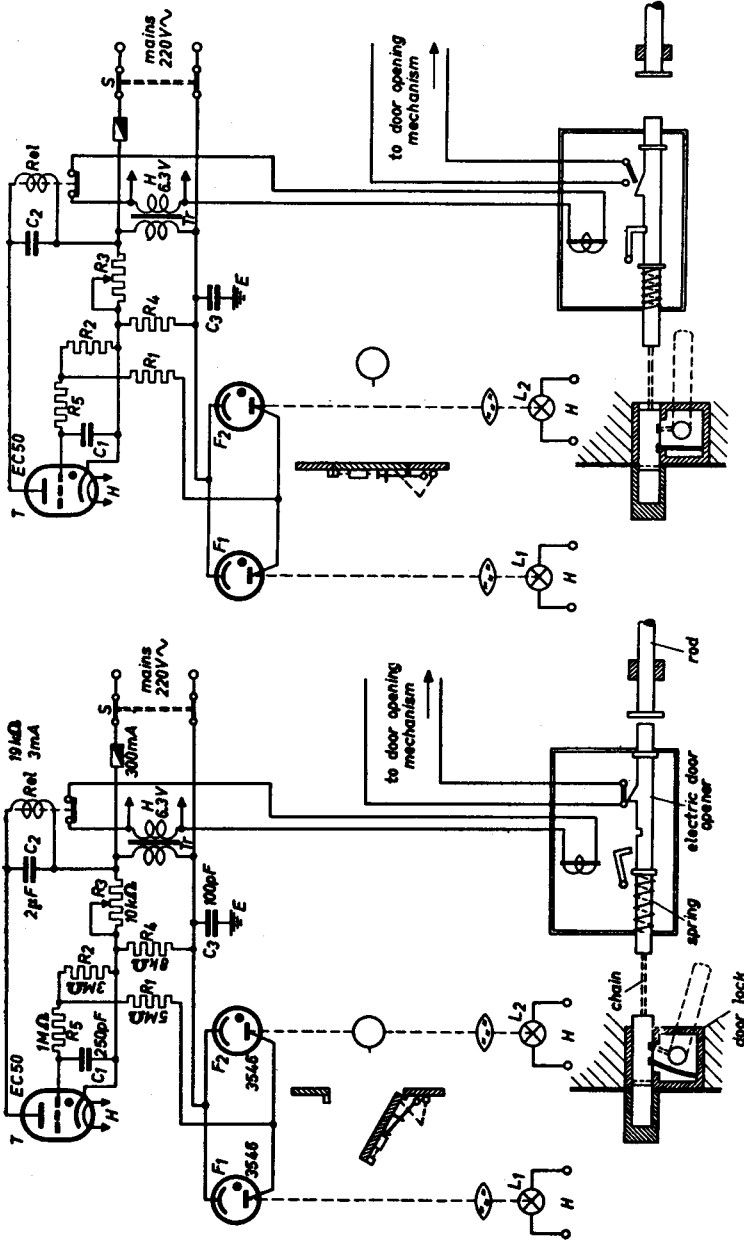


Fig. 10-6. Electronic door-opening device.

The gas-triode used is Type EC 50 which can be expected to have a long life in this class of circuit. The photocell is Type 3546. The type of lamp used to supply the light beams depends upon the distance between the lamps and the photocells, and also upon the optical system employed. Normally 6.3 volt lamps suffice, and these can be fed from the heater transformer for the thyatron. Potentiometer R_3 is fitted as a preset adjustment of the sensitivity of the photo-electric circuit.

Automatic Control of Oil-fired Boilers

A complete automatic control system for an oil-fired boiler must ensure that:

- (1) A good spark is available for firing the mixture.
- (2) Heating ceases automatically as soon as the desired temperature is reached.
- (3) Risk of over-heating the boiler is eliminated.
- (4) If any fault should occur the installation is immediately shut down and an alarm signal is given.

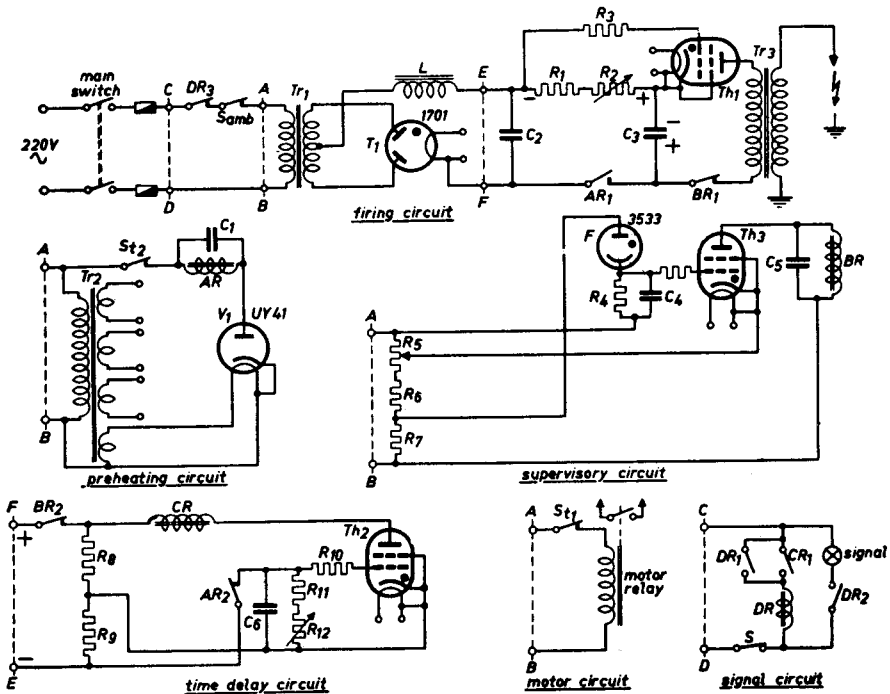


Fig. 10-7. Electronic control of oil furnaces.

Depending upon the particular needs of the installation, the electronic control system will consist of one or of several circuits. The system illustrated in fig. 10-7, for example, comprises the following six circuits:

- | | |
|--------------------------|------------------------------|
| (1) Firing circuit. | (4) Time delay circuit. |
| (2) Pre-heating circuit. | (5) Motor circuit. |
| (3) Supervisory circuit. | (6) Signal or alarm circuit. |

When the main switch is closed, the main transformer Tr_1 and the heater transformer Tr_2 are connected to the supply, and the motor relay is energised via contact St_1 of the boiler thermostat. The motor therefore starts and supplies air under pressure to the nozzle. After a few seconds the cathode of the rectifying tube V_1 will have warmed up, and the tube will pass current, energising relay AR to close contact AR_1 thus allowing capacitor C_3 to charge with the polarity shown in the diagram. The charging current of C_3 produces a voltage drop across resistors R_1 and R_2 which prevents the PL2D21 thyatron Th_1 from firing.

When, however, the capacitor is fully charged, the voltage drop across R_1 and R_2 collapses, and Th_1 ignites, discharging C_3 through the primary winding of the ignition transformer Tr_3 . Because of the inductance of the transformer winding, the discharge current will be maintained for sufficient time for C_3 to be partially charged with reverse polarity. The anode voltage of Th_1 therefore falls momentarily to a value less than the arc voltage value, and the tube ceases to conduct. Capacitor C_3 now charges again, and the cycle repeats.

At each discharge of C_3 a current surge flows through the primary winding of Tr_3 producing a high voltage pulse in the secondary circuit, which causes the spark for igniting the flame. The frequency of successive sparks can be adjusted by resistor R_2 .

Simultaneously with the operation of the ignition circuit, contact AR_2 is opened, thereby disconnecting a negative voltage from capacitor C_6 and from the control grid of thyatron Th_2 . C_6 can now discharge through R_{11} , R_{12} , the time of the discharge being adjustable by means of R_{12} . Assuming that the oil-vapour ignites during the period of this discharge, photocell F is illuminated by the flame, permitting thyatron Th_3 to fire, so that relay BR is energised, opening contact BR_1 to interrupt the firing circuit. At the same time contact BR_2 is opened, putting the time delay circuit out of operation. The installation is now operating.

If the temperature in the boiler reaches a pre-set temperature, boiler thermostats come into operation and, by opening contacts S_{11} and S_{12} , interrupt the motor circuit and the pre-heating circuit. The oil burner is thus extinguished, and since the photocell is then not illuminated, thyatron Th_3 becomes non-conductive and relay BR drops out. Contacts BR_1 and BR_2 therefore close and switch on the time delay circuit, but the firing circuit remains switched off because contact AR_1 is still open.

When the water temperature has fallen below the control value, the boiler thermostats close contacts S_{11} and S_{12} , the motor runs, the ignition circuit operates, and the time interval determined by $C_6 \cdot (R_{11} + R_{12})$ of the time delay circuit expires — in other words the installation starts up again as previously described.

When the ambient temperature reaches the value corresponding to the setting of the room thermostat, contact S_{amb} opens, switching off the complete installation which will remain shut down until the room temperature falls to the lower limit, whereupon the starting-up sequence recommences.

In the event of any irregularity or failure, such as, for example, failure of the ignition circuit, or failure of the gas mixture to ignite for any other reason, the photocell is not illuminated so that the anode circuit of thyatron Th_2 is not interrupted. After the expiration of the time delay interval therefore, i.e. when capacitor C_6 is almost completely discharged, Th_2 ignites and energises

relay CR . Contact CR_1 energises the self-holding relay DR to operate the alarm signal and contact DR_3 puts the whole installation out of action. The signal remains switched on until the circuit is interrupted manually by means of switch S or until the main switch is opened.

Should trouble occur while the installation is in operation, and the furnace flame is extinguished, contact BR_2 is closed by the supervisory circuit. Since capacitor C_6 is already discharged, the signalling circuit immediately operates.

Electronic Smoke Detector

A further example of a photo-electric relay, which may be used for such applications as smoke detection, is illustrated in fig. 10-8. Here, two photocells F_1 and F_2 are connected in a bridge circuit. Potentiometer R_6 is so adjusted that the high-vacuum amplifying tube V_1 receives only a small negative grid bias when the two photocells are equally illuminated. The anode current thus flowing through V_1 develops a voltage drop across resistor R_2 with the polarity indicated in the diagram. Thyatron T (PL_2D_{21}) is therefore cut off, and relay Rel is not energised.

If, however, owing to the presence of smoke, the illumination on photocell F_1 is reduced, the current through F_1 decreases, i.e. the effective resistance of the photocell increases. In consequence the potential at the control grid of V_1 becomes more negative, its anode current is reduced and with it the voltage drop across R_2 . Thyatron T would therefore ignite — after a 90 degree phase delay — but would pass full current only if its negative grid voltage is further reduced. Operation of the relay by reduced current might, however, cause fluttering, and to avoid this an alternating voltage is also applied to the thyatron grid. This

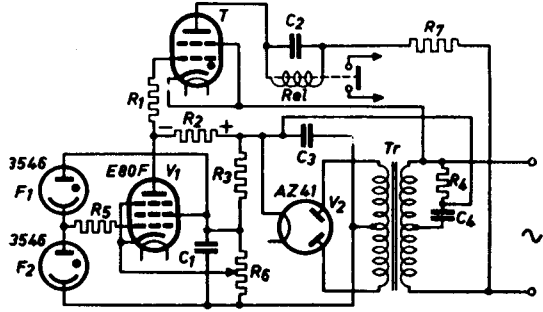


Fig. 10-8. Photoelectric smoke detecting circuit.

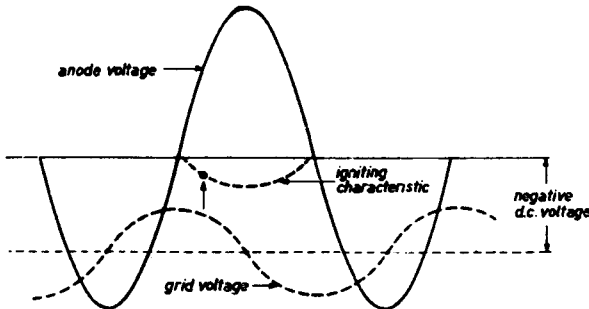


Fig. 10-9. Grid and anode voltage of tube T in fig. 10.6.

voltage is advanced about 90 degrees in phase with respect to the anode voltage of the thyatron as shown in fig. 10-9. From this figure it is seen that the ignition characteristic is intersected at the point indicated by the arrow when the direct negative grid voltage decreases. The thyatron therefore fires at only a small firing angle, and full current is applied to the relay coil.

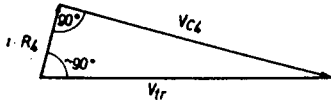


Fig. 10-10. Vector diagram of phase network included in fig. 10-6.

The alternating grid voltage for the thyatron is taken from resistor R_4 which is small compared with the reactance of capacitor C_4 . As is clearly shown in fig. 10-10, the voltage across R_4 is advanced by very nearly 90 degrees with respect to the alternating anode voltage, i.e. the voltage across the transformer primary.

Automatic Examination of Food Cans

The internal surface of tins used for packing preserved food is often lacquered in order to prevent corrosion of the metal by the acidity of the contents. It is necessary to examine each container before filling, and to reject any in which the lacquer is perforated. This can be done by the simple but reliable electronic device illustrated in fig. 10-11.

The tins, travelling on a conveyor belt, are filled with water or a dilute solution of common salt. The electrode E , which is connected to the control grid of thyatron T , is dipped into each container in turn. Normally the thyatron receives a negative grid bias due to the voltage drop appearing across resistor R_3 , and is thus prevented from firing. If, however, the lacquer coating of the tin is defective, a small current flows through R_2 , producing a voltage drop with the polarity shown. The thyatron thereupon ignites and energises relay Rel , setting into operation a mechanical device which removes the defective tin and at the same time momentarily opens switch S so that the thyatron is again extinguished and the circuit is ready to test the next tin.

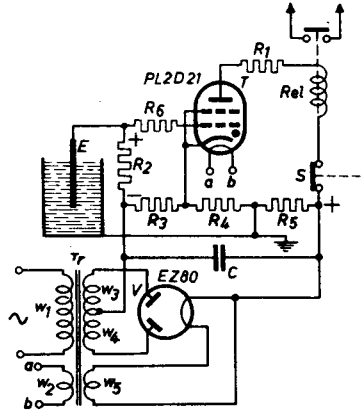


Fig. 10-11. Circuit for testing lacquered tins.

Automatic Reversing Control for Electric Motors

Finally, fig. 10-12 shows an electronic relay circuit for the automatic reversal of an electric motor within predetermined limits. Such a device can be used, for example, for reversing the raising and lowering movements of cranes; for automatically reversing machine tools or other mechanisms operating periodically between pre-set limits; and for automatic temperature control regulators.

The motor M is provided with two field windings and can therefore run either in one direction or the reverse direction according to whether relay Rel_1 is energised or not. The slider of the potentiometer R_{10} is coupled to the motor by gearing in such a way that, for example, in the case of a crane, raising or

lowering of the load results in movement of the potentiometer in the direction indicated in the diagram.

Potentiometers R_2 and R_4 can be set to adjust the lower and upper limits of travel respectively. Relay Rel_1 is so connected that when it is energised the motor runs in the direction to lower the load.

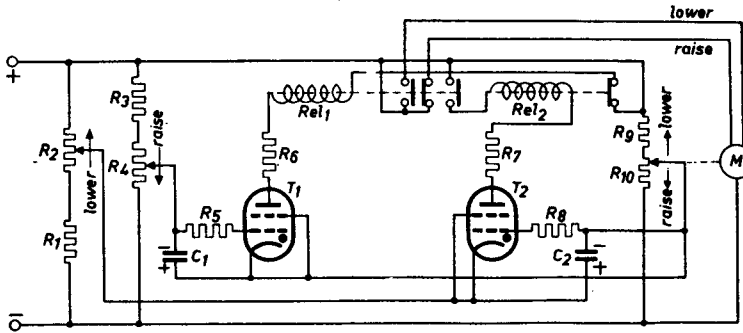


Fig. 10-12. Reversible motor controlling circuit.

In the diagram the relay is shown unenergised. The load is therefore raised, and the position of the slider of potentiometer R_{10} is moved in the corresponding direction so that the voltage difference between the sliders of R_4 and R_{10} appearing across capacitor C_1 is reduced. This voltage is applied to the control grid of thyatron T_1 as negative bias, preventing T_1 from firing. The thyatron will fire, however, when this voltage is reduced to a sufficiently small value. Relay Rel_1 will then be energised and will reverse the motor.

The relay also closes the anode circuit of thyatron T_2 . This tube does not ignite at this stage because the difference of potential between the sliders of R_2 and R_{10} charges capacitor C_2 with the polarity shown, thus applying a negative bias to the control grid of T_2 . This voltage decreases slowly, however, as the load is lowered, until T_2 fires and energises relay Rel_2 which interrupts the anode circuit of T_1 . Rel_1 thus falls out, causing the motor to reverse and also interrupting the anode circuit of T_2 . The original condition is again reached, and the process repeated.

11. ELECTRONIC COUNTING CIRCUITS

In mass production plants it is often required to count objects such as components or finished products as they pass a given point, say on a belt conveyor. By means of a photocell and electric relay an impulse can be produced for every piece scanned, but a device which will register the number of these impulses is then required. There are, of course, suitable mechanical counting devices of many kinds. For some applications, however, the counting speed of these devices may be too slow, in which case an electronic "reduction circuit" or "scaler" can be used whereby the counting rate may be increased up to many thousands of impulses per second. Again, in some cases the impulses to be counted may be too feeble to operate a mechanical counter, and here an entirely electronic counting circuit may be employed instead of a mechanical device.

„Scale-of-two” Circuit

A simple circuit which reduces the number of impulses registered by the relay in the ratio of 1 : 2 (“scale-of-two” circuit), thus doubling the speed of counting, is illustrated in fig. 11-1.

The photocell *F* is illuminated momentarily at regular intervals, for example by a light beam reflected from the objects to be counted. The corresponding short positive voltage impulses appearing across resistor *R*₅ are applied to the control grids of the two thyratrons *T*₁ and *T*₂ via capacitors *C*₁ and *C*₂ respectively. As the ignition voltages of these two tubes will inevitably differ slightly, one of the tubes, say *T*₁, will ignite first,

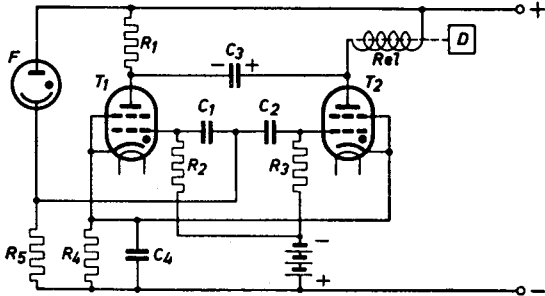


Fig. 11-1. Scale-of-two counting circuit.

and its anode current, flowing in the common cathode resistor *R*₄, will increase the negative grid bias of *T*₂, thus preventing this tube from firing.

Moreover, the anode potential of *T*₁ will now be equal to the arc voltage value and, until capacitor *C*₃ has charged with the polarity indicated, the same applies to the anode potential of *T*₂. When a second positive impulse occurs *T*₂ will ignite since its anode potential will have once more reached the value of the supply voltage as the result of *C*₃ becoming charged. *T*₂ having ignited, however, its anode potential immediately falls to the arc voltage value, and because of the charge on *C*₃, the anode potential of *T*₁ drops to a considerably lower value and *T*₁ is extinguished. Meanwhile, the anode current of *T*₂ flows in relay *Rel*, and the mechanical counting device *D* is operated.

Because *T*₁ is now non-conducting and *T*₂ is ignited, *C*₃ charges with reverse polarity so that when a third impulse is produced *T*₁ ignites and *T*₂ is extinguished since the voltage across *C*₃ brings the anode potential of *T*₂ below the arc voltage value. The fourth impulse, however, will again fire *T*₂ and extinguish *T*₁, at the same time operating the counting mechanism once more. It is thus seen that the counter *D* is operated only at every second impulse.

Ring Counting Circuit

One form of ring counting circuit has already been described in Chapter 8. A similar device, which is a development of the circuit described above, is shown in fig. 11-2.

Small inert gas-filled tetrode thyratrons, Philips Type PL 2D 21, are employed. In this type of tube the control characteristics can be shifted to the region of positive control grid voltages by applying a negative potential to the second grid, as shown in fig. 11-3. This is achieved for all the tubes in the circuit of fig. 11-2 since their cathodes are given a fixed positive potential by the voltage developed across resistor *R*₀. In order to ignite any tube, therefore, it is necessary to apply a positive potential of a predetermined value between its control grid and cathode.

The operation of this circuit can best be followed by assuming that tube *T*₁₀

is ignited so that a positive voltage drop appears across R_{19} and is applied to the control grid of thyatron T_1 but is not sufficiently great to cause that tube to ignite. Furthermore, it should be assumed that capacitor C_8 is charged with the polarity as shown.

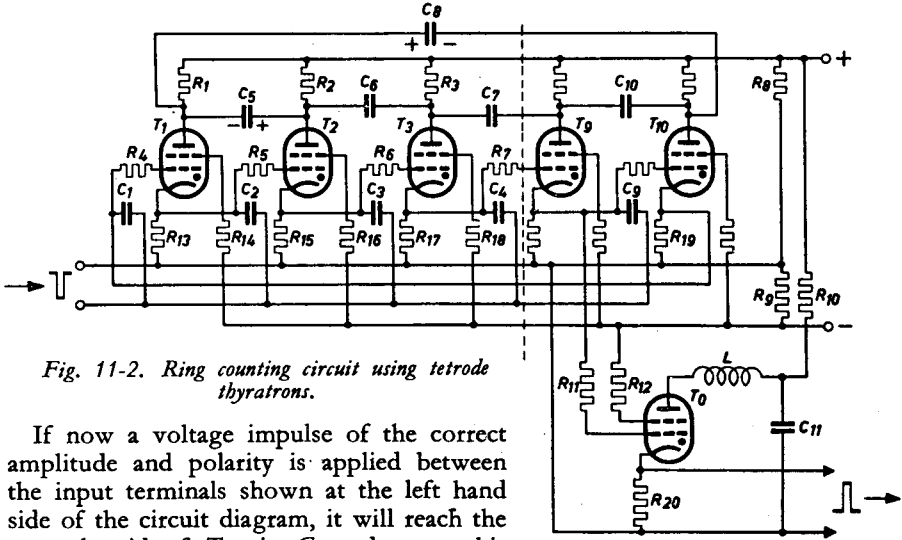


Fig. 11-2. Ring counting circuit using tetrode thyatrons.

If now a voltage impulse of the correct amplitude and polarity is applied between the input terminals shown at the left hand side of the circuit diagram, it will reach the control grid of T_1 via C_1 and cause this tube to ignite. This impulse of course also appears at the control grids of the other tubes, but they will remain extinguished because they have no positive grid bias. Simultaneously with the ignition of T_1 , T_{10} is extinguished because the voltage across C_8 causes the anode potential of T_{10} to drop momentarily below the arc voltage value.

Due to the anode current that flows through R_{13} , the control grid of T_2 now receives a positive bias and at the same time C_5 is charged with the polarity shown. When the next impulse appears, therefore, T_2 will fire and T_1 will be extinguished. In a similar way further impulses cause the ignition and extinction of the successive tubes until at length T_9 is ignited. The next impulse ignites T_{10} and T_0 . When T_0 ignites, C_{11} can discharge through the inductor L and resistor R_{20} .

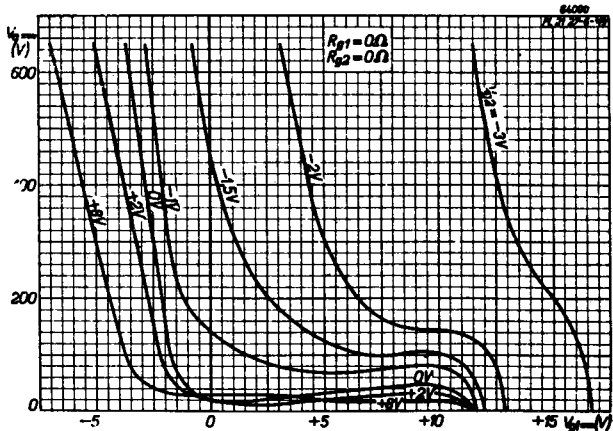


Fig. 11-3. Shifting the control characteristic of a PL2D21 tetrode thyatron.

As described in an earlier chapter, however, T_0 rapidly becomes non-conducting due to the effect of the inductance, so that only a momentary current surge flows through R_{20} , producing a voltage impulse which is applied to the counting ring of the second decade.

Further input impulses now operate the tubes of the first decade once more in succession until the 20th impulse fires T_0 again, producing the second impulse for the counting ring of the second decade.

The Eccles-Jordan ("Flip-Flop") Circuit

Another commonly used counting circuit is that shown in fig. 11-4. Instead of using thyratrons, double triodes are employed in the so-called Eccles-Jordan or "flip-flop" multivibrator circuit. The two electrode systems, A and B of the double triode, each has its own anode resistor R_1 and R_2 and there is also a common anode resistor R_3 . A common cathode resistor R_{10} is employed but the cathode circuit of section A has an additional resistor R_9 which can be short-circuited by switch S . A glow discharge lamp D , in series with a current limiting resistor R_8 , is connected between the anode and cathode of section A and serves as a signal lamp.

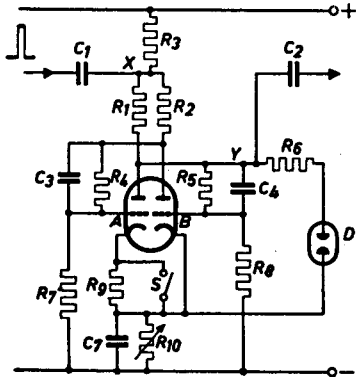


Fig. 11-4. Eccles-Jordan circuit ("flip-flop" circuit).

The circuit is symmetrical, but in practice these currents will differ slightly because of the inevitable tube tolerances. Imagine that the anode current of the section A is slightly higher than that of section B . The voltage drop across resistor R_1 will therefore be greater than that across resistor R_2 . Now R_1 , R_3 and R_8 constitute a voltage divider which, together with a voltage drop across cathode resistor R_{10} , determines the effective grid bias of section B of the double triode. The increased voltage drop across R_1 mentioned above therefore reduces the grid potential of section B , so that the anode current of this section decreases, thus reducing the voltage drop across the anode resistor R_2 . This in turn increases the grid potential of section A of the tube via the voltage divider composed of R_2 , R_4 and R_7 , and this causes a further increase in the anode current of section A . It is thus seen that this d.c. feedback results in an increase of the anode current of section A and a corresponding reduction in the anode current of section B until ultimately section B is completely cut off and section A is operating at full anode current.

In describing the operation of this circuit, switch S is assumed to be closed. Impulses delivered from a photocell with or without an amplifier stage, are applied via capacitor C_1 . These signals are assumed to have the square wave form shown in the diagram, that is to say, point X in the circuit becomes momentarily negative with respect to the positive high tension line.

It will be convenient first to consider the behaviour of the double triode when no pulses are applied. The anode currents of the two sections A and B should theoretically be equal since the

It will be clear that it is also possible to obtain the reverse effect where section *B* conducts and section *A* is cut off, but it will be recognised that these two conditions are the only stable conditions of the circuit. This mode of operation can be compared with that of a two-way tumbler switch and it is from this analogy that the name "flip-flop" is derived.

The condition in which section *A* of the double diode is cut off and section *B* is conducting can be obtained by opening switch *S* for a short time, thus reducing the grid potential of section *A*, causing the anode current to decrease

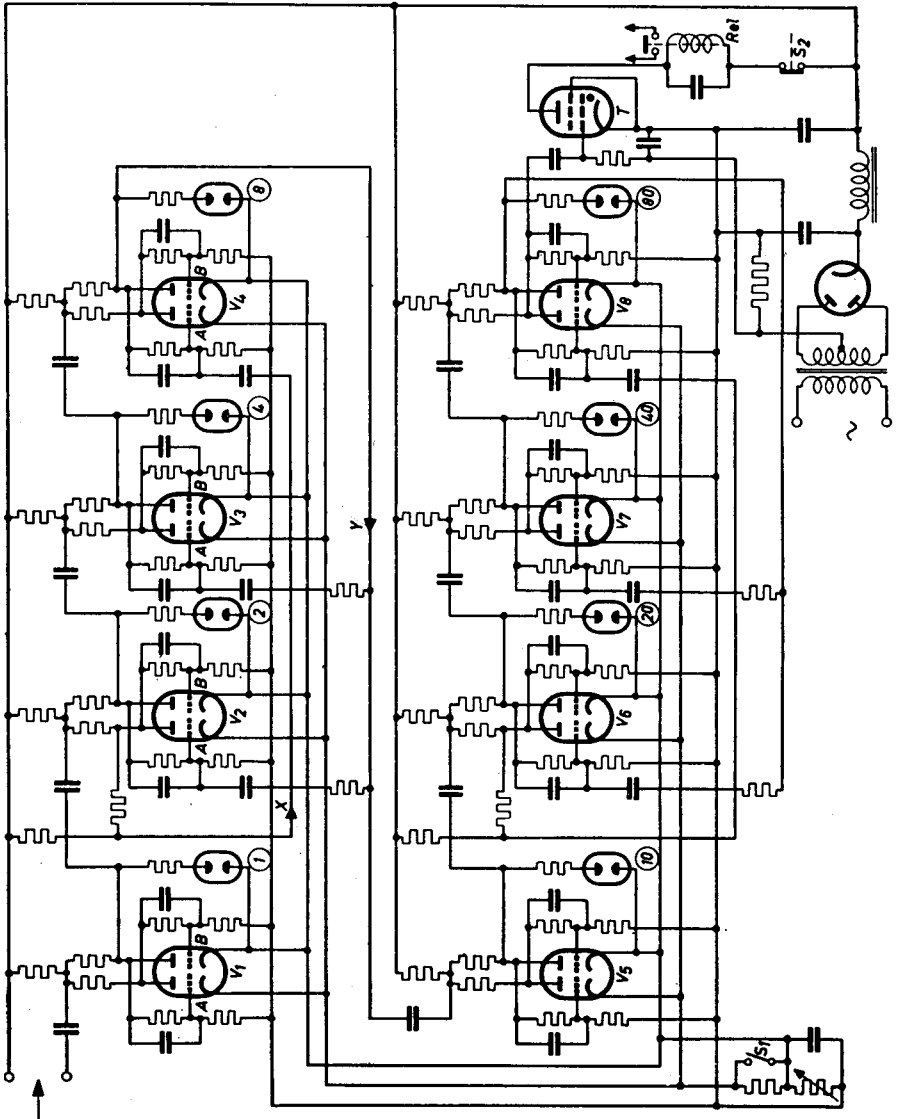


Fig. 11-5. Flip-flop counting circuit for two decades.

and with it the voltage drop across R_1 . This increases the grid potential of section B causing the anode current also to increase. The voltage drop across R_2 is therefore also increased and the grid potential of section A further reduced, this process being cumulative until section A is cut off. At this point the full supply voltage appears between the anode and cathode of section A and the lamp D connected across these electrodes lights up. If, however, section B is blocked, the voltage between the anode and cathode of section A which is now conducting is only small and the lamp does not light.

The operation of the circuit when the negative impulse is applied via C_1 can now be considered. Assuming that section A of the valve is cut off and system B is conducting, the negative grid potential of section B will be increased via the voltage divider R_1, R_5, R_8 and the anode current of that section will decrease so that the circuit changes over to the alternative stable operating condition in which section B is cut off and section A conducts.

A second negative impulse results in the circuit changing over to the original condition at which A is cut off and B conducts. If, however, a positive impulse is applied it will have no effect on the circuit beyond a momentary increase in the current flowing through whichever section of the tube happens to be conducting at the time.

Every time section A of the circuit is cut off a positive voltage impulse appears at the anode which lights the signal lamp D and can also be applied to another "flip-flop" stage via capacitor C_2 , while every time section B is cut off and A is made conducting the signal lamp is extinguished and a negative pulse appears at the anode.

Fig. 11-5 shows a counting circuit which is suitable for counting up to two decades, i.e. from 1 to 9 and from 10 to 90. Each decade requires four double triodes connected in the "flip-flop" circuit. The second decade chain includes a final stage which operates a relay.

The B section of each double triode is associated with its own glow discharge lamp and these lamps show the numbers 1, 2, 4, 8, 10, 20, 40 and 80. When the B section of a given tube is cut off, the associated signal lamp lights. When the counting system has to be put into operation, switch S_1 is opened for a short time thus ensuring that all the A sections of the tubes are cut off and all signal lamps are extinguished.

The negative impulses which are to be counted are now applied to the input terminals of the circuit, and their effect upon the first decade chain is illustrated diagrammatically in fig. 11-6. The first impulse switches over high-vacuum tube V_1 so that section A conducts and section B is cut off so that lamp 1 will light and a positive impulse is transmitted to tube V_2 which, however, has no effect on this tube. The second impulse switches T_2 back to its original condition so that section A is cut off and section B conducts. Lamp 1 is therefore extinguished, but a negative impulse is transferred to V_2 so that its section A conducts and section B is cut off causing lamp 2 to light. The successive stages are easily explained by reference to fig. 11-6 when it is remembered that only a negative pulse causes a stage to change over, a positive impulse having no action whatsoever.

The number of pulses counted is equal to the sum of the numerals illuminated on the signal lamps. The operation of the second decade chain is initiated by the tenth impulse. It will be observed from fig. 11-6 that after the ninth im-

pulse, section A of V_1 is conducting, section A of V_2 is non-conducting, section A of V_3 is also non-conducting and section A of V_4 is conducting. The tenth impulse switches V_1 so that section A is cut off and a negative impulse is transmitted to V_2 which is also switched. V_2 produces a positive impulse which has no appreciable effect on V_3 . A negative impulse is, however, taken from the anode of section A of V_2 and is applied via line X to the grid of section A of V_4 (see fig. 11-5). Because section A of V_4 has been conducting, V_4 is switched and produces a negative impulse which is applied via line Y to tubes V_2 , V_3 and V_5 . Tubes V_2 and V_3 are therefore switched so that their B sections become conducting and their respective lamps are extinguished and V_5 is switched so that its section B is cut off and lamp 10 is illuminated. Further negative impulses will operate the first decade chain once more until the twentieth impulse extinguishes lamp 10 and illuminates lamp 20 and so on.

When finally the hundredth impulse is applied, section A of V_8 produces a positive impulse which is applied to the final stage which includes a small inert gas filled thyratron T whose control grid is normally negatively biased. The positive impulse due to the hundredth impulse causes this tube to ignite and the relay connected in its anode circuit is energised.

If the counting device is used for such processes as bottling tablets or packing cigars, the relay can be used to interrupt the bottling or packing process and to substitute an empty pack for the one already filled. The anode circuit includes a switch S_2 which is operated automatically by the packing mechanism, thus extinguishing the thyratron and allowing the packing process to recommence. This counting device can, of course, be extended by adding further decade chains. Furthermore, the circuit can be used for counting at a point remote from the signalling position. Finally, by using several photocells in a suitable circuit the device can be made selective so that only objects of a certain shape or colour will be registered and counted. This application may be useful not only in industry but also for traffic control in streets, exhibitions and garages or at factory doors, etc.

Circuits with Decade Counter Tubes

In the electronic counting circuits discussed above the condition of the counter had to be indicated by means of special signal lamps or similar devices, in contrast

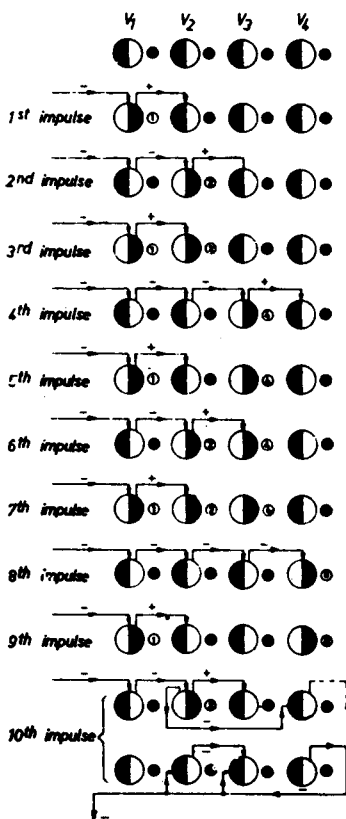


Fig. 11-6. Diagram showing the operation of the counting circuit of fig. 11-5. Black sections are non-conducting.

to apparatus in which a new type of counter tube, the E 1 T, is used (see fig. 11-7). This tube, by means of which an entire decade can be counted, itself indicates the number of counts carried out.

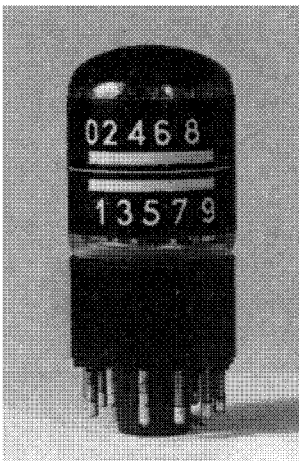


Fig. 11-7. Decade counter tube E 1 T.

The E 1 T decade counter tube is in fact a small cathode-ray tube, the ribbon-shaped electron beam of which can occupy 10 stable positions. In doing so, luminescent marks are formed at 10 different points on the glass envelope, which has a fluorescent screen; these marks are situated opposite the figures "0" to "9" which are indicated on a mask. At each pulse applied to the tube the beam advances one step, until finally position "9" is passed, and the beam returns to position "0". The tube then produces a pulse which is fed to the input of the next decade, as a result of which the latter advances one step.

Owing to its small dimensions (diameter of the envelope 35 mm, length 75 mm), the E 1 T counter tube is particularly suited for application in counters and electronic computers, in which it can also serve as a "memory". The circuits are simpler than those for the binary system, and the number of double triodes required is only 25 to 30% of that required for this system. By means of a single E 1 T and a single double triode E 90 CC (in a flip-flop circuit) per stage, it is possible to count up to 30 000 pulses per second.

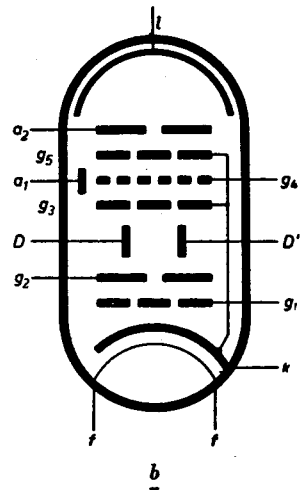
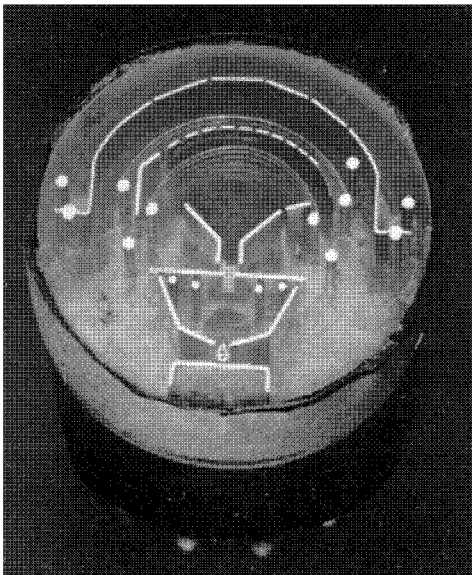


Fig. 11-8. Cross-section of the decade counter tube E 1 T (a) and diagrammatic representation of this tube (b).

Fig. 11-8a shows a cross section of the E 1 T, the customary symbol of which is shown in fig. 11-8b, all electrodes which are not connected externally being omitted. The electrons emitted by the indirectly heated rectangular cathode k are focused by the control grid g_1 , various focusing electrodes and the acceleration electrode g_2 so that a ribbon-shaped electron beam is obtained. The focusing electrodes, the two suppressor grids g_3 and g_5 and a screen are internally connected to the cathode, whilst an auxiliary anode is internally connected to the acceleration electrode. The fluorescent screen is provided with an electrically conductive coating, which is connected to the highest available potential, in order to prevent the formation of disturbing charges.

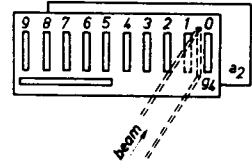


Fig. 11-9. Diagrammatic representation of the slotted electrode g_4 .

Fig. 11-9 represents schematically the slotted electrode g_4 ; attention is drawn to the horizontal slot at the left. Assume the anode voltage V_{a2} (and also the voltage of the right-hand deflecting plate D') to have such a value that the electron beam impinges on the slotted electrode g_4 to the right of the slot marked 0. Obviously no current will then flow to the anode a_2 , the entire beam being intercepted by g_4 . When the voltage at a_2 and D' is lowered, the beam will be deflected to the left, and the number of electrons flowing towards the anode a_2 will steadily increase. If the anode voltage is further reduced, a gradually increasing proportion of the beam is intercepted by the separation between the slots 0 and 1 and the anode

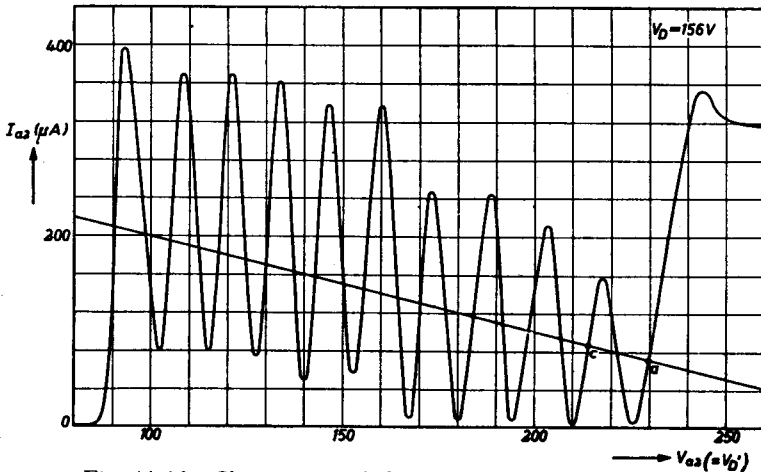


Fig. 11-10. Characteristic of the decade counter tube E 1 T.

current thus decreases again. Between the slots 4 and 5 the horizontal slot also becomes operative, so that the anode current no longer drops back to zero but fluctuates around a higher average value.

The actual characteristic is shown by fig. 11-10; the load line for a resistance of $1\text{ M}\Omega$ has also been plotted, in accordance with the circuit shown in fig. 11-11. In this circuit the anode a_2 and the deflecting plate D are connected to $+300\text{ V}$ via an external resistor of $1\text{ M}\Omega$, whereas g_4 is fed via a series resistor

of 47 kΩ, and D is connected to +156 V via a leak resistor. It will be assumed that the beam occupies position 0 , corresponding to the point of intersection a

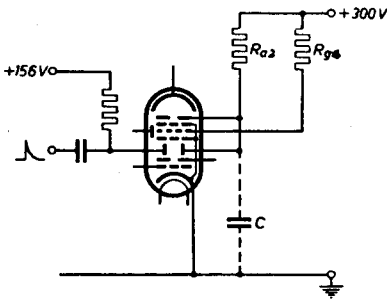


Fig. 11-11. Basic circuit of the decade counter tube.

(fig. 11-10) of the load line with the tube characteristic. If the voltage at D were gradually increased, the beam would be deflected to the left if the voltage at D' did not increase simultaneously, thereby counteracting the effect of the voltage increase at D . In fact, as the beam is deflected, it will be intercepted by g_4 , so that the beam current and the voltage drop across the external load decreases, which results in a rise of the voltage at D' .

Since the slope of the characteristic at a greatly exceeds that of the load line, the deflection to the left will be extremely small, in other words the position is stable.

This also applies, conversely, to a gradual decrease of the voltage at D , for in that case the proportion of the beam reaching a_2 increases. As a result of the large difference between the external and the internal resistance of the tube, the individual positions are also stable when the anode voltage is subject to slow fluctuations, provided these are not excessive.

When, however, a positive pulse with a steep front is applied to D , via the coupling capacitor, the behaviour of the tube will be entirely different. In that case the beam will be deflected to the left, but D' cannot follow the sudden voltage rise because the capacitance C (formed by the inter-electrode and wiring capacitances) must be discharged via R_{a2} . If the amplitude of the pulse is sufficiently large to shift the beam to the vicinity of the following stable position i , the beam will remain there and will not be returned to its original position by the trailing edge of the pulse, provided its slope is not too steep. In other words, the slope of the leading edge of the pulse must exceed the rate of change of the voltage across the capacitance C as determined by the time constant $C \cdot R_{a2}$, whereas the slope of the trailing edge must be smaller than this rate of change (fig. 11-12). The condition is therefore imposed that, for reaching a high counting speed, the value of C should be kept as small as possible, although this speed is mainly determined by the "reset time", that is the time the beam requires to return from position 9 to position 0 .

After nine pulses have been applied to D , the beam will be at the slot 9 . If a tenth pulse is now applied, this must ensure in the first place that the beam is reset from 9 to 0 , and secondly that a pulse is fed to the following decade stage, so that this is advanced one position. This is achieved in the following way. The beam is deflected by the tenth pulse so that it impinges on the reset anode a_1 . The current then suddenly flows from a_1 through R_5 (fig. 11-13), as a result of which a negative-going pulse is applied to the grid of left electrode system of the double triode E 90 CC. This tube forms part of a monostable multivibrator,

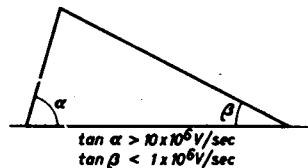


Fig. 11-12. Waveform of the pulse required for controlling the decade counter tube.

which is triggered by this pulse. As a result, a single powerful pulse is produced at the anode of the right-hand electrode system, after which the double

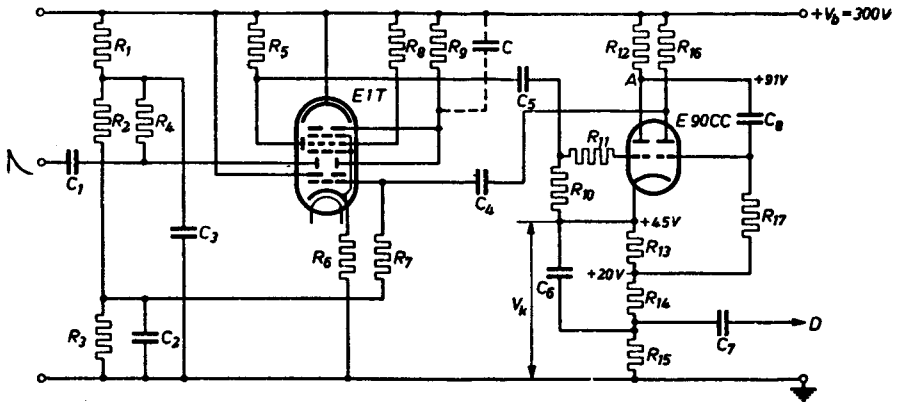


Fig. 11-13. Circuit diagram illustrating the resetting of the electron beam.

triode returns to its original, stable condition. This negative pulse is fed to g_1 of the counter tube of the first stage, thereby temporarily suppressing the beam. The potential at a_2 and D' then starts to rise to the highest value V_b according to the time constant $R_9 \cdot C$, so that the returning beam is deflected entirely to the right, hence to position θ . The duration of the pulse must therefore at least be equal to the time the voltage across C requires to rise, but on the other hand it should not be much longer than this time, because the counting speed would thereby be reduced.

A positive pulse is simultaneously produced across the non-bypassed cathode resistor R_{15} of the E 90 CC. This pulse is fed to the deflecting plate D of the following stage via C_7 , thus advancing this stage one position.

The first stage must be preceded by a pulse shaper, which also consists of a monostable multivibrator equipped with an E 90 CC. The circuit values of this multivibrator are so chosen that at the output a pulse of very short duration is obtained, and a counting speed of 30 000 pulses per second can be reached. A pulse of very short duration may be used, for in this case there is no preceding counter tube which has to be reset to position θ . Such a pulse shaper should, moreover, be preceded by a differentiating network to prevent the multivibrator from being operated once again by pulses the duration of which exceeds the time required by the multivibrator to return to its original position. Using this network, either positive or negative going pulses may be applied to the input. Their amplitude may range from approximately 20 to 50 volts; the applied pulses should, moreover, be roughly square-shaped. A sinusoidal input voltage can be converted into a square-wave voltage by means of a double limiter.

Fig. 11-14 shows the complete circuit of a two-stage counter with an input stage and a voltage divider (R_{41} , R_{42} and R_{43}), from which up to seven stages can be fed with the voltages $+11.9$ V and $+156$ V for g_1 and D respectively. The parts of the circuit separated by broken lines represent interchangeable

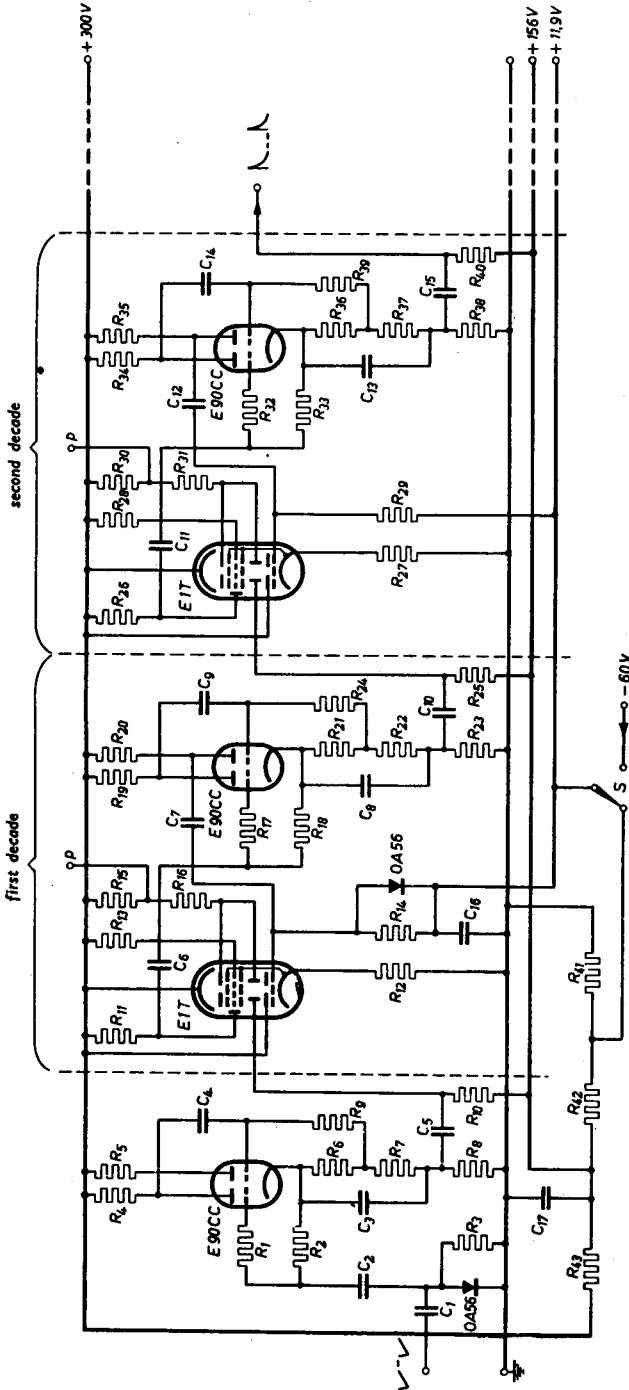


Fig. 11-14. Complete circuit of a two-stage decade counter.

COMPONENT VALUES

R_1	$= 5.6 \text{ k}\Omega \pm 10\%$	R_{11}	$= R_{14} = 4.7 \text{ k}\Omega \pm 2\%$	C_3	$= 82 \text{ pF} \pm 2\%$
R_2	$= 0.56 \text{ M}\Omega \pm 10\%$	R_{12}	$= R_{15} = 2.7 \text{ k}\Omega \pm 2\%$	C_4	$= 39 \text{ pF} \pm 5\%$
R_3	$= 39 \text{ k}\Omega \pm 5\%$	R_{13}	$= R_{16} = 1 \text{ k}\Omega \pm 1\%$	C_5	$= 0.0068 \mu\text{F} \pm 10\%$
R_4	$= 39 \text{ k}\Omega \pm 2\%$	R_{14}	$= R_{17} = 0.15 \text{ M}\Omega \pm 2\%$	C_6	$= 220 \mu\text{F} \pm 10\%$
R_5	$= 3.3 \text{ k}\Omega \pm 2\%$	R_{15}	$= R_{18} = 15 \text{ k}\Omega \pm 2\%$	C_7	$= 0.0068 \mu\text{F} \pm 10\%$
R_6	$= 4.7 \text{ k}\Omega \pm 2\%$	R_{16}	$= R_{19} = 5.6 \text{ k}\Omega \pm 1\%$	C_8	$= 68 \text{ pF} \pm 2\%$
R_7	$= 2.7 \text{ k}\Omega \pm 1\%$	R_{17}	$= R_{20} = 68 \text{ k}\Omega \pm 1\%$	C_9	$= 68 \text{ pF} \pm 2\%$
R_8	$= 1 \text{ k}\Omega \pm 1\%$	R_{18}	$= R_{21} = 68 \text{ k}\Omega \pm 1\%$	C_{10}	$= 680 \text{ pF} \pm 5\%$
R_9	$= 0.1 \text{ M}\Omega \pm 1\%$	R_{19}	$= R_{22} = 470 \text{ pF} \pm 10\%$	C_{11}	$= 0.39 \mu\text{F} \pm 20\%$
R_{10}	$= 15 \text{ k}\Omega \pm 2\%$	R_{20}	$= R_{23} = 0.0068 \mu\text{F} \pm 10\%$	C_{12}	$= 0.15 \mu\text{F} \pm 20\%$
		R_{21}	$= R_{24} = 2\%$	C_{13}	$= 2\%$
		R_{22}	$= R_{25} = 1\%$	C_{14}	$= 1\%$
		R_{23}	$= R_{26} = 1\%$	C_{15}	$= 1\%$
		R_{24}	$= R_{27} = 1\%$	C_{16}	$= 1\%$
		R_{25}	$= R_{28} = 1\%$	C_{17}	$= 1\%$
		R_{26}	$= R_{29} = 1\%$		
		R_{27}	$= R_{30} = 1\%$		
		R_{28}	$= R_{31} = 1\%$		
		R_{29}	$= R_{32} = 1\%$		
		R_{30}	$= R_{33} = 1\%$		
		R_{31}	$= R_{34} = 1\%$		
		R_{32}	$= R_{35} = 1\%$		
		R_{33}	$= R_{36} = 1\%$		
		R_{34}	$= R_{37} = 1\%$		
		R_{35}	$= R_{38} = 1\%$		
		R_{36}	$= R_{39} = 1\%$		
		R_{37}	$= R_{40} = 1\%$		
		R_{38}	$= R_{41} = 1\%$		
		R_{39}	$= R_{42} = 1\%$		
		R_{40}	$= R_{43} = 1\%$		
		R_{41}	$= R_{44} = 1\%$		
		R_{42}	$= R_{45} = 1\%$		
		R_{43}	$= R_{46} = 1\%$		
		R_{44}	$= R_{47} = 1\%$		
		R_{45}	$= R_{48} = 1\%$		
		R_{46}	$= R_{49} = 1\%$		
		R_{47}	$= R_{50} = 1\%$		
		R_{48}	$= R_{51} = 1\%$		
		R_{49}	$= R_{52} = 1\%$		
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		R_{60}	$= R_{63} = 1\%$		
		R_{61}	$= R_{64} = 1\%$		
		R_{62}	$= R_{65} = 1\%$		
		R_{63}	$= R_{66} = 1\%$		
		R_{64}	$= R_{67} = 1\%$		
		R_{65}	$= R_{68} = 1\%$		
		R_{66}	$= R_{69} = 1\%$		
		R_{67}	$= R_{70} = 1\%$		
		R_{68}	$= R_{71} = 1\%$		
		R_{69}	$= R_{72} = 1\%$		
		R_{70}	$= R_{73} = 1\%$		
		R_{71}	$= R_{74} = 1\%$		
		R_{72}	$= R_{75} = 1\%$		
		R_{73}	$= R_{76} = 1\%$		
		R_{74}	$= R_{77} = 1\%$		
		R_{75}	$= R_{78} = 1\%$		
		R_{76}	$= R_{79} = 1\%$		
		R_{77}	$= R_{80} = 1\%$		
		R_{78}	$= R_{81} = 1\%$		
		R_{79}	$= R_{82} = 1\%$		
		R_{80}	$= R_{83} = 1\%$		
		R_{81}	$= R_{84} = 1\%$		
		R_{82}	$= R_{85} = 1\%$		
		R_{83}	$= R_{86} = 1\%$		
		R_{84}	$= R_{87} = 1\%$		
		R_{85}	$= R_{88} = 1\%$		
		R_{86}	$= R_{89} = 1\%$		
		R_{87}	$= R_{90} = 1\%$		
		R_{88}	$= R_{91} = 1\%$		
		R_{89}	$= R_{92} = 1\%$		
		R_{90}	$= R_{93} = 1\%$		
		R_{91}	$= R_{94} = 1\%$		
		R_{92}	$= R_{95} = 1\%$		
		R_{93}	$= R_{96} = 1\%$		
		R_{94}	$= R_{97} = 1\%$		
		R_{95}	$= R_{98} = 1\%$		
		R_{96}	$= R_{99} = 1\%$		
		R_{97}	$= R_{100} = 1\%$		

units, consisting of the input stage and supply unit, which can be plugged into the apparatus. All units, with the exception of that for the first stage, are identical; in this stage a germanium diode is shunted across the grid leak R_{14} to prevent the potential of g_1 rising at high counting speeds. This diode can be omitted in the following stages, where the counting speed is at least a factor 10 lower.

The connections p can be used for measuring the beam currents. By means of the switch S a voltage of -60 V instead of $+11.9$ V can be temporarily applied to the control grids g_1 of all tubes, for resetting the beams to zero.

Measuring Shutter Speeds of Cameras

For mass production of cameras and for making repairs it is essential to determine the shutter speeds exactly. It was not until electronic means were available that such measurements could be carried out with sufficient accuracy. Apart from being very simple, the method of measuring shutter speeds by means of electronic devices equipped with the decade counter tube E 1 T offers several advantages, the most important of which are that the measurements can be carried out in the workshop and that no calibration is required. The time during which the shutter remains open is indicated directly in figures by the counter tubes.

Fig. 11-15 shows the principle for measuring the shutter speed. A beam of

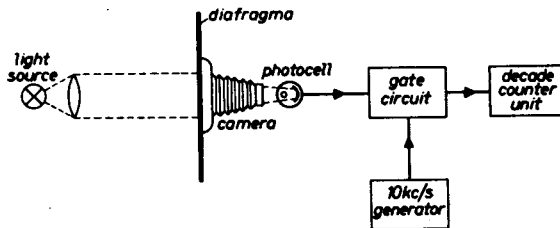


Fig. 11-15. Block diagram of a device for measuring shutter speeds.

light is directed on to a diaphragm, which replaces the film in the camera, passes through the shutter and the lens, and finally impinges on a photocell. The diaphragm has an extremely narrow aperture so that the measured time corresponds to the time during which an extremely small part of the film is exposed.

The photocell controls a gating circuit, which is connected between a 10 kc/s signal generator and a decade counter. When no light impinges on the photocell, the connection between the generator and the counter is interrupted. When, however, the beam of light falls on the photocell, this connection is established, and the number of cycles the gating circuit passes is registered by the counter. When the shutter is operated the counter thus indicates a number that corresponds directly to the shutter speed in ten thousandths of a second.

The measurement is extremely accurate because fluctuations of the light source have no effect. In fact, the device does not react to the amplitude of the pulses but merely to their presence. The accuracy of the measurement depends in the first instance on the accuracy with which the signal generator has been

calibrated and on its frequency constancy; it is quite feasible to reduce the inaccuracy caused thereby to less than 1%. For measuring very fast shutters, it may be advisable to choose a signal generator with a somewhat higher frequency, for example 20 kc/s to 30 kc/s, thereby further improving the accuracy of measurement.

The beam of light should be as narrow as possible in order to obtain reliable results. For that reason the aperture of the diaphragm, which is substituted for the film, should be extremely narrow. It has been proved that a slot of 0.1 mm

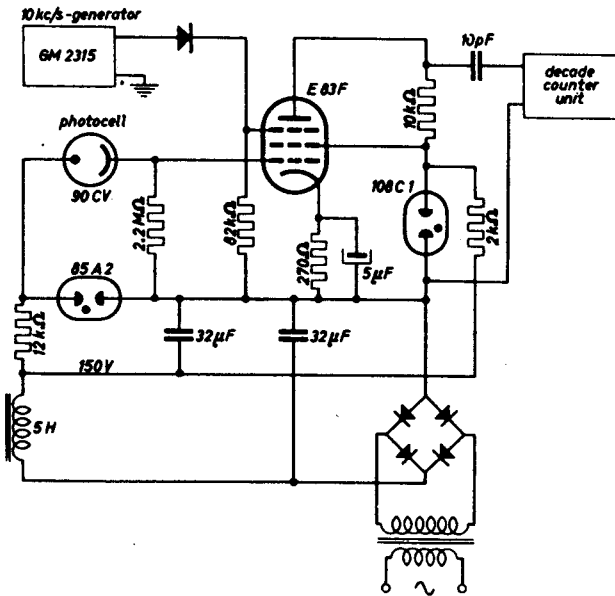


Fig. 11-16. Gate circuit employed in fig. 11-15.

is sufficient, provided a fairly strong source of light is used. The photocell must be of the high-vacuum type because gas-filled photocells have considerable inertia. A Philips signal generator type GM 2315 can be used to advantage.

The decade counter can be constructed according to the circuit shown in fig. 11-14. The operation of the gating circuit will be explained by means of fig. 11-16. The pentode E 83 F is connected so that the voltage pulse originating from the photocell is applied to the first grid; the output voltage of the signal generator GM 2315 is applied to the third grid after having been rectified by means of the germanium diode. The latter is so connected that the pulses at this grid are positive-going. The load resistance of the E 83 F has a value of 10 k Ω ; the counter is connected to this resistor via a capacitor of 10 pF. When no light impinges on the photocell, the gain of the tube will be fairly low, so that the amplitude of the pulses produced across the load resistance is insufficient to operate the counter. When, however, the photocell is illuminated, a positive bias is applied to the first grid of the E 83 F, so that its gain is raised and the pulses are counted.

Predetermined Counter

Apart from performing simple counting operations, decade counter tubes can be used in appropriate circuits for carrying out multiplications in electronic computers, in sorting machines and so forth. In these cases provision must be made that after a predetermined number of pulses has been applied, a certain action, such as the putting into operation of a second counter or a relay, is performed. In these devices the primary cycle of counts may be either incidental or periodical.

It has already been shown that the beam in the decade counter tube is reset from position 9 to 0 at the tenth pulse, a pulse being passed simultaneously to the following counter stage. In the same way the following stages are also operated. When, for example in a four-decade counter the number 9999 is reached, all tubes will be reset to zero at the following pulse, and the pulse produced by the fourth stage may be used for other purposes. This pulse may thus serve to indicate that the number 10 000 has been reached, or, for example, to initiate or to terminate a certain action. If this is required to take place after a predetermined number of counting pulses it is obviously sufficient to apply the complement of this number to the counter before the counting is started. If, for example, it is desired that a signal is produced after the 4796th pulse, the four stages must be adjusted to the number $10\ 000 - 4796 = 5204$ before the counting starts.

It has been mentioned previously that for each stable position of the beam of the counter tube the potential at the deflection plate D' and the anode a_2 has a given value. The difference between the deflection voltages of two successive figures is approximately 14 V. It will thus be clear that if a variable direct voltage is applied to D' and a_2 via a voltage divider, the corresponding counter stage can be adjusted to any digit between 0 and 9 by choosing a different tapping of this voltage divider. A diode is also connected between this tapping and the deflection plate, with its cathode connected to the tapping. The voltage divider may consist of a number of resistors connected in series with a ten-pole switch, the positions of which may be marked with the complementary digits, that is with the digits which form the predetermined number.

Fig. 11-17 shows the complete circuit of a four-decade predetermined counter. The main difference with respect to the circuit of fig. 11-14 consists in the addition of the double diodes EB 91, one system of which is used in combination with each counter tube. A voltage, corresponding to the selected digit and determined by the position of the switches S_p to S_n , is temporarily applied to the right-hand deflection plate D' of each counter tube via a diode system. For this purpose the stage for adjusting the counter to the predetermined number is also equipped with an electronic switch formed by the double triode E 90 CC. By means of this switch the left-hand extremity of the voltage dividing chain $R_{80} \dots R_{88}$ is separated from the negative terminal of the supply voltage once the selected number has been adjusted, before counting is started. A positive voltage of 300 V is thereby applied to the cathode of the diode systems of the EB 91, so that these become non-conducting. The electron beams of the counter tubes, however, remain at the positions to which they have been adjusted.

Let us assume that the decade counter tube of the first stage has been adjusted to the digit 4 as in the previous example. After 6 pulses have been applied, the beam will return to zero, and the tube will continue to count in the normal

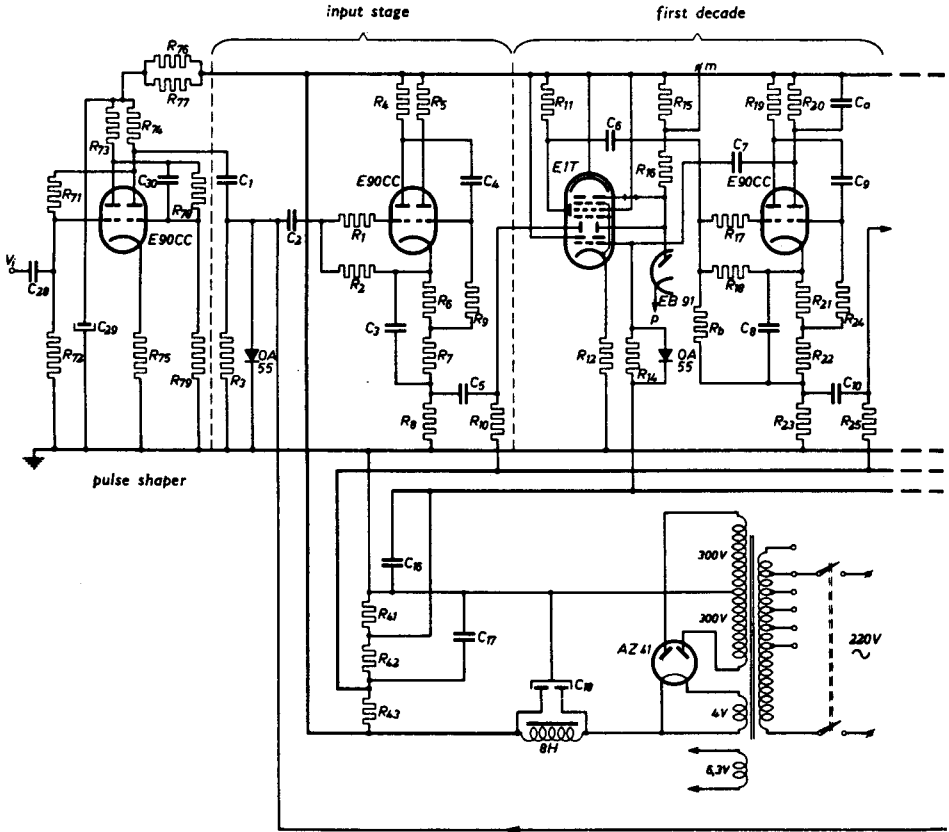
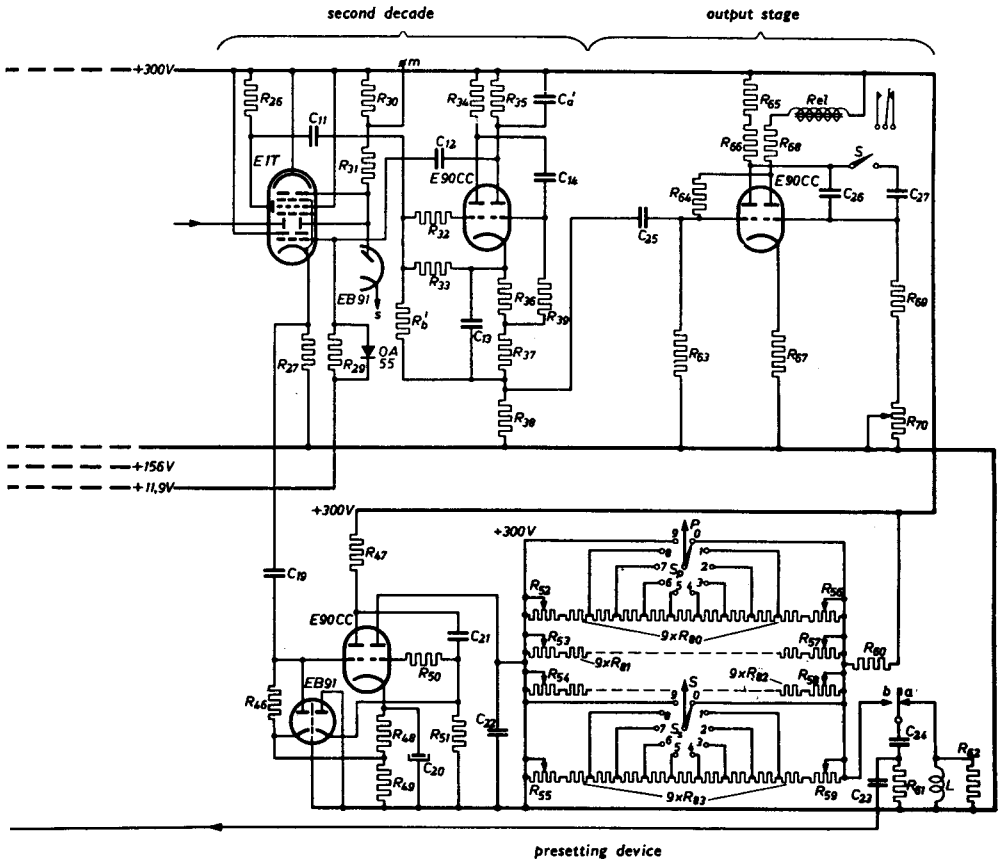


Fig. 11-17. Circuit of a

COMPONENT

R_1	=	56 k Ω \pm 10%, 1/2 W
R_2	=	0.56 M Ω \pm 10%, 1/2 W
R_3	=	56 k Ω \pm 10%, 1/2 W
R_4	=	39 k Ω \pm 2%, 2 W
R_5	=	3.3 k Ω \pm 2%, 1/2 W
R_6	=	4.7 k Ω \pm 2%, 1 W
R_7	=	2.7 k Ω \pm 2%, 1/2 W
R_8	=	1 k Ω \pm 1%, 1/2 W
R_9	=	0.1 M Ω \pm 1%, 1/2 W
R_{10}	=	15 k Ω \pm 2%, 1/2 W
$R_{11} = R_{30}$	=	39 k Ω \pm 5%, 1/2 W
$R_{12} = R_{31}$	=	15 k Ω \pm 1%, 1/2 W
$R_{13} = R_{32}$	=	0.33 M Ω \pm 10%, 1/2 W
$R_{14} = R_{33}$	=	10 k Ω \pm 10%, 1/2 W
$R_{15} = R_{34}$	=	1 M Ω \pm 1%, 1/2 W
$Rb = Rb'$	=	8.2 M Ω \pm 2%, 1/2 W
$R_{17} = R_{35}$	=	5.6 k Ω \pm 10%, 1/2 W
$R_{18} = R_{36}$	=	0.56 M Ω \pm 5%, 1/2 W
$R_{19} = R_{37}$	=	39 k Ω \pm 2%, 2 W
$R_{20} = R_{38}$	=	3.3 k Ω \pm 2%, 1/2 W
$R_{21} = R_{39}$	=	4.7 k Ω \pm 2%, 1/2 W

$R_{22} = R_{37}$	=	2.7 k Ω \pm 2%, 1/2 W
$R_{23} = R_{38}$	=	1 k Ω \pm 1%, 1/2 W
$R_{24} = R_{39}$	=	0.15 M Ω \pm 2%, 1/2 W
R_{25}	=	15 k Ω \pm 2%, 1/2 W
R_{26}	=	5.6 k Ω \pm 1%, 1/2 W
R_{27}	=	68 k Ω \pm 1%, 1 W
R_{28}	=	68 k Ω \pm 1%, 1 W
R_{29}	=	0.56 M Ω \pm 10%, 1/2 W
R_{41}	=	18 k Ω \pm 2%, 1/2 W
R_{42}	=	1.5 k Ω \pm 2%, 1/2 W
R_{43}	=	3.9 k Ω \pm 2%, 1/2 W
R_{44}	=	33 k Ω \pm 10%, 1/2 W
R_{45}	=	1 M Ω \pm 10%, 1/2 W
$R_{46} - R_{49}$	=	10 k Ω
R_{50}	=	3.9 k Ω \pm 2%, 1/2 W
R_{51}	=	100 Ω \pm 10%, 1/2 W
R_{52}	=	680 Ω \pm 10%, 1/2 W
R_{53}	=	0.1 M Ω \pm 2%, 1/2 W
R_{54}	=	1 M Ω \pm 2%, 1 W
R_{55}	=	15 k Ω \pm 5%, 1/2 W
R_{56}	=	47 k Ω \pm 5%, 1 W



predetermined decade counter.

VALUES

R_{47}	= 10	$k\Omega \pm 2\%$, $\frac{1}{2}$ W	$C_8 = C_{13}$	= 68	pF $\pm 2\%$
R_{48}	= 6.8	$k\Omega \pm 5\%$, $\frac{1}{2}$ W	$C_9 = C_{12}$	= 270	pF $\pm 10\%$
R_{49}	= 0.33	$M\Omega \pm 10\%$, $\frac{1}{2}$ W	C_9	= 100	pF $\pm 2\%$
R_{70}	= 2.6	$M\Omega$	C_{10}	= 680	pF $\pm 5\%$
R_{71}	= 56	$k\Omega \pm 2\%$, $\frac{1}{2}$ W	C_{14}	= 180	pF $\pm 2\%$
R_{72}	= 56	$k\Omega \pm 2\%$, $\frac{1}{2}$ W	C_{16}	= 0.39	μF
R_{73}	= 8.2	$k\Omega \pm 2\%$, $\frac{1}{2}$ W	C_{17}	= 0.15	μF
R_{74}	= 8.2	$k\Omega \pm 2\%$, $\frac{1}{2}$ W	C_{18}	= 2×50	μF , 450 V
R_{75}	= 12	$k\Omega \pm 2\%$, $\frac{1}{2}$ W	C_{19}	= 470	pF $\pm 10\%$
R_{76}	= 56	$k\Omega \pm 2\%$, 1 W	C_{20}	= 25	μF , 25 V
R_{77}	= 56	$k\Omega \pm 2\%$, 1 W	C_{21}	= 560	pF $\pm 10\%$
R_{78}	= 56	$k\Omega \pm 2\%$, $\frac{1}{2}$ W	C_{22}	= 180	pF $\pm 10\%$
R_{79}	= 56	$k\Omega \pm 2\%$, $\frac{1}{2}$ W	C_{23}	= 15	pF $\pm 10\%$
$R_{80} - R_{83}$	= elk 9×4.7	$k\Omega \pm 2\%$, $\frac{1}{4}$ W	C_{24}	= 0.47	μF
C_1	= 220	pF $\pm 10\%$	C_{25}	= 22	pF $\pm 10\%$
C_2	= 220	pF $\pm 10\%$	C_{26}	= 0.058	μF $\pm 10\%$
C_3	= 100	pF $\pm 2\%$	C_{27}	= 0.47	μF
C_4	= 39	pF $\pm 2\%$	C_{28}	= 1	μF
C_5	= 0.0068	μF $\pm 10\%$	C_{29}	= 25	μF , 300 V
$C_6 = C_{11}$	= 220	pF $\pm 10\%$	C_{30}	= 100	pF $\pm 10\%$
$C_7 = C_{12}$	= 0.0068	μF $\pm 10\%$	L	= 1	mH

way. The counter tubes of the other stages will act in the same way, until — after the 4795th pulse has been applied — the number 9999 is registered. The output stage, which again consists of an E 90 CC connected as a monostable multivibrator and which has a relay in its anode circuit, then comes into operation. Moreover, the negative-going pulse, which is produced at the cathode of the last counter tube when its beam is reset to zero, is applied to the left triode section of the above-mentioned E 90 CC operating as an electronic switch. The resulting positive-going pulse at the anode of this section is fed to the grid of the right triode section via C_{21} , so that the latter section passes current during a short interval. As a result, all the counter tubes are reset to the predetermined figure, so that the cycle can start anew.

Provision has also been made in the circuit for a static check to be carried out. This can be achieved by means of the circuit drawn at the extreme right in fig. 11-17. When the switch knob $a - b$ is pressed, the potential at all tapings of the voltage dividers will drop to such an extent before the capacitor C_{24} is charged that the beams of all counter tubes are advanced to position 9. If this does not occur, the switch makes poor contact. If the switch is subsequently released so that it returns to position a , the capacitor C_{24} will be discharged via R_{61} and R_{62} shunting the inductor L . The negative-going pulse thus produced is fed to the input of the predetermined counter and causes the beams of all tubes to be advanced to the predetermined figures.

Multiplying circuit

It is possible to construct electronic computers in a simple way by using predetermined counters. Fig. 11-18 shows, by way of example, the block diagram of a device for multiplying n factors.

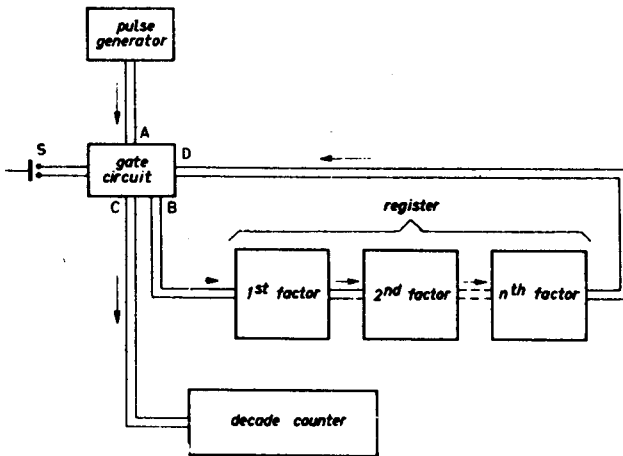


Fig. 11-18. Block diagram of a multiplying device for n factors.

The clock pulses supplied by a pulse generator are passed through a gate circuit to a decade counter and a register, which consists of n predetermined counters. The gate circuit is opened by closing switch S . As soon as the number of pulses passed is equal to the product of the n factors, the gate circuit is closed

again by the reset pulse produced by the last counter tube of the n^{th} predetermined counter. The product is then indicated by the decade counter unit.

In practice the use of such a circuit is restricted because the multiplication of numbers consisting of two or more digits would take too much time. By slightly modifying the circuit it is, however, possible to have several arithmetical operations carried out simultaneously. To multiply $x_1 x_2 x_3$ by $y_1 y_2 y_3$ it is, for example, possible to multiply x_3 simultaneously by y_1, y_2 and y_3 , and subsequently x_2 simultaneously by y_1, y_2 and y_3 , etc. The partial results are then continuously added by the decade counter. A rough calculation shows that at an operating frequency of the counter of 30 000 pulses per second, less than 200 milliseconds are required on an average for multiplying two numbers of 16 digits each.

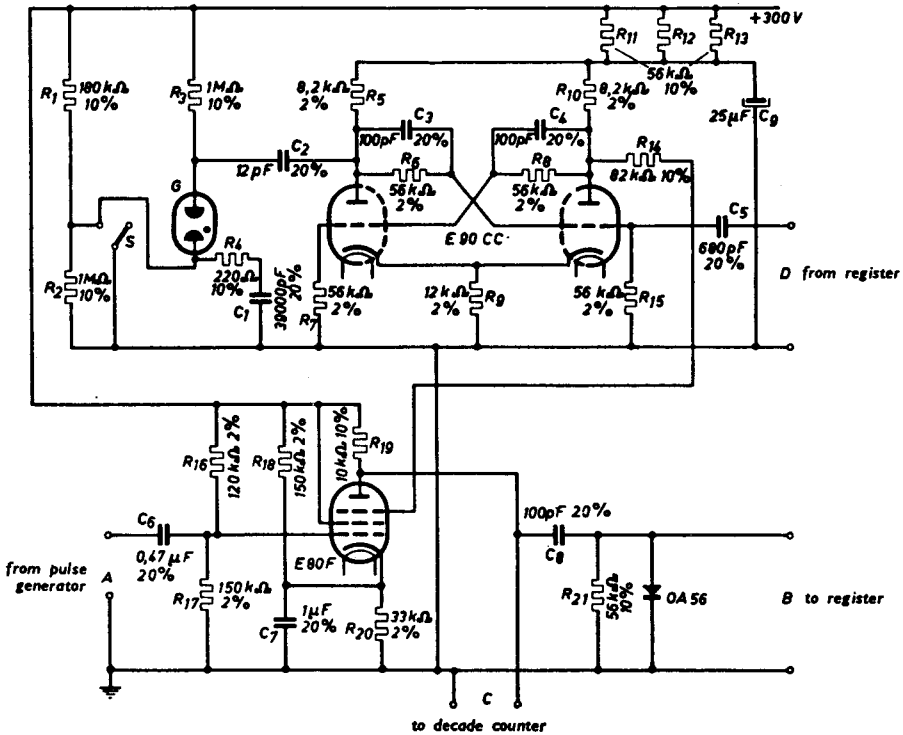


Fig. 11-19. Gate circuit applied in fig. 11-18.

Fig. 11-19 shows the complete gate circuit, consisting of a double triode E 90 CC connected as a bi-stable multivibrator and a pentode E 80 F operating as a gating tube. At rest, the left triode section of the E 90 CC is cut off, whereas the right section is conducting. A voltage of 95 V is thus applied to the third grid of the gating tube since this grid is connected to the anode of the right triode section; as a positive voltage of 140 V is applied to the cathode of the gating tube, the latter is cut off. The anode current is approximately 0.1 mA and the screen-grid current 2.25 mA under these conditions.

If switch S is now temporarily closed, the glow discharge lamp G will be ignited, and the negative-going pulse thus produced is applied, via C_2 , to the

anode of the left triode section. (When switch S is closed again, the glow discharge lamp will be extinguished since C_1 is charged). The pulse brings the multivibrator into the other stable condition, so that the left triode section becomes conducting whereas the right section is cut off. The voltage at the third grid of the gating tube is thus raised to 140 V, which results in the pulses applied to the control grid being amplified; the output pulses at the terminals B and C can be fed to the calculating device and the decade counter unit respectively.

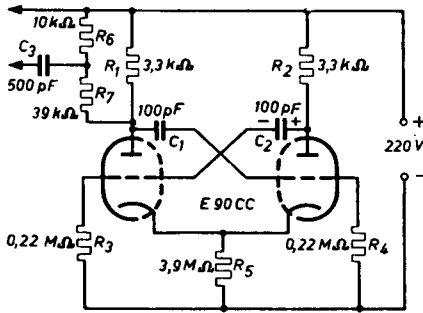


Fig. 11-20. Multivibrator for generating clock pulses.

The gate circuit is closed by a positive-going pulse which is supplied by the calculating device and applied to the control grid of the right triode section via C_5 . The double triode is thus returned to its original condition so that the pulses originating from the generator are no longer passed.

The simple multivibrator circuit shown in fig. 11-20 can be used for producing the clock pulses. Its operation

may be explained as follows. Assume C_2 to be charged with the polarity indicated, and the left triode section to be cut off, whereas the right triode section is conducting; C_1 will then be charged almost up to 250 V. As soon as C_2 is sufficiently discharged via R_3 , the left triode section becomes conducting. This causes the voltage at the left anode to decrease, whilst the right triode section is cut off as a result of the charge across C_1 . Subsequently C_1 is discharged via R_4 until the right section becomes conducting again and the left section is cut off. The frequency of the pulses thus produced is determined by the values of C_1 , C_2 and R_3 , R_4 ; with the values indicated in the circuit diagram, this frequency is approximately 10 kc/s.

Subtracting and Dividing by means of Decade Counter Tubes

Subtractions can be carried out by means of a circuit according to the block diagram shown in fig. 11-21. The pulses are applied to two predetermined

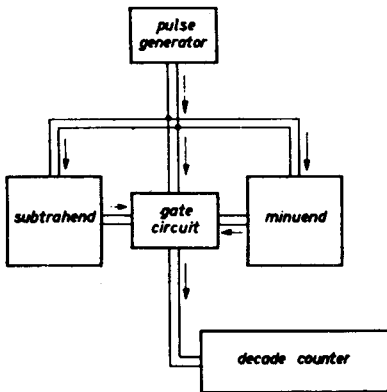


Fig. 11-21. Block diagram of a device for carrying out subtractions.

counters which have been preset to the subtrahend and the minuend. The last stage of the counter, preset to the subtrahend, opens the gate circuit, so that the decade counter starts to register the pulses until the last stage of the counter, preset to the minuend, again closes the gate circuit. The difference is then recorded by the decade counter unit.

For carrying out divisions a circuit according to the block diagram of fig. 11-22 may be used. The predetermined counters are now preset to the divisor and the dividend. Once switch S is closed, the pulses are fed to the counter adjusted to the divisor, and this supplies a pulse to the decade counter after each completed cycle. As soon as the counter preset to the dividend has received a number of pulses that is equal to the dividend, this counter closes the gate circuit again. The quotient is then registered by the decade counter, the remainder being indicated by the counter which has been preset to the divisor.

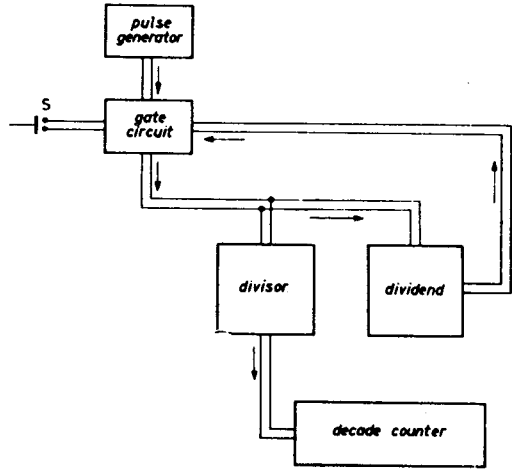


Fig. 11-22. Block diagram for carrying out divisions.

Reading out of the Number of Counts

It is often desired to read out the result of a calculation carried out by the decade counter, so that it can be recorded on a strip of paper, for example. This may be achieved by a circuit according to fig. 11-23.

The counter tube of each decade is connected to a gate circuit; a series of 10 pulses is applied both to the counter tube and to the gate circuit. Assume the counter tube to indicate the figure 4; after the 6th pulse a fly-back pulse will be produced by the counter tube, and this opens the gate circuit. The following 4 pulses return the beam of the counter tube to its original position, and also reach the recording device by way of the gate circuit.

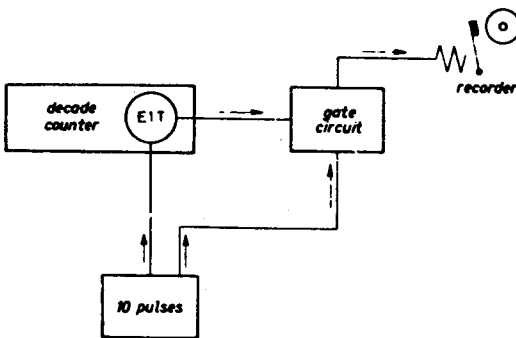


Fig. 11-23. Block diagram of a reading-out device.

12. ELECTRONIC TIMERS

A timer is a device which causes or permits some process or action to commence at a particular moment and to cease after the expiration of a predetermined period. Its operation may be automatic and continuous, or may be controlled manually or otherwise so that it operates once only ("single stroke" operation). Electronic timing devices have a wide range of application. They are employed, for example, for the automatic control of electric welding machines; for process control in chemical laboratories; and for timing exposures in radiography and studio photography.

The operation of all electronic timers is based upon the time taken for a capacitor to charge or discharge through a resistor. Before considering some practical timing circuits, therefore, it will be useful to examine the relationships existing in such a circuit.

Starting from the well-known fact that the voltage across a capacitor is $v = q/C$, where q is the charge and is equal to:

$$q = \int i \cdot dt,$$

it can be shown that when a capacitor, charged to a voltage V_0 , is discharged through a resistor R as shown in fig. 12-1,

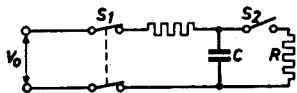


Fig. 12-1. Discharge of a capacitor through a resistor.

$$\frac{1}{C} \cdot \int i \cdot dt + R \cdot i = 0 \quad (12.1)$$

whence $i = I_0 \cdot e^{-t/RC}, \quad (12.2)$

and $v = V_0 \cdot e^{-t/RC}. \quad (12.3)$

The voltage v across the capacitor decreases with time exponentially, as shown in fig. 12-2, and it is obvious that, after a time interval $t = T = R \cdot C$, v will have decayed to the value

$$v = V_0/e = 0.368 V_0. \quad (12.4)$$

$R \cdot C$ is termed the "time constant" of the circuit.

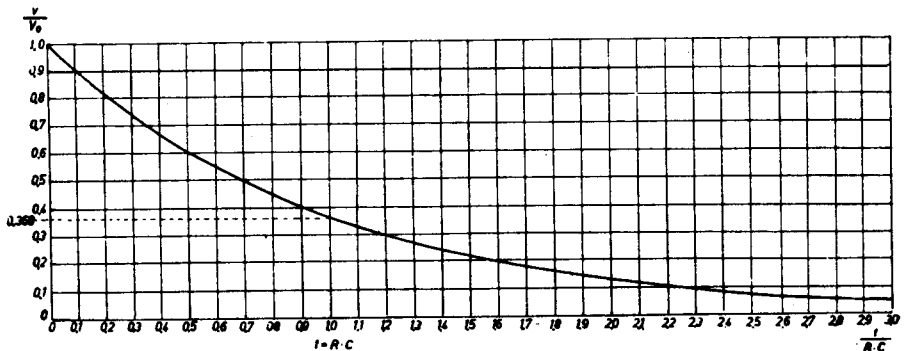


Fig. 12-2. Voltage decrease as a function of time when discharging a capacitor through a resistor.

In the general case, where v has fallen to a value V_0/n :

$$\frac{1}{n} = e^{-t/RC}, \tag{12.5}$$

or

$$t = -RC \cdot \log_e \frac{1}{n}. \tag{12.6}$$

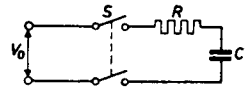


Fig. 12-3. Charge of a capacitor through a resistor.

It should be noted that the discharge time t is purely a function of the capacitance and the resistance, and is quite independent of the voltage V_0 . To obtain the value of t in seconds, R must be expressed in megohms and C in microfarads.

In the case where a capacitor is charged by a voltage V_0 through a resistor R as shown in fig. 12-3, the voltage across the capacitor after a time interval t is:

$$v = V_0 (1 - e^{-t/RC}). \tag{12.7}$$

After a time interval $T = RC$, therefore:

$$v = V_0 (1 - e^{-1}) = 0.632 V_0. \tag{12.8}$$

To take a numerical example, if $R = 1$ megohm, and $C = 8$ microfarads, and it is desired that the voltage across the capacitor should rise from zero to 10 volts in $T = RC = 8$ seconds, then the applied voltage must be:

$$V_0 = \frac{10}{0.632} = 15.8 \text{ V.}$$

Consider finally the case when a capacitor C is charged to a voltage $+V_0$ and is then discharged through resistor R against an opposing voltage $-V_0$ (see fig. 12-4). The reversal of the charge from $+V_0$ to $-V_0$ will again take place exponentially. At $v = 0$ half this change will be reached, so that, according to equ. (12.5), $n = 2$. The charge on the capacitor will therefore have fallen to zero after a time interval:

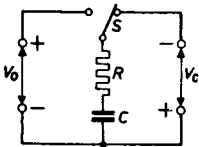


Fig. 12-4. Charging a capacitor with reverse polarity.

$$t = 0.693 RC. \tag{12.9}$$

This means that after $0.693 RC$ seconds the voltage across C will have just fallen to zero, irrespective of the value of the charging voltage V_0 .

Timer for Photographic Enlarger

A timer operating on the principles described above may use the voltage across the capacitor for firing a thyatron, the anode current of which operates a relay to switch a process on or off. Fig. 12-5 is the circuit diagram of such a timer, suitable for controlling the exposure time when enlarging photographs. The tube used is the inert gas-filled tetrode thyatron PL2D21 which is particularly suited for this application as it can be operated at an average anode current of 100 milliamperes.

When the mains supply is switched on, positive control grid and anode voltages are applied to the thyatron via rectifier V_1 . The thyatron therefore ignites and energises the relay, thus interrupting the lamp circuit and connecting capacitor

C_1 to rectify V_2 so that C_1 is charged to a potential of -100 volts. Although its control grid is now at a negative potential, the thyatron is not extinguished because the anode voltage has not been removed.

However, if switch S_2 is now thrown over from one position to the other,

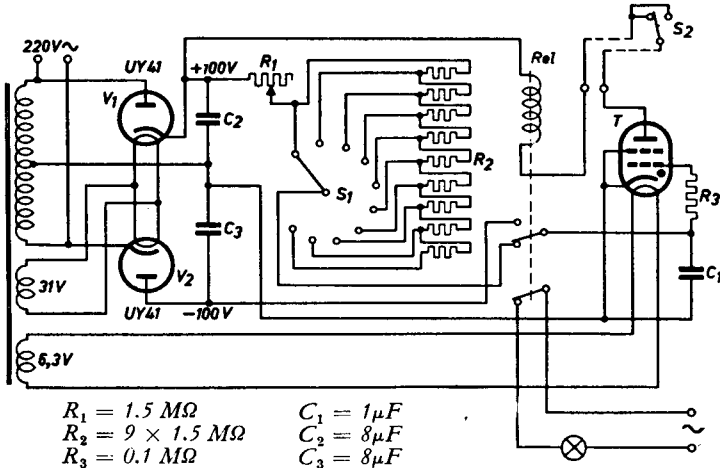


Fig. 12-5. Timing circuit for controlling the time of exposure.

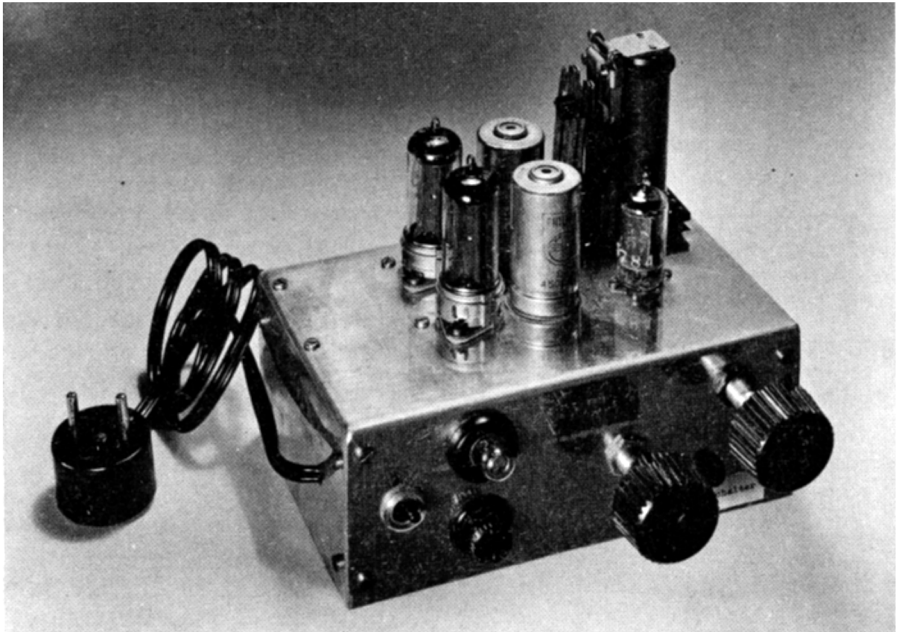


Fig. 12-6. Experimental unit using the circuit of fig. 12-5.

the anode supply will be interrupted for a few milliseconds and, since the de-ionisation time of the thyatron is very short, the tube remains extinguished owing to the negative control grid voltage supplied by the charge on C_1 . The relay is thus de-energised, and closes the lamp circuit. At the same time C_1 is connected to a 100-volt supply of opposite polarity (+100 volts) via resistors R_1 and R_2 . R_2 consists of nine resistors which can be switched in by means of the step switch S_1 , to adjust the discharge time up to a maximum of approximately 10 seconds. When the capacitor has discharged down to a voltage of about -2 volts, the thyatron ignites again, energising the relay and interrupting the lamp circuit. The timing circuit has now returned to its original condition, and the lamp will remain extinguished until switch S_2 is operated once more.

The time interval, the value of which can be calculated by equation (12.9), can be varied by means of resistors R_1 and R_2 . By using resistors of greater value, the control range can be extended up to approximately three minutes. A photograph of an experimental timer using this circuit is reproduced in fig. 12-6.

Timer giving Two Successive Intervals

An electronic timing circuit which produces two successive time intervals, each of which can be varied independently of the other, is shown in fig. 12-7. This circuit, however, is suitable only for time intervals which are large compared with the duration of one cycle of the mains supply. (A circuit to which this limitation does not apply, since it is operated from a d.c. supply, is shown in fig. 12-9).

When the circuit is at rest, thyratrons T_1 , T_2 and T_4 are extinguished, whereas thyatron T_3 is ignited. Capacitor C_1 now charges via the grid-to-cathode conduction of T_2 , and acquires the polarity indicated in the diagram.

Capacitor C_3 is also charged with the polarity indicated as a result of the anode current of T_3 flowing through R_6 .

If switch S is now temporarily closed, T_1 ignites, and the relay Rel is energized. Although this results in the cathode of T_2 being connected to the lower supply line, this tube cannot ignite until C_1 has almost completely discharged through R_1 .

When T_2 fires, the positive half cycles of the supply voltage which hitherto had been charging capacitor C_2 through R_3 are practically short-circuited, and the voltage across C_2 appears as a negative voltage at the control grid of T_3 , which is thus immediately blocked. C_3 can now discharge through resistor R_6 , the discharge time being determined by the setting of this variable resistor.

When C_3 has discharged almost completely in this way, T_4 fires. The resulting voltage across C_4 is applied to the screen grid of T_1 , so that this tube is also extinguished, and the relay is de-energised. T_2 is thereby also extinguished, so that T_3 cuts off and T_4 ignites, the original condition thus being restored.

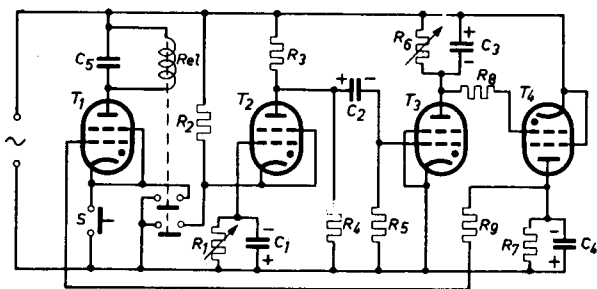


Fig. 12-7. Timing circuit for two successive intervals.

Continuously-Acting Timer

A very simple circuit for continuously producing the same time cycle, and therefore suitable for use as a relaxation oscillator, is shown in fig. 12-8. Capacitor C is charged via variable resistor R_1 by a direct voltage. During the charging period the voltage drop across R_1 applies a negative potential to the control grid of the thyatron which is therefore blocked. When C_1 is fully charged, however, the tube fires, permitting the capacitor to discharge through inductor L . Owing to the self-inductance of L , which tends to maintain the discharge current, C is momentarily re-charged with the opposite polarity, thus extinguishing the thyatron PL2D21. The cycle then repeats. L may represent the primary winding of a transformer, in the secondary winding of which considerable

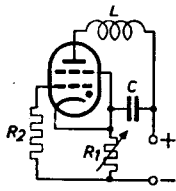


Fig. 12-8. Timer producing periodically the same interval.

voltage peaks are generated due to the high current surges flowing in the primary winding (see fig. 10-5). The values of L and C should be so chosen that the permissible peak current of the thyatron is not exceeded.

Another Timer giving Two Successive Intervals

In the circuit shown in fig. 12-9 two PL 2D 21 thyratrons are alternately ignited and extinguished for periods which can be independently varied by adjusting the value of resistors R_1 and R_4 . When the direct supply voltage is switched on, a positive voltage surge is transferred via C_1 to the control grid of T_1 , ensuring that this tube fires first. The anode potential of T_1 now drops to approximately 8 volts (the arc voltage value) and the control grid voltage falls to approximately cathode potential. Capacitors C_3 and C_4 can therefore charge via R_2 almost up to the full supply voltage.

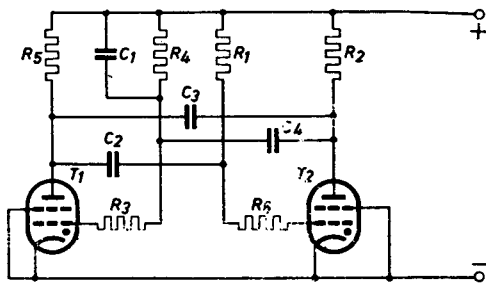


Fig. 12-9. Timer for two successive time intervals.

When tube T_2 ignites, its anode potential similarly drops to the arc voltage value, while the anode voltage of T_1 momentarily falls to a still lower value due to het charge on C_3 . T_1 is thus extinguished and remains non-conducting because of the negative potential applied to its control grid via C_4 . When, however, C_4 has discharged through R_2 and R_4 , T_1 can again ignite. By this time, however, C_2 and C have been charged in the reverse direction so that, when T_1 fires, T_2 is extinguished and remains extinguished until C_2 has discharged via R_1 and R_5 , after which the cycle repeats.

In order to switch on and off the processes to be controlled, suitable relays may be inserted in place of resistors R_2 and R_5 .

Another Single-Stroke Photographic Timer

Fig. 12-10 shows a simple alternative circuit for controlling the exposure time in photographic studios. Although designed for a 110 volt a.c. supply

it may, of course, be operated on other a.c. supply voltages provided suitable tappings are arranged on the primary of the filament transformer.

Before making an exposure, switches S_2 and S_3 are in the position shown in the diagram. When the main switch S_1 is closed, the thyatron PL2D21 ignites, energising relay *Rel* so that the relay contacts switch on L_1 (the red "safe light" in the studio) and extinguish the main lighting L_2 . If necessary, L_2 can be switched on by means of S_3 while adjusting the enlarging apparatus, but S_3 must be opened again before making the automatically-timed exposure.

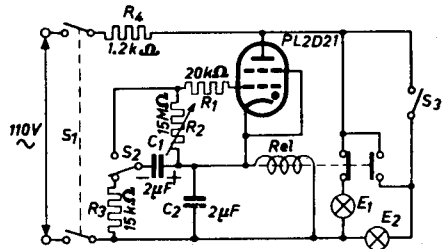


Fig. 12-10. Another timing circuit suitable for photographic studios.

To make the exposure, switch S_2 is thrown over to its other position. Capacitor C_1 , which has hitherto been charged to approximately 100 volts by the voltage drop across the relay winding, is now connected to the control grid of the thyatron, and the tube is extinguished. The relay is thus de-energised, and its contacts open the circuit of the red lamp L_1 and switch on the main light L_2 . C_1 now discharges via R_2 , this resistor being variable in order to adjust the time of exposure. When C_1 is almost discharged, the thyatron again ignites, energising the relay to switch off L_2 , thus completing the exposure, and at the same time switching on the red lamp.

The exposure time can be adjusted over a range of from 1 to 50 seconds; the use of a step switch with suitable fixed value resistors is recommended.

Timer for Small Electric Spot-Welder

The quality of a weld made by a spot-welding machine depends greatly on the accurate timing of the individual welds. In order to minimise the volume of material which is heated to welding temperature, the time of each weld should be as short as possible consistent with producing a satisfactory weld, and this time may sometimes correspond to only a few cycles of the alternating current supply. A circuit for an electronic weld timer suitable for small table welding machines is shown in fig. 12-11.

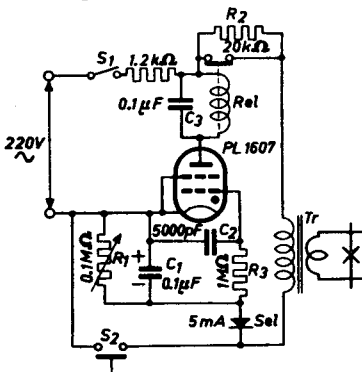


Fig. 12-11. Electronic weld timer.

The thyatron PL1607 thus receives a negative control grid voltage and is prevented from firing. When pedal switch S_2 is closed, the welding current flows, and since S_2 short-circuits the selenium rectifier *Sel* and R_1 , the charging of C_1 ceases and C_1 discharges through R_1 . The discharge time, which is determined by the value of R_1 , controls the welding time because

as soon as C_1 is almost completely discharged the thyatron fires and energises the relay, thus interrupting the welding transformer circuit. When S_2 is reopened, C_1 commences to charge again, the thyatron will be extinguished, and the original "rest" condition is restored.

Theoretically the welding current could be interrupted by premature opening of S_2 , but in practice this is not possible because the welding time corresponds to only 1 to 4 cycles and a mechanical pedal switch cannot generally be opened and closed in such a short time.

Alternative Welding Timer Circuit

Fig. 12-12 is the circuit of a welding timer used in the U.S.A. and conforming with the standards of the National Electrical Manufacturers' Association (N.E.M.A.).

When the circuit is at rest, with switch S open, current flows from the cathode to the grid of the high-vacuum tube V , and charges capacitor C_1 with the polarity indicated. The actual voltage appearing across C_1 is determined by the position of the slider of potentiometer R_2 . Immediately the starting switch S is closed, the thyatron T (PL2D21) is ignited so that relay Rel is energised and its contacts close; this is the start of the welding time.

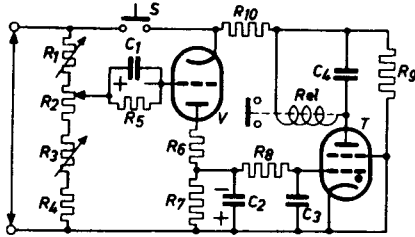


Fig. 12-12. NEMA weld timer circuit.

At the same time, closing of switch S completes the grid circuit of V , but anode current will not flow because of the heavy negative grid voltage applied by capacitor C_1 . However, this capacitor now discharges through R_5 and the negative voltage at the grid of V decays until V begins to pass current. Capacitor C_2 now charges with the polarity indicated in the diagram so that the control grid of the thyatron T is driven negative, eventually extinguishing T and ending the welding time. The welding time is controlled by adjusting the value of R_2 ; the variable resistors R_1 and R_3 serve for the initial calibration of the dial of R_2 .

Timer for Controlling Four Stages in the Welding Operation

For the complete automatic control of welding it is often required to time four successive phases in the operation, each time interval being independently adjustable. These four periods are:

- (1) The "squeeze" time during which the welding electrodes close on the work and build up sufficient mechanical pressure.
- (2) The "weld" time, during which current flows through the electrodes and the work.
- (3) The "hold" time during which the welding current is cut off but the pressure of the electrodes is maintained, allowing the molten metal to solidify and form a strong joint.
- (4) The "off" time during which the solenoid valve opens the electrodes and moves the welding arm to the new location ready to make the next weld.

An electronic timing device suitable for this form of control is shown in the

circuit of fig. 12-13. The four time intervals are produced by four timing circuits, each equipped with a small inert gas-filled tetrode thyratron type PL2D21. In order to understand the operation of this device the circuit of tube T_1 may first be studied.

The anode of this tube is connected through the coil of relay Rel_2 to terminal A of the secondary winding of the mains transformer Tr . The cathode of T_1 is also connected to terminal A via resistor R_7 , and the control grid is connected via resistors R_5 and R_6 to the slider of potentiometer R_3 — point B on the voltage

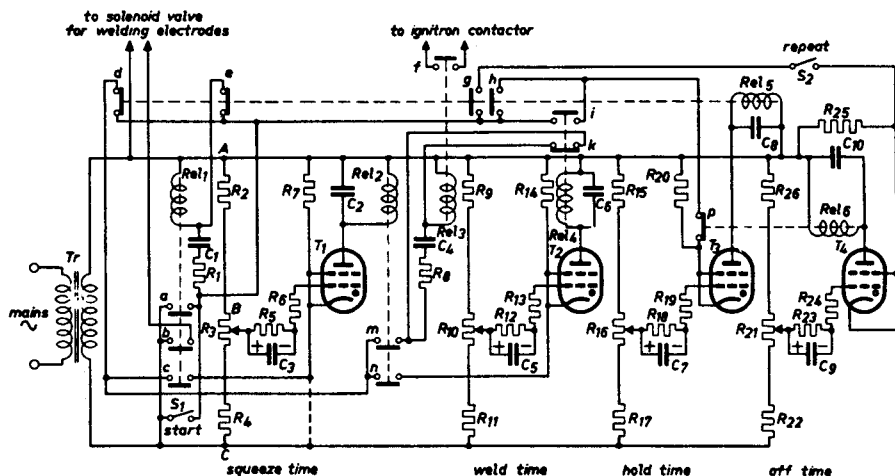


Fig. 12-13. Four-time weld timer.

divider chain formed by R_2 , R_3 and R_4 . Between points A and B , therefore, there exists an alternating voltage, the value of which can be varied by means of R_3 .

During those half cycles of the mains supply during which point B is positive with respect to point A , current will flow via the grid-to-cathode conduction of T_1 through R_5 , R_6 and R_7 , and the voltage drop across R_5 charges capacitor C_3 with the polarity shown. It is clear that the voltage appearing across C_3 is determined by the value of the alternating voltage between A and B , that is to say upon the position of the slider of potentiometer R_3 . C_3 will charge almost up to the peak value of this voltage because it cannot discharge to any great extent during the half cycles when B is negative with respect to A .

Now imagine that by some means or other the cathode of tube T_1 is connected to terminal C of the a.c. supply as indicated by the dotted line in fig. 12-13. The anode circuit of T_1 is immediately completed, but the tube remains non-conducting because the voltage applied to its control grid consists of the negative direct voltage derived from capacitor C_3 and the alternating voltage existing between B and C . However, C_3 now commences to discharge through R_5 and the negative voltage at the grid of T_1 slowly decays until, after a time interval predetermined by the setting of R_3 , tube T_1 ignites. The curve representing the voltage appearing at the control grid of T_1 during this time is given in fig. 12-14.

The complete operation of the circuit of fig. 12-13 can now be explained. When the circuit is at rest with S_1 open, none of the cathodes of tubes T_1 to T_4 is connected to terminal C of the a.c. supply, so all the tubes are non-conducting; however, capacitors C_3 , C_5 , C_7 and C_9 are charged via the grid-to-cathode conduction of their respective tubes in the manner already described. To start a welding sequence switch S_1 is closed, energising relay Rel_1 via contact e . The solenoid valve is thus switched on via relay contact b and the "squeeze time" commences. At the same time Rel_1 holds itself in circuit via contact a , and the

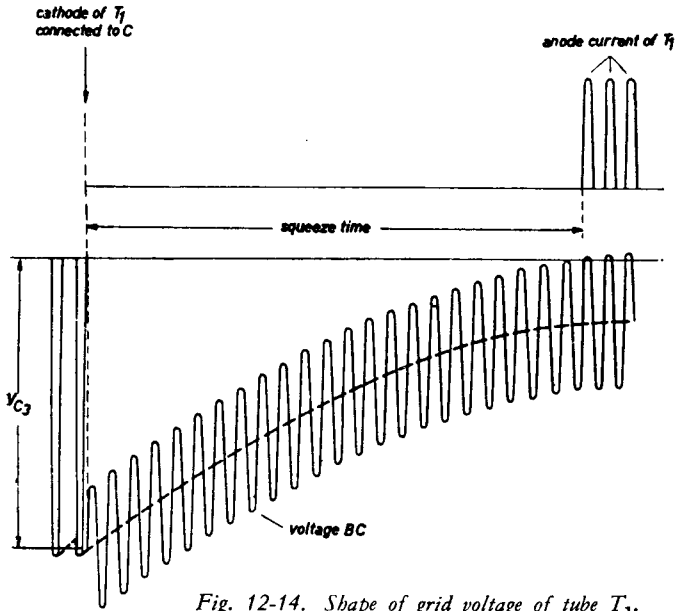


Fig. 12-14. Shape of grid voltage of tube T_1 .

cathode of T_1 is connected to terminal C of the a.c. supply via contacts c , d and a . Capacitor C_3 now discharges through R_5 , controlling the "squeeze time" until T_1 ignites, energising Rel_2 . Contact m closes the circuit of Rel_3 via contact k . Rel_3 switches on the ignitron contactor, thus starting the weld time.

Simultaneously the cathode of T_2 is connected to terminal C via contacts n , d and a , and C_5 begins to discharge through R_{12} thus producing the weld time. When C_5 has discharged sufficiently to allow T_2 to ignite, the anode current of T_2 energises relay Rel_4 , which interrupts the circuit of Rel_3 by contact k .

At the same time the cathode of T_3 is connected to terminal C via contacts p , i and a , and the "hold" time commences. Capacitor C_7 now discharges through R_{18} , producing the "hold" time, until T_3 ignites, energising Rel_5 . Contact b is thus closed, short-circuiting contact i and thus ensuring that the anode circuit of T_3 will not be interrupted when Rel_4 is de-energised.

At the same time contact d is opened thus interrupting the anode circuits of T_1 and T_2 and extinguishing these tubes so that Rel_2 and Rel_4 are de-energised; and finally contact e interrupts the circuit of Rel_1 thus operating the solenoid valve which opens the welding electrodes.

If, during this time, switch S_2 were open, the welding cycle would now cease

and the circuit be restored to its "rest" condition, switch S_1 being opened and therefore T_3 being extinguished.

If, however, switch S_2 were closed during the welding cycle, the cathode of T_4 would be connected to terminal C via contact g and S_1 and the period of "off" time would commence. Capacitor C_9 now discharges through R_{23} , determining the "off" time, until T_4 fires and energises Rel_6 , interrupting the cathode circuit of T_3 at contact p and extinguishing that tube. Rel_5 is thus de-energised and interrupts the circuit of T_4 at contact g , extinguishing T_4 .

At the same time, however, contact e closes again, and since switch S_1 is still closed, Rel_1 is once more energised, thus starting the next complete welding cycle. This process is repeated as long as S_1 is closed. If this switch is opened before expiration of a complete welding period, this period will nevertheless be completed before the circuit returns to its "rest" condition.

Another Welding Timer Producing Four Time Intervals

A somewhat similar circuit which produces four time intervals is given in fig. 12-15. The anodes of the four PL2D21 thyratrons, T_1 to T_4 , are fed from

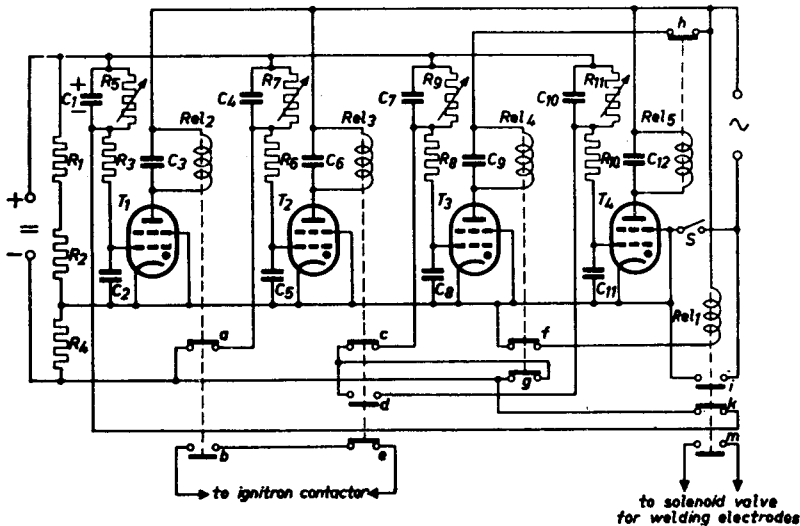


Fig. 12-15. Another timing circuit producing four time intervals.

an a.c. supply, and their control grids are fed from a d.c. source. The cathodes of the tubes receive a positive potential from a tapping on the voltage divider formed by resistors R_1 , R_2 and R_4 . In the condition shown in fig. 12-15 the control grids of T_1 , T_2 and T_3 are connected to the negative side of the d.c. supply via relay contacts k , a and c respectively, and these tubes are therefore non-conducting. Tube T_4 is, however, conducting, since the connection between its control grid and the negative side of the d.c. supply is interrupted by relay contact d .

To start a weld, switch S is closed, energising relay Rel_1 which holds itself

on by contact i . At the same time, contact m switches on the circuit of the solenoid valve which operates the welding electrodes, thus starting the "squeeze" time, and contact k opens to interrupt the connection between the grid of T_1 and the negative line. C_1 now discharges through R_5 , controlling the squeeze time. When the voltage drop across C_1 has fallen to a value approximately equal to that of the voltage across $R_1 + R_2$, tube T_1 ignites and energises Rel_2 . Contact b now closes the circuit which operates the ignitron contactor, thus starting the welding time, and contact a opens the connection between the control grid of T_2 and the negative line, allowing C_4 to discharge through R_7 , thereby controlling the welding time.

When T_2 ignites, Rel_3 is energised and interrupts the ignitron contactor circuit at contact e , thus completing the welding time. Relay Rel_3 also closes contact d , connecting the control grid of T_4 with the negative line so that T_4 is extinguished and C_{10} charges; finally contact c breaks the connection between the control grid of T_3 and the negative line, allowing C_7 to discharge through R_9 and thus control the hold time.

When the potential of the control grid of T_3 has reached a value approximately equal to that of the cathode, T_3 ignites, energising Rel_4 which opens contacts f and g . Contact f de-energises Rel_1 and the solenoid valve circuit is interrupted at contact m , thus causing the welding electrodes to open and the "off" time to commence. Moreover the closing of contact k extinguishes T_1 and de-energises Rel_2 , contact a of which closes and extinguishes T_2 . Relay Rel_3 is thus de-energised, but although contact c is thus closed, T_3 remains conducting since the connection between its control grid and the negative line is still interrupted at contact g . Contact d also interrupts the grid circuit of T_4 , allowing C_{10} to discharge through R_{11} ; the "off" time is governed in this way.

Finally, T_4 ignites and relay Rel_5 operates to open contact b , thus interrupting the anode circuit of T_3 . Relay Rel_4 now operates to close contacts f and g . If switch S is still closed Rel_1 is now energised and the whole process repeats. If S has been re-opened, however, Rel_1 remains unenergised and the circuit remains inoperative until S is again closed.

13. INDUSTRIAL RECTIFIER CIRCUITS

In industrial plants it often happens that a d.c. supply is needed for charging batteries, supplying d.c. motors or magnetic clutches, for arc lamps or for electro-chemical processes, but only an a.c. supply is available. Conversion of a.c. into d.c. is then necessary, and one of the following three methods must be selected:

- (1) By a rotary converter; i.e. an a.c. motor driving a d.c. generator.
- (2) By selenium rectifiers.
- (3) By thermionic tube rectifiers.

Rectification by means of thermionic tubes offers several important advantages compared with rotating converters. No foundations are required; the apparatus has no moving parts; maintenance is simpler; and operation is noiseless and without vibration. Compared with the selenium rectifier, rectifying tubes have the advantages of smaller dimensions, greater independency in temperature and in most cases higher efficiency. It is not surprising, therefore, that tube rectifiers

are widely used in industry, especially in view of the fact that modern rectifying tubes have an extremely long life (approximately 20 000 to 30 000 hours of operation).

Two-Phase Half-Wave Rectification

To simplify the calculations it will be assumed that the reactances of the transformer may be neglected (infinitely short commutation time) and that the mains and transformer voltages are purely sinusoidal. In fig. 13-1 the secondary voltage V_{tr} , the secondary current I_{tr} and the primary current I_p have been plotted as functions of time for a two-phase half-wave rectifier at a firing angle $\varphi_0 = 0$. The upper oscillogram is applicable to a purely resistive load, whereas the lower oscillogram is valid for a purely inductive load (infinitely large inductance in the cathode lead). In the latter case the current curves I_{tr} and I_p become rectangular.

If the arc voltage of the rectifying tubes is disregarded, the mean value of the rectified output voltage of an m -phase rectifier at full load (zero firing angle) is given by equ. (2.22) in Chapter 2. For $m = 2$ the value of M is 0.9. When the arc voltage V_{arc} cannot be neglected with respect to the transformer voltage, the output voltage is given by:

$$V_o = \frac{m}{2\pi} \int_{\frac{\pi}{2} - \frac{\pi}{m}}^{\frac{\pi}{2} + \frac{\pi}{m}} (\sqrt{2} \cdot V_{tr} \cdot \sin \varphi - V_{arc}) d\varphi = M \cdot V_{tr} - V_{arc}. \tag{13.1}$$

According to this expression the arc voltage must be subtracted from the output voltage as a constant, and it may therefore be disregarded in the following calculations for the sake of simplicity.

The r.m.s. value of the secondary current I_{tr} per phase can be calculated from the following general formula:

$$I_{tr} = \sqrt{\frac{1}{2\pi} \cdot \int_{\frac{\pi}{2} - \frac{\pi}{m}}^{\frac{\pi}{2} + \frac{\pi}{m}} i^2 \cdot d\varphi}, \tag{13.2}$$

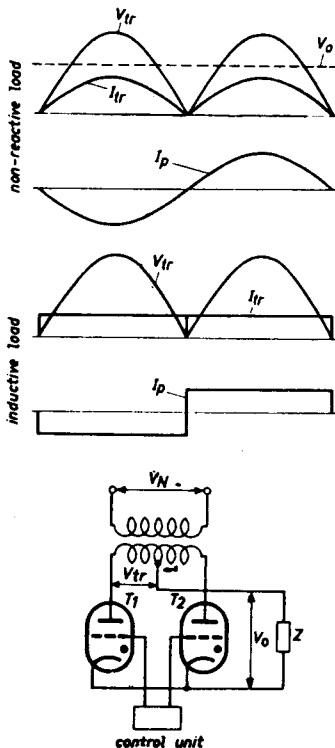


Fig. 13-1. Output voltage, output current and primary current of a two-phase, half-wave rectifier with a resistive and with an inductive load.

in which, when the load is purely resistive:

$$i = \hat{i} \sin \varphi. \quad (13.3)$$

Hence, for $m = 2$:

$$I_{tr} = \frac{1}{2} \hat{i} = \frac{1}{2} \cdot \frac{\pi}{2} \cdot I_o = 0.785 I_o, \quad (13.4)$$

in which I_o denotes the mean value of the rectified output current. When the load is inductive, then, according to fig. 13-1:

$$I_{tr} = \frac{1}{\sqrt{2}} \cdot I_o, \quad (13.5)$$

since I_{tr} now has a rectangular waveform and flows during intervals with a duration π .

The primary (mains) voltage V_N is related to the secondary voltage by the expression:

$$\frac{V_N}{V_{tr}} = n, \quad (13.6)$$

in which n denotes the transformer ratio.

The waveform of the primary current corresponds to that of the secondary current; with a resistive load this primary current is:

$$I_p = \frac{1}{n} \cdot \sqrt{2} \cdot I_{tr} = \frac{1}{n} \frac{\pi}{2\sqrt{2}} \cdot I_o, \quad (13.7)$$

which is to be attributed to the fact that, owing to the rectifying action of the tubes, the secondary current per phase can flow only during each second half cycle. This reveals the fundamental difference between a rectifier transformer and a normal current transformer, in which the transformation of the current is proportional to $1/n$.

When the load is inductive, then according to equ. (13.5):

$$I_p = \frac{1}{n} \sqrt{2} \cdot I_{tr} = \frac{1}{n} \cdot I_o. \quad (13.8)$$

In equ. (13.7) and (13.8) the magnetisation current of the transformer, which is usually less than 10% of the primary current, is not taken into account.

The rectified output power is obviously equal to $\mathcal{W}_o = I_o \cdot V_o$, whereas the apparent power of the transformer secondary is:

$$\mathcal{W}_{tr} = m \cdot I_{tr} \cdot V_{tr}. \quad (13.9)$$

which, in the case of a resistive load and for $m = 2$ gives:

$$\mathcal{W}_{tr} = 2 \cdot \frac{0.785}{0.9} \cdot I_o \cdot V_o = 1.74 \mathcal{W}_o, \quad (13.10)$$

and in the case of an inductive load:

$$\mathcal{W}_{tr} = \frac{2}{0.9 \cdot 1.41} \cdot I_o \cdot V_o = 1.57 \mathcal{W}_o. \quad (13.11)$$

The apparent power at the transformer primary is:

$$W_p = I_p \cdot V_N = \sqrt{2} \cdot \frac{0.785}{0.9} \cdot I_o V_o = 1.23 W_o \quad (13.12)$$

with a resistive load, and

$$W_p = \frac{\sqrt{2}}{0.9 \cdot 1.41} \cdot I_o V_o = 1.11 W_o \quad (13.13)$$

with an inductive load.

The nominal power of the transformer is determined by the arithmetic mean value of the apparent primary and secondary powers. Hence, in the case of a resistive load:

$$W_{(res)} = \frac{W_{tr} + W_p}{2} = 1.48 W_o \quad (13.14)$$

and in the case of an inductive load:

$$W_{(ind)} = 1.34 W_o \quad (13.15)$$

For designing a rectifier it is moreover important to know the quantities which determine the choice of the tubes, namely the peak current and the maximum peak inverse voltage.

In the case of a resistive load:

$$I_o = \frac{m}{2\pi} \cdot i \int_{\frac{\pi}{2} - \frac{\pi}{m}}^{\frac{\pi}{2} + \frac{\pi}{m}} \sin \varphi \, d\varphi, \quad (13.16)$$

which gives:

$$i = I_o \cdot \frac{\pi}{m \sin \pi/m}, \quad (13.17)$$

whence, for $m = 2$:

$$i = \frac{\pi}{2} \cdot I_o = 1.57 I_o. \quad (13.18)$$

In the case of an inductive load:

$$i = I_o. \quad (13.19)$$

According to what was said in Chapter 2, the inverse peak voltage per tube is:

$$V_{inv} = 2\sqrt{2} \cdot V_{tr} = 3.14 V_o. \quad (13.20)$$

Finally, the mean anode current per tube is generally given by:

$$I_a = \frac{I_o}{m}. \quad (13.21)$$

The application of the relations derived above will be illustrated by a simple example. It will be assumed that a two-phase rectifier is required to supply a current of 10 A at a voltage of 500 V at zero firing angle, the load being sub-

stantially inductive. According to equ. (13.19), (13.20) and (13.21) two thyatrons PL 105 can be used for this purpose. It follows from equ. (13.1) that the required transformer voltage V_{tr} per phase is $573 V_{rms}$. At $V_N = 220$ V, the transformer ratio n then becomes 1 : 2.6. According to equ. (13.5) the secondary current per phase is 7.1 A, the primary current being 26 A according to equ. (13.8); the latter value should be increased by 10% because of the magnetisation current. The apparent power at the transformer primary then amounts to $I_p \cdot V_N = 28,6 \times 220 \approx 6,3$ kVA, whereas the apparent power at the transformer secondary is approximately 8 kVA according to equ. (13.11). The nominal power of the transformer should therefore be 7.2 kVA.

Controlled two-phase half-wave Rectification

In the preceding section the firing angle of the rectifier was always assumed to be zero. The case will now be investigated of the firing angle φ_0 being greater than zero; this angle will be referred to the point of inter-

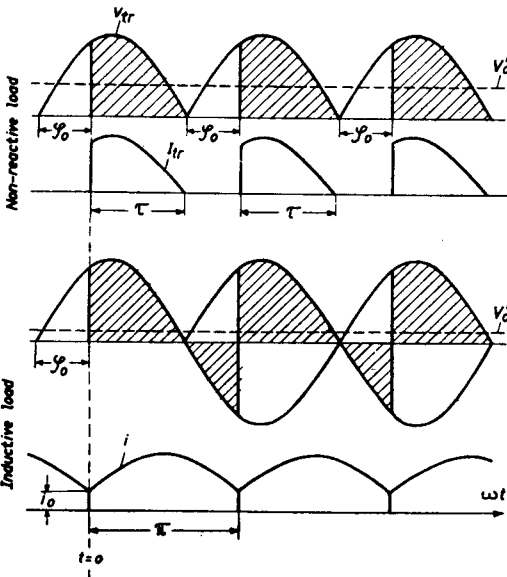


Fig. 13-2. Output voltage and output current of a controlled two-phase, half-wave rectifier with a resistive and with an inductive load.

section of two successive half waves of the transformer secondary voltage. In a two-phase rectifier this point coincides with the point of intersection on the abscissa. This condition is represented in fig. 13-2, both for a resistive and an inductive load, the firing angle being denoted by φ_0 . This figure reveals that in a two-phase rectifier the current flow becomes discontinuous when the load is resistive and φ_0 becomes greater than zero. When the load is purely inductive, this is no longer the case because the angle of current flow τ always remains equal to π at $\varphi_0 < 90^\circ$; this is because the tubes still pass current in a part of the range within which the secondary voltages are negative. The difference between the hatched areas above and below the zero axis is a measure of the voltage drop across the load. At $\varphi_0 = 90^\circ$ both areas obviously become equal, in other words the mean value of the output voltage V_o' becomes zero. Even without calculation it will be obvious that at a given firing angle the mean value of the output voltage V_o' is always lower with an inductive load than with a resistive load. In the latter case:

$$V_o' = \frac{\sqrt{2} V_{tr}}{\pi} (1 + \cos \varphi_0), \tag{13.22}$$

as derived in Chapter 3, whereas in the case of the purely inductive load:

$$V_o' = \frac{2\sqrt{2}}{\pi} \cdot V_{tr} \cos \varphi_0. \quad (13.23)$$

In fig. 13-3 the variation of V_o' has been plotted as a function of φ_0 for a purely inductive load (a) and for a resistive load (b). In the case of a mixed inductive and resistive load the function can be represented by a curve situated between (a) and (b) with a form similar to curve (c).

By way of example we will now consider the previously discussed two-phase rectifier

when the output voltage of 500 V is controlled. If the output voltage should be reduced to 250 V, the firing angle of the thyratrons must be $\varphi_0 = \cos^{-1} 0.5 = 60^\circ$ when the load is purely inductive. With a resistive load, however, the firing angle must be $\varphi_0 = 90^\circ$ for obtaining an output voltage of 250 V.

As shown by fig. 13-2, the current flow is continuous at a firing angle of $\varphi_0 < 90^\circ$ when the load is mainly inductive. The calculation of the current is based on the following general equation:

$$\sqrt{2} \cdot V_{tr} \cdot \sin(\omega t + \varphi_0) - Ri - L \cdot \frac{di}{dt} = 0, \quad (13.24)$$

which is applicable to the current circuit of a tube, provided its arc voltage is disregarded. Integration of equ. (13.24) gives:

$$i = \frac{\sqrt{2} \cdot V_{tr}}{\sqrt{R^2 + \omega^2 L^2}} \cdot \sin(\omega t + \varphi_0 - \Phi) + C \cdot e^{-tR/L}. \quad (13.25)$$

In this expression C represents an integration constant, whilst $\Phi = \tan^{-1} \omega L/R$ denotes the phase angle of a load. Since

$$\cos \Phi = \frac{R}{\sqrt{R^2 + \omega^2 L^2}},$$

it follows that in the case of $i = i_0$ at $t = 0$:

$$C = i_0 - \frac{\sqrt{2} \cdot V_{tr}}{R} \cdot \cos \Phi \cdot \sin(\varphi_0 - \Phi), \quad (13.26)$$

and

$$i = \frac{\sqrt{2} \cdot V_{tr}}{R} \cdot \cos \Phi \cdot \sin(\omega t + \varphi_0 - \Phi) + \left\{ i_0 - \frac{\sqrt{2} \cdot V_{tr}}{R} \cdot \cos \Phi \cdot \sin(\varphi_0 - \Phi) \right\} \cdot e^{-tR/L}. \quad (13.27)$$

According to fig. 13-2 it may now be assumed that i is again equal to i_0 at $\omega t = \pi$, in other words that the current curve of the one phase is linked up with that of the other phase. Hence:

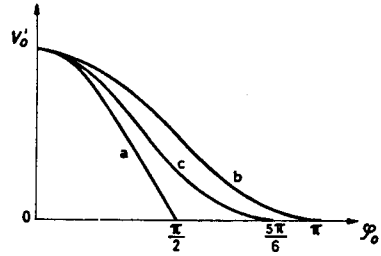


Fig. 13-3. V_o' as a function of φ_0 for (a) two- and more-phase rectifiers with an inductive load, (b) two-phase rectifiers with a resistive load, and (c) three-phase rectifiers with a resistive load.

$$i_o = \frac{\sqrt{2} \cdot V_{tr} \cdot \cos \Phi \cdot \sin(\Phi - \varphi_o) - [\sin(\varphi_o - \Phi)] e^{-\frac{R \pi}{\omega L}}}{R \cdot (1 - e^{-\frac{R \pi}{\omega L}})} \quad (13.28)$$

The complete formula for the current i is obtained by substituting this expression in equ. (13.27). The mean value of the direct current, I_o , can now be evaluated from the expression:

$$I_o = \frac{1}{\pi} \int_0^\pi i \cdot d\omega t, \quad (13.29)$$

which gives:

$$I_o = \frac{2 \sqrt{2} \cdot V_{tr} \cdot \cos \Phi}{\pi R} \cdot \left\{ \cos(\Phi - \varphi_o) + \frac{\omega L}{R} \cdot \sin(\Phi - \varphi_o) \right\}. \quad (13.30)$$

In view of the simplifying assumptions this expression is valid only in the case where the current flow is continuous, that is in the case of $\Phi \geq \varphi_o$.

The significance of equ. (13.30) will be illustrated by means of the example used previously, it being assumed that the load of the two-phase, controlled rectifier has a resistance of $R = 50 \Omega$. If the arc voltage of the thyatron is

disregarded, the r.m.s. value of the secondary voltage is $V_{tr} = 555$ V. The mean value of the direct current I_o appears to be 10 A at zero firing angle, irrespective of the value of the power factor. At $\Phi = 60^\circ$, i.e. $\cos \Phi = 0.5$ and $\varphi_o = 30^\circ$, I_o becomes 8.66 A, whereas at $\varphi_o = 60^\circ$, $I_o = 5$ A. This appears to be the lowest limit of the output current at which this is still continuous. If φ_o is further increased, the current will consist of discontinuous pulses and equ. (11.30) is no

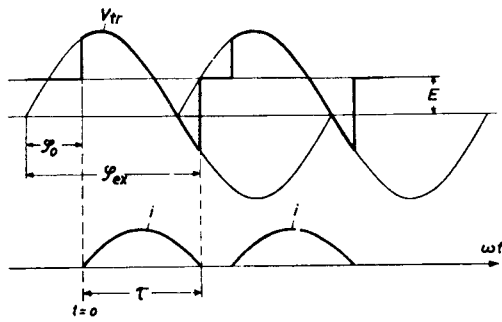


Fig. 13-4. Output voltage and output current of a controlled two-phase rectifier with a counter-e.m.f. in its load circuit.

longer applicable.

In electronic motor control devices the load of the rectifier consists of the armature with, as a rule, a smoothing choke connected in series. This is therefore again a mixed inductive and resistive load, but the load circuit also includes an e.m.f. E which is proportional to the speed of the motor (see fig. 13-4). Disregarding once again the arc voltage of the tubes, the mean value of the output voltage as a function of the firing angle is given by:

$$V_o' = \frac{1}{\pi} \int_{\varphi_o}^{\varphi_{ex}} (\sqrt{2} \cdot V_{tr} \cdot \sin \omega t - E) d\omega t, \quad (13.31)$$

in which φ_o represents the firing angle and φ_{ex} the extinction angle of the tubes. By introducing the voltage ratio

$$a = \frac{E}{\sqrt{2} \cdot V_{tr}} \quad (13.32)$$

integration of equ. (13.31) may be written:

$$V_o' = \frac{\sqrt{2} \cdot V_{tr}}{\pi} \cdot \left\{ \cos \varphi_0 - \cos \varphi_{ex} - a (\varphi_{ex} - \varphi_0) \right\}. \quad (13.33)$$

In a two-phase rectifier with a mixed inductive and resistive load and a counter-e.m.f. the current will not as a rule flow continuously, and this case will now be investigated more closely. According to fig. 13-4 the equation of the current circuit of one tube (arc voltage drop again neglected) is:

$$\sqrt{2} \cdot V_{tr} \cdot \sin (\omega t + \varphi_0) - E - Ri - L \cdot \frac{di}{dt} = 0. \quad (13.34)$$

The solution of this equation is:

$$i = \frac{\sqrt{2} \cdot V_{tr}}{\sqrt{R^2 + \omega^2 L^2}} \cdot \sin (\omega t + \varphi_0 - \Phi) - \frac{E}{R} + C \cdot e^{-tR/L}. \quad (13.35)$$

The value of C can be determined by making use of the fact that the current is now discontinuous, which means that $i = 0$ at $t = 0$ (cf. fig. 13-4). Hence:

$$C = -\frac{\sqrt{2} \cdot V_{tr}}{R} \cdot \cos \Phi \cdot \sin (\varphi_0 - \Phi) + \frac{E}{R}, \quad (13.36)$$

so that the complete expression for i becomes:

$$i = \frac{\sqrt{2} \cdot V_{tr}}{R} \cdot \left[\cos \Phi \cdot \sin (\omega t + \varphi_0 - \Phi) - a + \left\{ a - \cos \Phi \cdot \sin (\varphi_0 - \Phi) \right\} \cdot e^{-\omega t / \tan \Phi} \right]. \quad (13.37)$$

The current function thus consists of a periodical term and of a term which decreases exponentially with time. Fig. 13-4 reveals that the current flow is now discontinuous. A comparison with fig. 13-2 shows that this fact is to be attributed to the presence of the counter-e.m.f. The duration of the current pulse, denoted by the angle τ , will increase as E decreases, in other words as the voltage ratio a becomes smaller. The value of τ can be determined by putting equ. (13.37) equal to zero, i.e.:

$$a - \cos \Phi \cdot \sin (\tau + \varphi_0 - \Phi) = \left\{ a - \cos \Phi \cdot \sin (\varphi_0 - \Phi) \right\} \cdot e^{-\tau / \tan \Phi}. \quad (13.38)$$

this equation can be written in a symmetrical form by introducing the angle $\varphi_{ex} = \varphi_0 + \tau$:

$$\left\{ a - \cos \Phi \cdot \sin (\varphi_{ex} - \Phi) \right\} \cdot e^{\varphi_{ex} / \tan \Phi} = \left\{ a - \cos \Phi \cdot \sin (\varphi_0 - \Phi) \right\} \cdot e^{\varphi_0 / \tan \Phi}. \quad (13.39)$$

A lower limit is obviously set to the firing angle φ_0 because the thyatron cannot be fired until the instantaneous value of V_{tr} exceeds the sum of the

counter-e.m.f. and the arc voltage. If the latter is again neglected, it may thus be written:

$$\varphi_0 \geq \sin^{-1} a. \quad (13.40)$$

The equations derived will again be illustrated by means of the example mentioned above, it being assumed that the load circuit of the two-phase, controlled rectifier contains a counter-e.m.f. of 400 V, which gives:

$$a = \frac{400}{\sqrt{2} \cdot 555} \approx 0.5. \quad (13.41)$$

According to equ. (13.40) the smallest firing angle φ_0 is now 30° (at full load). It will be assumed that $\varphi_0 = 60^\circ$; the angle φ_{ex} is a function of the power factor of the load, and can be evaluated by means of equ. (13.39). Assuming $\Phi = 60^\circ$, then $\cos \Phi = 0.5$, which gives $\varphi_{ex} = 192^\circ$. It can then be calculated from equ. (13.33) that V_o' is approximately 80 V.

The peak value of the current can be calculated by means of equ. (13.37), which gives the instantaneous value of the current, by setting the first derivative of i equal to zero. The mean value of the current I_o can be calculated from equ. (13.29) by integrating between 0 and $\varphi_{ex} - \varphi_0$, that is over the duration τ of the current pulse. The actual evaluation of this integral is rather complicated and need not be dealt with here.

The comments on two-phase, controlled rectifiers will be concluded by investigating the case of a resistive load and a counter e.m.f. included in the load circuit, which applies for example to rectifiers for charging batteries.

The calculation will be based on the general equ. (13.33), in which it should be assumed that:

$$\varphi_{ex} = \pi - \sin^{-1} a \quad (13.42)$$

because the tubes cannot pass current beyond this point. Hence, for $m = 2$:

$$V_o' = \frac{\sqrt{2} \cdot V_{tr}}{\pi} \cdot \left\{ \cos \varphi_0 + \sqrt{1 - a^2} - a \cdot (\pi - \sin^{-1} a - \varphi_0) \right\}. \quad (13.43)$$

At full load of the rectifier, φ_0 must be put equal to $\sin^{-1} a$, which gives:

$$V_o = \frac{\sqrt{2} \cdot 2 V_{tr}}{\pi} (\sqrt{1 - a^2} - a \cdot \cos^{-1} a). \quad (13.44)$$

The function

$$\mathcal{L} = \sqrt{1 - a^2} - a \cdot \cos^{-1} a \quad (13.45)$$

is already known and has been plotted in fig. 2-10.

The calculation of the current can be based on equ. (13.34), L being put equal to zero. The mean value of the current can be evaluated by carrying out the integration from 0 to $\pi - \sin^{-1} a - \varphi_0$:

$$I_o = \frac{\sqrt{2} \cdot V_{tr}}{R \cdot \pi} \int_0^{\pi - \sin^{-1} a - \varphi_0} \left\{ \sin(\omega t + \varphi_0) - a \right\} d\omega t, \quad (13.46)$$

which gives:

$$I_o = \frac{\sqrt{2} \cdot V_{tr}}{R \cdot \pi} \cdot \left\{ \cos \varphi_0 + \sqrt{1 - a^2} - (\pi - \sin^{-1} a - \varphi_0) a \right\}. \quad (13.47)$$

In the case of the previously used example, a being again 0.5, the output voltage V_o is equal to 170 V at full load, according to equ. (13.44). At a firing angle of $\varphi_0 = 60^\circ$, however, $V_o' = 125$ V according to equ. (13.43). At $R = 50 \Omega$, the output current $I_o = 2.5$ A according to equ. (13.47).

Three-Phase Rectification

The investigation of three-phase rectifiers will be based on the comments given above. The frequently used delta-Y circuit is represented in fig. 13-5. When the load is resistive the output voltage and current consist of consecutive sinusoidal sections of duration $2\pi/3$ at full load. With an inductive load the secondary current assumes a rectangular waveform. The mean value of the output voltage of a three-phase rectifier is, according to equ. (2.22):

$$V_o = \sqrt{2} \cdot \frac{3}{\pi} \sin \frac{\pi}{3} V_{tr} = 1.17 V_{tr} \quad (13.48)$$

provided the simplifying assumptions made are permissible. With a resistive load the r.m.s. value of the secondary current per phase is given by the formula:

$$I_{tr} = \sqrt{\frac{1}{2\pi} \cdot \int_{30^\circ}^{150^\circ} i^2 \sin^2 \varphi d\varphi} = 0.485 i. \quad (13.49)$$

The mean value of the direct current is then:

$$I_o = \frac{3}{2\pi} \cdot \int_{30^\circ}^{150^\circ} i \cdot \sin \varphi d\varphi = \frac{3\sqrt{3}}{2\pi} \cdot i = 0.827 i. \quad (13.50)$$

When the load is inductive, which will usually be the case in three-phase rectification, the waveform of the primary current will be rectangular. According to equ. (13.2):

$$I_{tr} = \frac{1}{\sqrt{3}} \cdot I_o = 0.58 I_o. \quad (13.51)$$

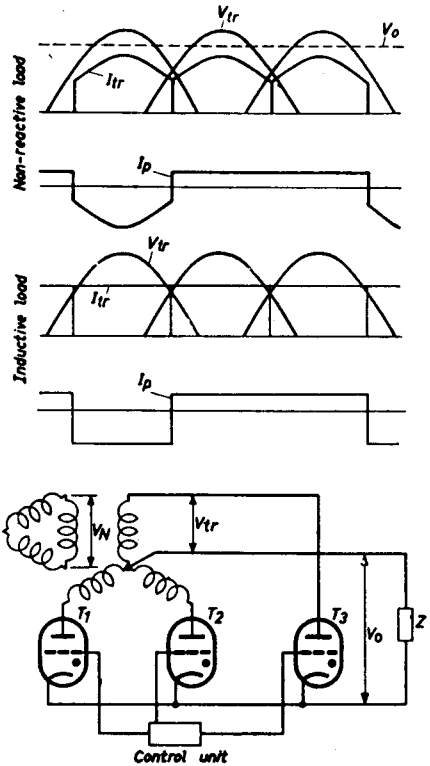


Fig. 13-5. Three-phase rectifier in delta-Y connection.

The transformer ratio may again be:

$$\frac{V_N}{V_{tr}} = n. \tag{13.6}$$

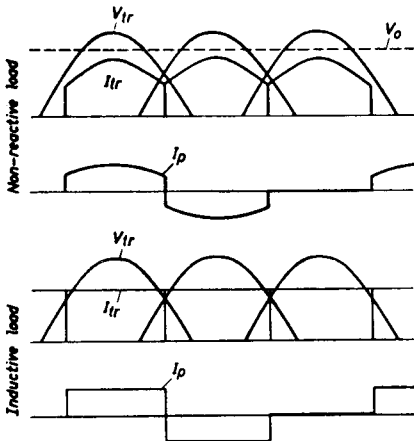
The waveform of the primary current of each phase is also rectangular, but the peak value of the first half cycle is twice that of the two subsequent half cycles. Hence:

$$\hat{I}_{p(1)} = \frac{2}{3} \cdot \frac{1}{n} \cdot I_o, \tag{13.52}$$

and

$$\hat{I}_{p(2)} = \frac{1}{3} \cdot \frac{1}{n} \cdot I_o. \tag{13.53}$$

The r.m.s. value of the primary current is therefore:



$$I_p = \frac{\sqrt{2}}{3} \cdot \frac{1}{n} \cdot I_o = 0.47 \cdot \frac{1}{n} \cdot I_o, \tag{13.54}$$

and

$$I_p = \sqrt{\frac{2}{3}} \cdot \frac{1}{n} \cdot I_{tr}. \tag{13.55}$$

The apparent power at the primary of the transformer is:

$$W_p = 3 V_N \cdot I_p = 1.21 W_o, \tag{13.56}$$

and that at the secondary of the transformer:

$$W_{tr} = 1.48 W_o, \tag{13.57}$$

which gives for the nominal power of the transformer:

$$W = 1.35 W_o. \tag{13.58}$$

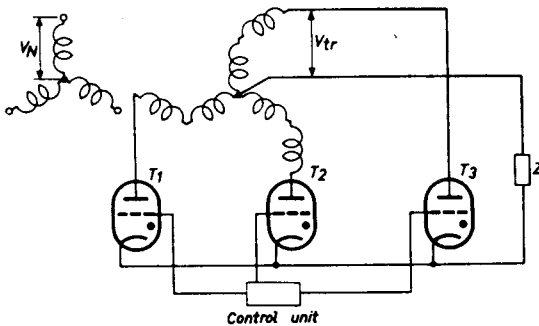


Fig. 13-6. Three-phase rectifier in Y-zigzag connection.

phase, half-wave rectification and for six-phase, half-wave rectification with interphase transformer (double Y-circuit) are also indicated.

	Three-phase, delta-Y half-wave rect.)	Three-phase, Y-zigzag (half-wave rect.)	Three-phase (full-wave rect.)	Six-phase, delta-Y (half-wave rect.)	Six-phase, double Y (half-wave rect.)
V_o	1.17 V_{tr}	1.17 V_{tr}	2.34 V_{tr}	1.35 V_{tr}	1.17 V_{tr}
I_{tr}	0.58 I_o	0.58 I_o	0.82 I_o	0.41 I_o	0.29 I_o
I_p	0.47 I_o/n	0.47 I_o/n	0.82 I_o/n	0.58 I_o/n	0.41 I_o/n
\mathcal{W}_p	1.21 \mathcal{W}_o	1.21 \mathcal{W}_o	1.05 \mathcal{W}_o	1.28 \mathcal{W}_o	1.05 \mathcal{W}_o
\mathcal{W}_{tr}	1.48 \mathcal{W}_o	1.71 \mathcal{W}_o	1.05 \mathcal{W}_o	1.81 \mathcal{W}_o	1.48 \mathcal{W}_o
\mathcal{W}	1.35 \mathcal{W}_o	1.46 \mathcal{W}_o	1.05 \mathcal{W}_o	1.55 \mathcal{W}_o	1.26 \mathcal{W}_o
I_o	3 I_a	3 I_a	3 I_a	6 I_a	6 I_a

Three-Phase Controlled Rectification

If the arc voltage is again neglected, the mean value of the direct voltage of a three-phase controlled rectifier with a resistive load is given, according to equ. (3.2), by the expression:

$$V_o' = \frac{\sqrt{2} \cdot 3}{\pi} \cdot \sin \frac{\pi}{3} \cdot V_{tr} \cdot \cos \varphi_0 = 1,17 V_{tr} \cdot \cos \varphi_0, \quad (13.59)$$

in which φ_0 denotes the firing angle. This expression is only valid when the period during which each anode passes current is not shortened, i.e. when:

$$\tau = \frac{2\pi}{3}. \quad (13.60)$$

This means that the limitation $\varphi_0 \leq \bar{\varphi}$ is imposed, where:

$$\bar{\varphi} = \frac{\pi}{2} - \frac{\pi}{3} = \frac{\pi}{6}. \quad (13.61)$$

When φ_0 exceeds $\bar{\varphi}$ the current per anode flows during a shorter period, given by the angle:

$$\tau' = \frac{\pi}{2} + \frac{\pi}{3} - \varphi_0 = \frac{5\pi}{6} - \varphi_0, \quad (13.62)$$

and the output voltage of the controlled rectifier becomes:

$$V_o' = \sqrt{2} \cdot \frac{3}{2\pi} \cdot V_{tr} \left\{ 1 - \sin \left(\varphi_0 - \frac{\pi}{3} \right) \right\}. \quad (13.63)$$

In rectification with three or more phases there is always a maximum firing angle $\bar{\varphi}$ within which the current flows continuously; at $m = 3$ this angle is, according to equ. (13.61), equal to 30° . Curve c of fig. 13-3 represents the output voltage of a three-phase rectifier, with a resistive load, as a function of the firing angle φ_0 .

With an inductive load the angle τ increases for each anode so that the current will flow continuously even at larger values of the firing angle. Since the secondary voltage becomes temporarily negative, the output voltage decreases. In that case equ. (3.2) is applicable, in other words V_o' becomes zero at $\varphi_0 = 90^\circ$, irrespective

of the number of phases. The output voltage V_o' as a function of the firing angle is then given by curve a of fig. 13-3.

When a counter-e.m.f. is included in the load circuit, calculations can be based on the general formula:

$$V_o' = \frac{m}{2\pi} \int_{\pi/2 - \pi/m + \varphi_0}^{\pi/2 - \pi/m + \varphi_{ex}} (\sqrt{2} \cdot V_{tr} \cdot \sin \varphi - E) \cdot d\varphi. \quad (13.64)$$

Provided the current flows continuously, it may be written:

$$\varphi_{ex} = \varphi_0 + \frac{2\pi}{m}, \quad (13.65)$$

which gives by integration:

$$V_o' = \sqrt{2} \cdot V_{tr} \left(\frac{m}{\pi} \cdot \sin \frac{\pi}{m} \cdot \cos \varphi_0 - a \right), \quad (13.66)$$

whereas in the case of the current being discontinuous:

$$V_o' = \frac{mV_{tr}}{\pi\sqrt{2}} \cdot \left[\sin \left(\frac{\pi}{m} - \varphi_0 \right) - \sin \left(\frac{\pi}{m} - \varphi_{ex} \right) - a \left(\varphi_{ex} - \varphi_0 \right) \right]. \quad (13.67)$$

Calculation of the Value of the Smoothing Choke

In designing multiphase rectifiers it is often desired to determine the inductance in the load circuit at which — for a certain firing angle — the current still flows continuously. To simplify the calculation the output voltage of an m -phase controlled rectifier may to a first approximation be put equal to the sum of the mean direct voltage V_o' and an alternating voltage the amplitude V_{\sim} of which is equal to the difference between the peak value of the secondary transformer voltage V_{tr} and the mean value of the direct voltage V_o' , and the frequency of which is equal to m times the mains frequency. Hence:

$$v = V_o' + V_{\sim} \cdot \sin m\omega t = V_o' + \sqrt{2} \cdot V_{tr} \cdot \left(1 - \frac{m}{\pi} \cdot \sin \frac{\pi}{m} \cdot \cos \varphi_0 \right) \cdot \sin m\omega t. \quad (13.68)$$

The load circuit can be described by the expression:

$$V_o' + \sqrt{2} \cdot V_{tr} \left(1 - \frac{m}{\pi} \cdot \sin \frac{\pi}{m} \cdot \cos \varphi_0 \right) \cdot \sin m\omega t = iR + L \cdot \frac{di}{dt} + E, \quad (13.69)$$

which gives:

$$i = I_o + I_{\sim} \cdot \sin (m\omega t - \Phi), \quad (13.70)$$

in which:

$$\Phi = \tan^{-1} \frac{m\omega L}{R}. \quad (13.71)$$

The current may thus also be considered as the sum of a d.c. component and an a.c. component with an amplitude I_{\sim} . In the boundary case of the current still just being continuous, $I_{\sim} = I_o$, that is:

$$i = I_o \{ 1 + \sin (m\omega t - \Phi) \}. \quad (13.72)$$

The d.c. component and the a.c. component must both satisfy equ. (13.69) individually, whence:

$$V_o' = I_o R + E, \quad (13.73)$$

and

$$\sqrt{2} \cdot V_{tr} \left(1 - \frac{m}{\pi} \cdot \sin \frac{\pi}{m} \cdot \cos \varphi_0 \right) = I_o \cdot \sqrt{R^2 + m^2 \omega^2 L^2}. \quad (13.74)$$

Disregarding R^2 with respect to $m^2 \omega^2 L^2$, this means that the lower limit of L , as a function of the firing angle φ_0 or of the lowest value of the current I_o that still flows continuously, is given by:

$$L = \frac{\sqrt{2} \cdot V_{tr} \left(1 - \frac{m}{\pi} \cdot \sin \frac{\pi}{m} \cdot \cos \varphi_0 \right)}{m \omega I_o}. \quad (13.75)$$

Influence of the Grid Control

For *vertical* control an alternating voltage is applied to the grids of the tubes that lags 90° with respect to the anode voltage and is superimposed on a direct voltage which may vary between positive and negative values. Denoting the variable direct voltage by v_s and the amplitude of the phase-shifted alternating voltage by V_s , we may write:

$$v_s = V_s \cos \varphi_0, \quad (13.76)$$

if the ignition characteristic of the thyatron is assumed to coincide with the zero line (see fig. 13-7). The firing angle will vary from 0 to 180° when v_s is varied from $+V_s$ to $-V_s$. Substitution of the ratio v_s/V_s for $\cos \varphi_0$ in equ. (13.22) and (13.23), which represent the output voltage of a two-phase rectifier with a resistive and an inductive load respectively, gives:

$$V_o' = \frac{\sqrt{2} \cdot V_{tr}}{\pi} \cdot \left(1 + \frac{v_s}{V_s} \right), \quad (13.77)$$

and

$$V_o' = \frac{2\sqrt{2}}{\pi} \cdot V_{tr} \cdot \frac{v_s}{V_s}, \quad (13.78)$$

respectively, in other words the output voltage becomes a linear function of the control voltage v_s . These functions have been plotted in fig. 13-8 for an inductive load (a) and for a resistive load (b).

The output voltage of m -phase rectifiers with an inductive and a resistive load and continuously flowing current was given by equ. (3.2), in which the firing angle was, however, calculated with respect to the point of intersection of two consecutive half sine waves, whereas the firing angle φ_0 in equ. (13.76) is

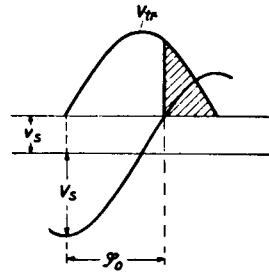


Fig. 13-7. Vertical control.

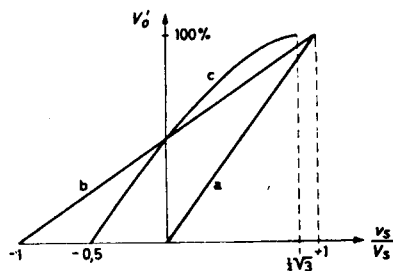


Fig. 13-8. Control characteristics (a) of a two-phase rectifier with an inductive load, (b) of a two-phase rectifier with a resistive load, and (c) of a three-phase rectifier with an inductive load.

related to the point of intersection of a half sine wave with the abscissa. It will therefore be useful to introduce the angle:

$$a = \frac{\pi}{2} - \frac{\pi}{m} + \varphi_0 \tag{13.79}$$

so that now:

$$\cos a = \frac{v_s}{V_s} \tag{13.80}$$

Equ. (3.2) thus becomes:

$$V_o' = MV_{tr} \cos \left(\alpha - \frac{\pi}{2} + \frac{\pi}{m} \right) = MV_{tr} \left\{ \frac{v_s}{V_s} \cdot \cos \left(\frac{\pi}{2} - \frac{\pi}{m} \right) + \sqrt{1 - \frac{v_s^2}{V_s^2}} \cdot \sin \left(\frac{\pi}{2} - \frac{\pi}{m} \right) \right\} \tag{13.81}$$

It is seen that for $m > 2$ the relation between the control voltage and the output voltage is no longer linear. For v_s/V_s the following limitation holds:

$$\frac{\pi}{2} - \frac{\pi}{m} \leq \cos^{-1} \frac{v_s}{V_s} \leq \pi - \frac{\pi}{m} \tag{13.82}$$

provided the current remains continuous. At $m = 3$ the control characteristic of the rectifier is represented by curve c of fig. 13-8, limited by

$$\frac{1}{2} \sqrt{3} \geq \frac{v_s}{V_s} \geq -\frac{1}{2}$$

corresponding to the limits of 30° and 120° for a .

The simplest method of obtaining *horizontal* control consists in applying to the grid an alternating voltage which lags by a certain angle α with respect to the secondary transformer voltage V_{tr} . This grid voltage is usually taken from a phase-shifting network consisting of a resistance R and a reactance ωL or $1/\omega C$. According to Chapter 3 the phase shift is given by

$$\alpha = 2 \tan^{-1} \omega RC, \tag{13.83}$$

in the case of an RC -network, and by

$$\alpha = 2 \tan^{-1} \frac{\omega L}{R}, \tag{13.84}$$

in the case of an RI -network.

The latter case will be investigated first. It may then be written:

$$\cos \frac{\alpha}{2} = \frac{R}{\sqrt{R^2 + \omega^2 L^2}}, \tag{13.85}$$

and

$$\cos \alpha = \frac{2 R^2}{R^2 + \omega^2 L^2} - 1. \tag{13.86}$$

In two-phase rectifiers α may be set equal to φ_0 ; it then follows from equ. (13.22) that with a resistive load:

$$V_o' = \frac{\sqrt{2} \cdot V_{tr}}{\pi} \cdot \frac{2 R^2}{R^2 + \omega^2 L^2} \quad (13.87)$$

and from equ. (13.23) that with an inductive load:

$$V_o' = \frac{2\sqrt{2}}{\pi} \cdot V_{tr} \cdot \left(\frac{2 R^2}{R^2 + \omega^2 L^2} - 1 \right). \quad (13.88)$$

It is thus seen that for controlling the output voltage from zero to its full value with a resistive load, R or L would have to be varied over a range from zero to infinity; with an inductive load, however, it is sufficient to vary L over a range from R/ω to zero.

For RC -networks:

$$\cos \alpha = \frac{2}{\omega^2 C^2 R^2 + 1} - 1, \quad (13.89)$$

which gives

$$V_o' = \frac{\sqrt{2} \cdot V_{tr}}{\pi} \cdot \frac{2}{\omega^2 C^2 R^2 + 1}, \quad (13.90)$$

for a resistive load, and

$$V_o' = \frac{2\sqrt{2}}{\pi} \cdot V_{tr} \cdot \left(\frac{2}{\omega^2 C^2 R^2 + 1} - 1 \right) \quad (13.91)$$

for an inductive load.

In this case, too, it appears to be necessary to vary R or C from zero to infinity when the load is resistive; with an inductive load it is sufficient to vary R from zero to $1/\omega C$ or to vary C from zero to $1/\omega R$.

In m -phase rectifiers supplying a continuous current, equ. (3.2) is applicable, and when an RL -network is used we may then write:

$$V_o' = M V_{tr} \left\{ \left(\frac{2 R^2}{R^2 + \omega^2 L^2} - 1 \right) \cos \left(\frac{\pi}{2} - \frac{\pi}{m} \right) + \sqrt{1 - \left(\frac{2 R^2}{R^2 + \omega^2 L^2} - 1 \right)^2} \cdot \sin \left(\frac{\pi}{2} - \frac{\pi}{m} \right) \right\}. \quad (13.92)$$

The limits of the control range are then:

$$\frac{\pi}{2} - \frac{\pi}{m} \leq \cos^{-1} \left(\frac{2 R^2}{R^2 + \omega^2 L^2} - 1 \right) \leq \pi - \frac{\pi}{m} \quad (13.93)$$

at $m = 3$, and in the case of $V_o' = V_o$:

$$\frac{2 R^2}{R^2 + \omega^2 L^2} - 1 = \cos \frac{\pi}{6} = \frac{1}{2} \sqrt{3}, \quad (13.94)$$

whereas in the case of $V_o' = 0$:

$$\frac{2 R^2}{R^2 + \omega^2 L^2} - 1 = \cos \frac{2}{3} \pi = -0.5. \quad (13.95)$$

From equ. (13.94):

$$R = 3.74 \omega L \tag{13.96}$$

for $V_o' = V_o$, whereas from equ. (13.95):

$$R = 0.577 \omega L, \tag{13.97}$$

for $V_o' = 0$. To control a three-phase rectifier from zero to its full output power it is therefore necessary to vary R over a range of approximately 1 : 6.5.

If an RC -network is used, the limits of the control range are, according to equ. (13.89):

$$\frac{\pi}{2} - \frac{\pi}{m} \leq \cos^{-1} \left(\frac{2}{\omega^2 C^2 R^2 + 1} - 1 \right) \leq \pi - \frac{\pi}{m} \tag{13.98}$$

Hence, at $m = 3$:

$$\omega RC = 0.265 \tag{13.99}$$

for $V_o' = V_o$, and

$$\omega RC = \sqrt{3} = 1.73 \tag{13.100}$$

for $V_o' = 0$.

Simple Two-Phase Battery Charger

The circuit of a simple two-phase rectifier for charging four 6-volt motor car batteries at a current of 6 amperes is shown in fig. 13-9. A Philips 367 rectifying tube is used (see fig. 13-10). As already explained in Chapter 2, the instantaneous value of the current in a rectifying circuit is determined by the difference between the instantaneous value of input voltage and the sum of

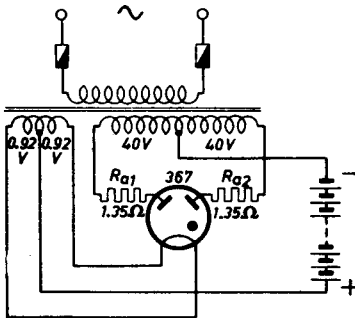


Fig. 13-9. Two-phase rectifier circuit for charging batteries.

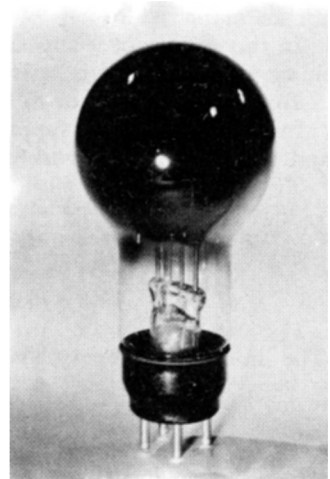


Fig. 13-10. Type 367 rectifying tube.

the battery and the arc voltage of the tube, divided by the total resistance in the circuit (see equation (2.3)). Since the resistance presented by the transformer secondary is usually so small that the maximum permissible tube current would be greatly exceeded, it is necessary in practice to insert in the circuit an additional

current limiting resistor or a choke. Assuming the transformer resistance to be about 10% of the total resistance in the rectifying circuit, the value of the anode resistors R_{a1} , R_{a2} to be inserted in the two anode leads, calculated according to equation (2.9) will be 1.35 ohm each.

The power losses in anode resistors for larger charging rectifiers would be so great that the efficiency of the device would be seriously reduced. A current limiting device which dissipates very little power is therefore preferred, and can take the form of a choke or a transformer with sufficiently high stray inductance. The choke should be inserted in the transformer primary circuit in order to avoid premagnetisation and thus reduction of inductance by the current flowing in the secondary. The choke should be so designed that the voltage drop is about 10% of the mains voltage under no-load conditions. If a load is present at the transformer secondary the primary current increases, and the voltage drop across the choke therefore increases also. Thus the voltage across the transformer primary and consequently the secondary voltage and the charging current are reduced.

These effects are illustrated in fig. 13-11 where the voltage and current occurring in the primary circuit of a two-phase rectifier are shown. Under no-load conditions the secondary voltage V_2 of the transformer is sinusoidal. Assuming the voltage drop across the choke inserted in the primary circuit to be 10% of the mains voltage, the ratio of secondary to primary voltage of the transformer becomes (see Chapter 2):

$$\frac{V_2}{V_1} = \frac{V_2}{0.9 V_N} = \frac{V_o + V_{arc}}{0.9 V_N a \sqrt{2}}, \quad (13.101)$$

V_1 being the primary voltage of the transformer and V_N the mains voltage. If current flows in the rectifying circuit, the secondary voltage (assuming the resistance of the rectifying circuit to be zero) will be approximately:

$$V_2' = V_o + V_{arc}, \quad (13.102)$$

V_2' being independent of the momentary value of the mains voltage. Obviously the same is valid for the primary voltage V_1' (see fig. 13-11), the difference between this rectangular voltage shape and the sinusoidal mains voltage being taken by the choke. V_1' may be written according to equation (13.101):

$$V_1' = 0.9 V_N a \sqrt{2} = V_1 a \sqrt{2}. \quad (13.103)$$

The primary transformer voltage therefore decreases more as the voltage ratio a becomes smaller, i.e., as V_o becomes smaller and the secondary current becomes larger.

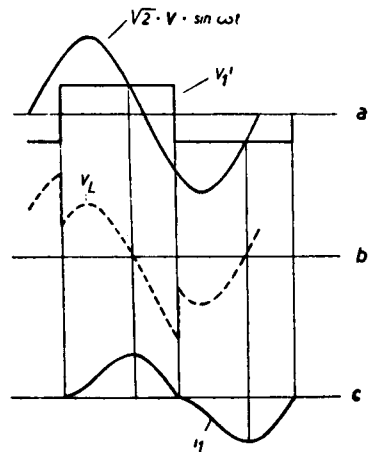


Fig. 13-11. Voltage and current waves occurring in the primary of a two-phase rectifying circuit.

A Larger Battery-Charging Equipment

The circuit of a rectifier which permits charging of 36 lead cells at a current of 50 amperes is shown in fig. 13-12. It is a four-phase rectifier using a Scott transformer fed from the three-phase mains. The circuit corresponds to the basic circuit No. 6 of fig. 2-16.

The two-anode rectifying tube, Philips Type 1749 A, has an auxiliary electrode which must be connected with one side of the filament via a resistor of about 10 000 ohms when low voltage ignition is desired. This connection is already made by the tube manufacturer. As the tubes need a preheating time of about

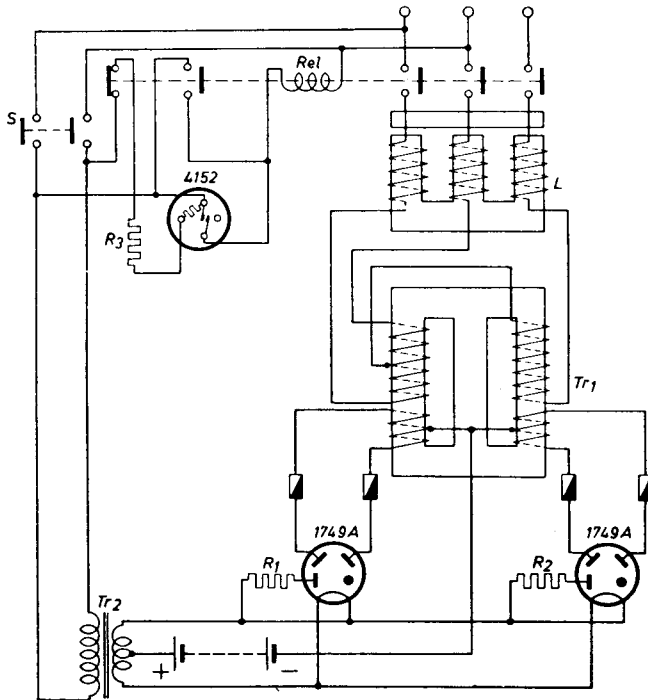


Fig. 13-12. Four-phase rectifying circuit.

$1\frac{1}{2}$ minutes before the anode voltage is applied, a bi-metal relay (Philips 4152) is employed. This device closes a circuit after the required time interval, thus energising relay *Rel* which switches on the main circuit. The choke *L* inserted in the primary circuit has either a variable air gap or winding taps in order to adjust the charging current accurately to its maximum value.

Rectifier with Constant Output Voltage

The circuit of a three-phase rectifier whose output voltage is held constant automatically is shown in fig. 13-12. The anodes of the three thyatron tubes Th_1 , Th_2 , Th_3 are connected to a three-phase power transformer Tr_1 , and the grid control of the tubes consists of three alternating voltages derived from transformer Tr_2 , and retarded in phase by 90 degrees with respect to the cor-

responding anode voltages, and a direct voltage variable between positive and negative values. This is clearly a "vertical" control of the type already described in Chapter 3 (see fig. 3-14). The variable direct voltage is taken from a bridge circuit consisting of the centre-tapped secondary winding of transformer Tr_3 , resistor R_4 and high-vacuum tube V_2 .

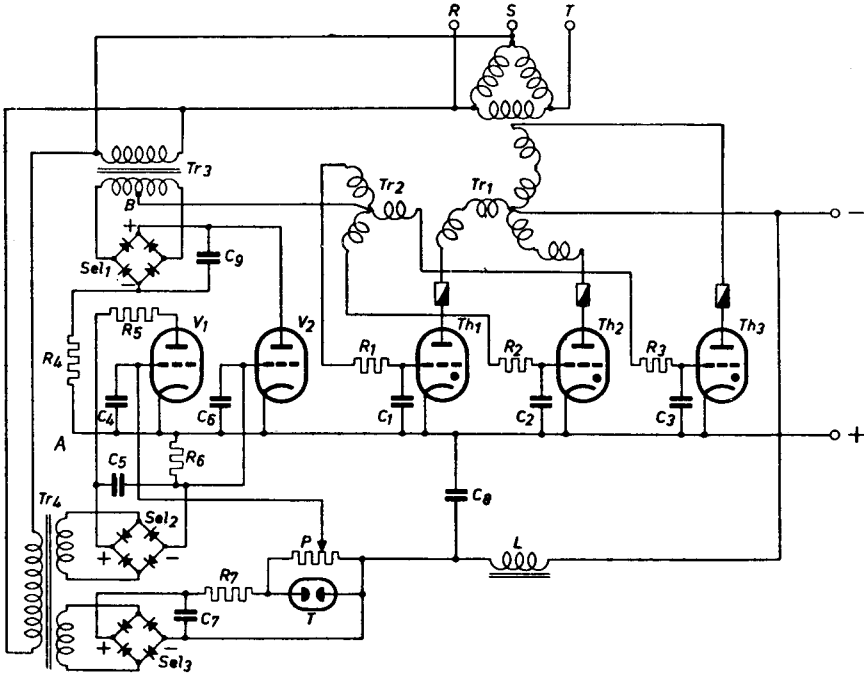


Fig. 13-13. Three-phase rectifying circuit the output voltage of which is held constant automatically (simplified).

Assume for the moment that tube V_2 receives a grid bias of such an amount that the d.c. resistance represented by the tube is equal to R_4 . The bridge will then be in balance and there will be no potential difference between points A and B . If, however, the grid voltage of V_2 is made more negative, the d.c. resistance of V_2 increases, and point B becomes positive with respect to point A , i.e. the firing angle of the thyristors becomes smaller. The grid bias of V_2 is represented by the voltage drop across resistor R_6 which carries the anode current of another high-vacuum tube, V_1 . This tube derives its anode voltage from selenium rectifier Sel_2 . Between control grid and cathode of V_1 a voltage is applied which is equal to the difference between the rectifier output voltage smoothed by a filter L, C_8 and a stabilised reference voltage which can be adjusted by means of potentiometer P .

If the rectifier output voltage tends to exceed the chosen value of the reference voltage, the control grid of V_1 becomes negative, thus reducing the anode current. The negative control grid voltage of V_2 consequently decreases, reducing the d.c. resistance of this tube. As a result, point B becomes negative with respect

to point A , and the firing angle of the thyratrons increases, thus reducing the output voltage of the rectifier.

Automatic Voltage Control with Saturable Core Reactor

A very common industrial requirement is a d.c. supply at about 110 volts which can deliver currents of some 75 to 150 amperes, the output voltage being adjustable over a limited range, either automatically or by hand.

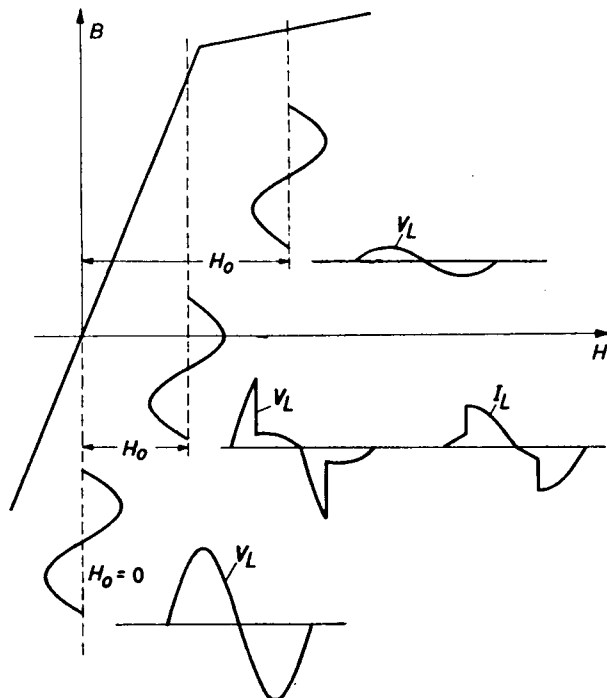


Fig. 13-14. Operation of a premagnetised reactor.

Thyratrons cannot normally be used in this case, since the anode current of this type of tube is generally limited to about 15 to 25 amperes. On the other hand, ignitron tubes would be too expensive because the ignition process is rather complicated and calls for equally complex auxiliary circuits. However, gas-filled high-current rectifying tubes with heated cathode can be used in connection with a premagnetised, i.e. d.c. polarised, choke inserted in the primary circuit, the premagnetisation being controlled electronically. Such a choke or saturable core reactor has an a.c. winding and a d.c. winding, the latter consisting of many turns of a fine wire. Its operation may be briefly explained by reference to fig. 13-14.

As long as the iron core is not premagnetised the working point lies on the steep part of the magnetising characteristic, an alternating current of given value causing a comparatively large fluctuation of the flux density B . The voltage drop appearing across the choke is then correspondingly high. If, however, a

current flows through the d.c. winding (this current may be only a few milliamperes because of the large number of turns) a polarising force H_0 is produced, shifting the working point into the saturation region of the iron core. Since the permeability μ is greatly reduced in this region, the inductance of the a.c. winding is lowered correspondingly.

In fig. 13-15 the voltage-current diagram of a saturable-core reactor is shown, and it can be seen that the a.c. resistance of such a device can be varied over a wide range by altering the premagnetising ampere-turns.

A schematic view of a saturable-core reactor is given in fig. 13-16. The a.c. winding is divided into two parts connected in series. The magnetic fields thus produced assist each other, but their influences are neutralised so far as the d.c. winding placed on the centre leg of the core is concerned. This avoids reaction of the a.c. winding upon the d.c. winding whereby high voltages might be induced because of the large number of turns in the d.c. coil. It is, of course, equally possible to connect the two halves of the a.c. coil in parallel. Several other constructions with different core shapes are also in use.

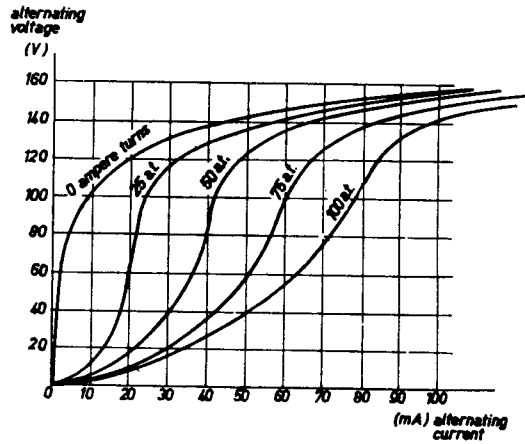


Fig. 13-15. Voltage-current diagram of a saturable core reactor.

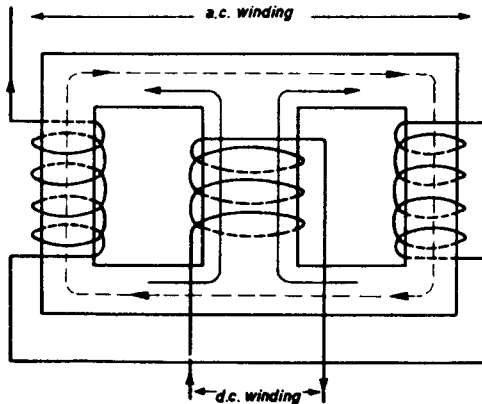


Fig. 13-16. Schematic view of a saturable-core reactor.

The use of a saturable-core reactor connected in series with the transformer primary has the advantage that the electronic control circuitry is independent of the number of secondary phases. Four- or six-phase rectifying circuits with

adjustable output voltage can therefore be easily designed. It is true that the range of adjustment of output voltage is not very large, if a variable reactor is used, but it is sufficient for many practical applications.

In fig. 13-17 the circuit of a six-phase rectifier is shown using three two-anode rectifying tubes 1849 (see fig. 2-17), the output voltage of which is held constant automatically. When switch *S* is closed, transformer *Tr*₂ is connected to the mains, and the heaters of all the tubes commence to warm up. After the pre-

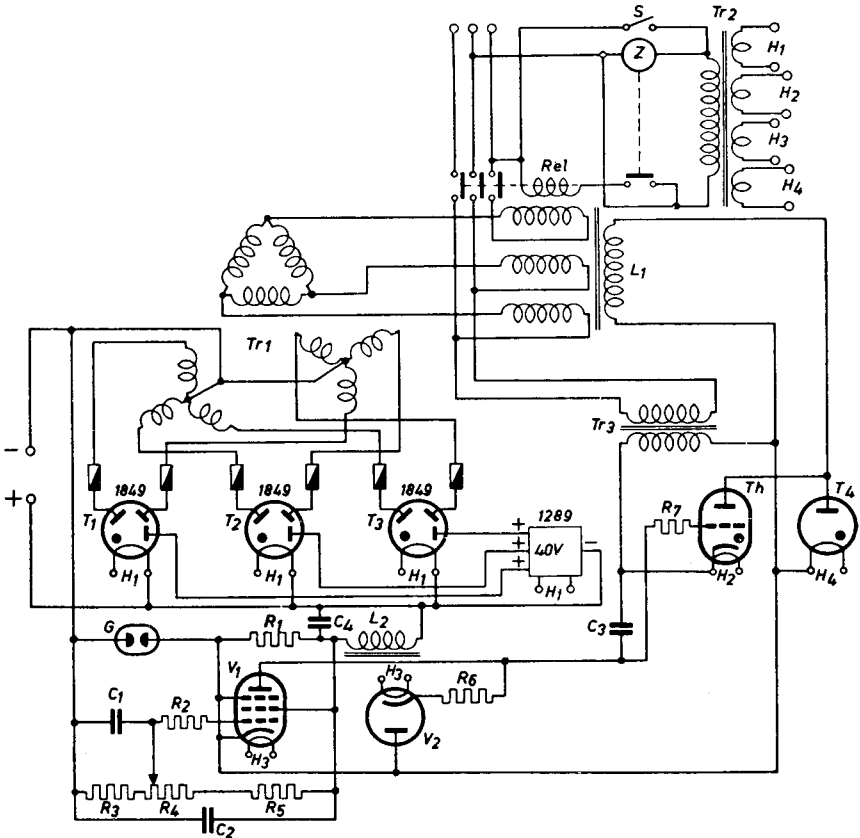


Fig. 13-17. Electronically controlled six-phase rectifying circuit.

heating time, the contact of the time delay relay *Z* is closed, energising relay *Rel* and thus connecting *Tr*₁ and *Tr*₂ to the mains. The three rectifying tubes *T*₁, *T*₂, *T*₃ (Type 1849) require for ignition a positive auxiliary voltage of about 40 volts to be applied to the ignition electrode. This voltage is taken from the auxiliary ignition unit, Type 1289, which is connected to the filament supply voltage of the rectifying tubes. The primary circuit of the power transformer *Tr*₁ contains the saturable-core reactor *L*₁. If the premagnetisation is increased by increasing the current in the d.c. winding, the a.c. resistance represented by *L*₁ is reduced, thus increasing the output voltage of the rectifier. This voltage is smoothed by a filter *L*₂, *C*₂ and applied via *R*₁ to the voltage reference

tube G which produces a very constant voltage of about 85 volts. It will be seen that this reference voltage appears between the negative terminal of the rectifier and the cathode of the high-vacuum tube V_1 . The control grid of this tube receives an adjustable voltage of approximately 84 volts via the voltage divider R_3, R_4, R_5 , so that the actual control grid bias is approximately -1 volt. Tube V_1 forms together with V_2, R_6 and C_3 a phase-shifting network which is separately

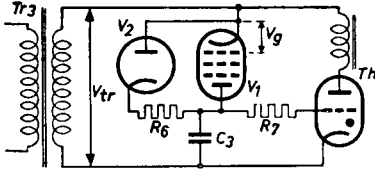


Fig. 13-18. Phase network included in the circuit of fig. 13-17.

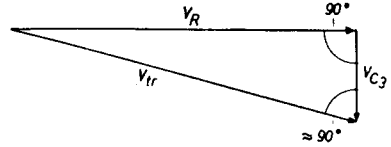


Fig. 13-19. Vector diagram of the phase network of fig. 13-18.

illustrated in fig. 13-18. The capacitor C_3 is charged alternately positive and negative by the transformer voltage V_{tr} through the inverse-parallel connected tubes V_1 and V_2 .

It may be assumed that tube V_1 when its control grid is biased to -1 volt, represents the same resistance as V_2 and R_6 connected in series. The positive and negative voltage half-cycles appearing across C_3 will then be of equal value and the combination of V_1, V_2 and R_6 can be symbolised by a resistor R with alternating current flowing through it. If R is made large compared with $1/\omega C_3$ the voltage vector diagram is as shown in fig. 13-19, i.e. the alternating voltage across C_3 is retarded in phase by nearly 90 degrees with respect to the transformer voltage, and the thyatron Th ignites during each positive half cycle of the transformer voltage with nearly 90 degrees delay.

If the negative control grid bias of V_1 is increased to -2 volts, the anode current of this tube is reduced, and C_3 receives an additional positive direct voltage which increases the superimposed a.c. voltage, thereby decreasing the firing angle of Th . On the other hand, if the control grid bias of V_1 becomes less negative, the anode current of this tube will increase, causing an additional negative charge of capacitor C_3 which in turn results in an increase of the firing angle of Th .

It will thus be seen that “vertical” control of thyatron Th is employed and is so sensitive that a voltage difference of approximately 1 volt at the control grid of V_1 may suffice to bring Th from passing full current to the non-conducting condition.

In the anode circuit of the thyatron the d.c. winding of the reactor L_1 is inserted which consists of many turns of fine wire and therefore represents a highly inductive load. A rectifying tube T_4 is connected in parallel, and this arrangement is often called a “zero-point anode”.

The operation of this tube may be explained by means of figs. 13-20 and 13-21. Fig. 13-20 shows a thyatron Th with an inductive load, fed by an alternating voltage source. Th being ignited (point A), the current increases only slowly, but flows during the whole interval X and becomes zero at point B , though the alternating voltage has changed its sign in the meanwhile. This is because the voltage appearing across the inductance L maintains the current also during

a part of the negative half cycle of the supply voltage, until the energy stored in the inductance has discharged.

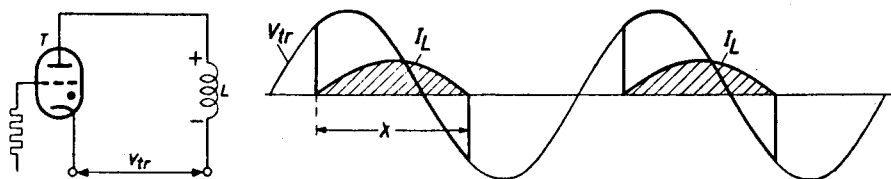
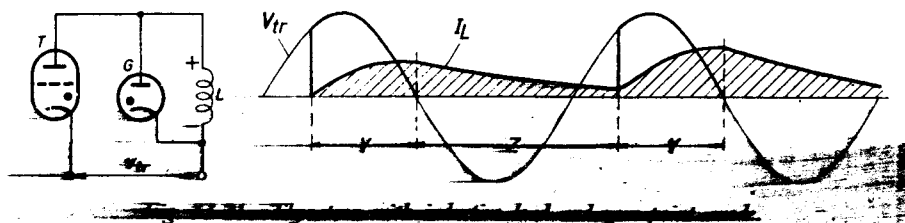


Fig. 13-20. Thyatron with inductive load.

If, however, according to fig. 13-21 a rectifying tube T is connected in parallel with L , the current flow is changed. Obviously Th and T cannot pass current at the same time. If Th is ignited, again a slowly increasing current flows through Th and L , but only during interval Y . The voltage appearing across L now produces a current flow through T and thus through L without being influenced by the negative half cycle of the supply voltage. This current continues to flow during the whole interval Z . Obviously the average current flowing through L can be influenced by the firing angle of Th . The earlier Th is ignited in each positive half cycle the greater will be the amount of energy stored in L and thus the current during interval Z .



Compared with the usual two-phase rectifying circuit, the circuit containing a zero-point anode has the advantage that only one controlled rectifying element is required, thus reducing the number of circuit components; moreover a two-phase power transformer with twice the secondary voltage is not required. The zero-point anode circuit is suitable for all cases in which the load is highly inductive, for instance, in rectifiers supplying current for the field windings in electronic motor control equipment.

A zero-point anode is often included also in two-phase rectifiers with highly inductive load. In the absence of such a tube an increased positive anode voltage appears across the tube which has not yet ignited in cases where the firing angle is large. This is illustrated in fig. 13-22. A firing angle of 90 degrees is assumed, and the voltage and current curves without (a) and with (b) zero-point anode are shown. In the first case the peak value of positive anode voltage is about $2\sqrt{2}$ times the r.m.s. transformer voltage. If the thyatron tubes employed cannot withstand this increased anode voltage a zero-point anode should be provided (fig. 13-22b). Moreover, this device reduces the reactive power requirement of the rectifier caused by the fact that the centre point of the current-time-integral per tube is retarded with respect to the peak value of the corresponding phase

voltage. As is seen, the inverse voltage of the zero-point anode does not exceed the peak value of the transformer voltage.

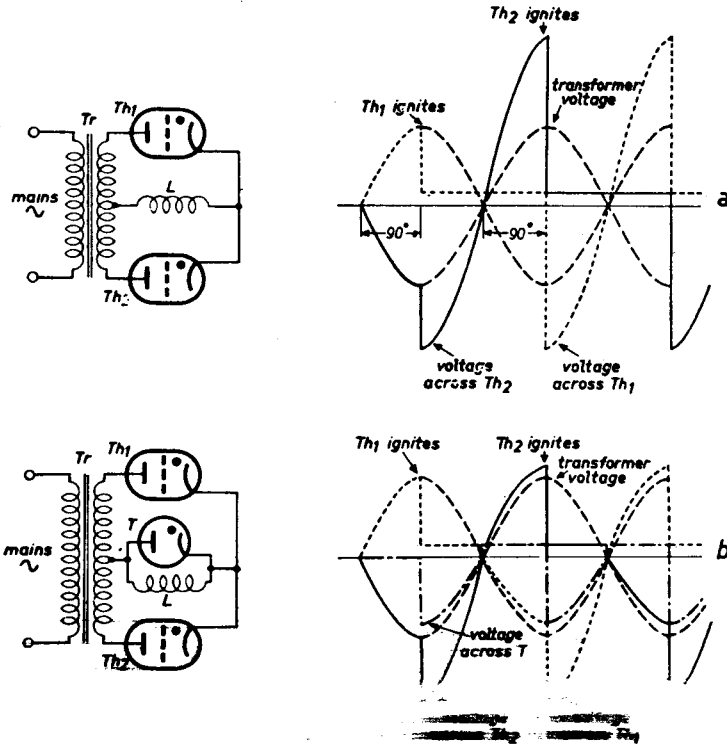


Fig. 13-22. Full-wave rectifying circuit without (a) and with (b) zero-point anode.

Returning to the circuit of fig. 13-17, it will be observed that a zero-point anode T_4 is used for the supply to the d.c. winding of the saturable core reactor L_1 . If the output voltage of the rectifier tends to exceed the value predetermined by the setting of R_4 the negative control grid bias of T_1 is reduced, thus increasing the firing angle of T_b . Consequently the premagnetising current of L_1 is reduced, the a.c. resistance of L increases and in turn causes a reduction of the rectifier output voltage.

Transformerless 3-phase Rectifier

A considerable proportion of the cost of rectifier equipment is that of the power transformer, so that substantial economy can be effected by omitting this component. The simplest form of transformerless three-phase rectifier is that in which each rectifying tube is connected between one phase (U , V or W) and the neutral wire (0) of the power line via the load, a suitable impedance Z_a being inserted in the anode lead of each tube.

This type of circuit, however, is inadmissible in many cases because the d.c. circuit is completed via the power line, so that the neutral wire carries the total direct current. This may conflict with the regulations issued by the electricity

supply authorities. This difficulty can be overcome by using a bridge arrangement in which the d.c. circuit is confined to the rectifier, and no direct current flows through the neutral wire (fig. 13-23).

Since in this case the filaments of the tubes are not all at the same potential, single anode tubes must be used. Six Philips Type 1177 tubes are used in the circuit illustrated, producing a total output current I_o of 75 amperes. A positive direct voltage of about 40 volts must be applied to the auxiliary electrodes of these tubes in order to facilitate ignition. For this reason additional windings are provided on the filament transformers Tr_1 to Tr_4 , producing voltages which are rectified by selenium rectifiers Sel_1 to Sel_4 .

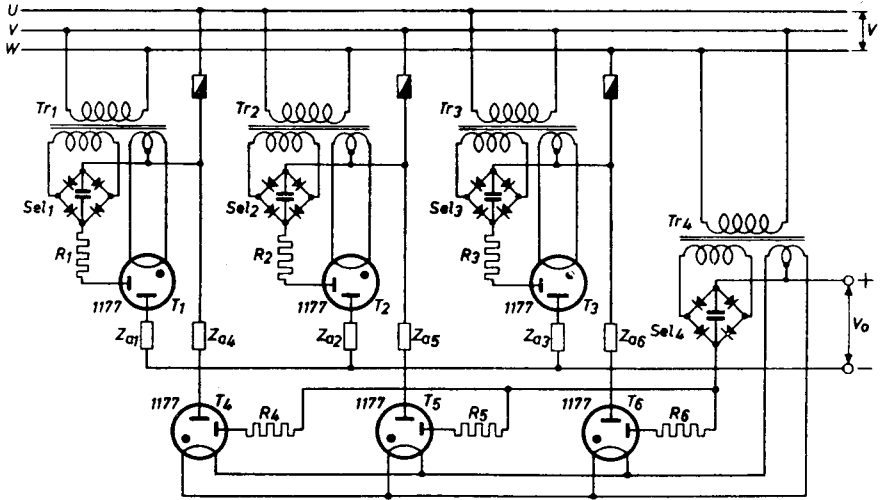


Fig. 13-23. Three-phase bridge circuit of a transformerless power rectifier.

The output voltage V_o of this bridge rectifying circuit is:

$$V_o = qV - 2(V_{arc} + V_z), \tag{13.104}$$

where q is a factor depending on the number of phases, V is the r.m.s. value of the mains voltage, V_{arc} is the arc voltage of the rectifying tubes, and V_z is the voltage drop across the impedance Z_a included in the anode circuit of each tube. In the case of a three-phase rectifier, $q = 2 \times 1.17/\sqrt{3} = 1.35$. In practice the voltage drop V_z should be approximately 3.5% of the output voltage V_o , which gives in the case considered:

$$1.07 V_o = 1.35 V - 2 V_{arc}, \tag{13.105}$$

or, V_{arc} being 12 volts for the 1177 rectifying tube,

$$V_o = 0.93 (1.35 V - 24) \text{ V}. \tag{13.106}$$

Thus, when the mains voltage between lines is 380 volts an output voltage of approximately 455 volts can be obtained.

An anode impedance Z_a must be included in each anode circuit to safeguard the tubes against possible overloading and to damp transient currents. This

impedance must perform the functions both of the inductance X_L provided by the power transformer in conventional circuits, and of the rated minimum anode resistance R_a (quoted in the technical data of the rectifying tube). The required inductance X_L is obtained by including a coil in the anode circuit, whilst if required an ohmic resistance is added to make up the prescribed value of R_a . The coil should preferably be air-spaced and must not contain a core, since this would be saturated by the d.c. component of the current flowing through the circuit.

Obviously Z_a is equal to the resistive and reactive components added in quadrature, i.e.

$$Z_a = \sqrt{R_a^2 + X_L^2}. \tag{13.107}$$

Considering $V_Z = 0.035 V_o$, and assuming that the anode current has approximately a square-wave form,

$$Z_a \approx \frac{0.035 V_o}{I_{a\text{ peak}}} \approx \frac{0.035 V_o}{I_o}. \tag{13.108}$$

Hence, Z_a can be calculated approximately, permitting the required value of X_L to be determined from equation (13.4).

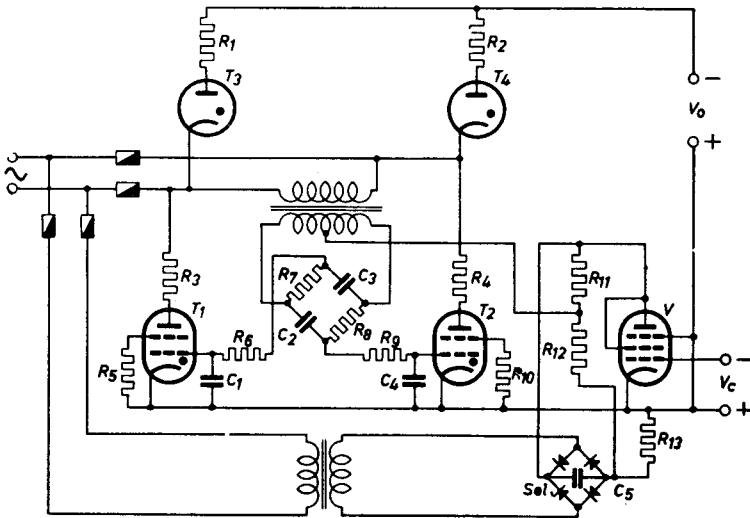


Fig. 13-24. Controlled two-phase bridge rectifying circuit without power transformer (simplified circuit).

Another bridge rectifying circuit is illustrated in fig. 13-24. Here, the output power is easily controlled by means of small alterations of a control voltage V_c . Since in a bridge rectifying circuit two rectifying elements connected in series are always conductive, it will be sufficient to control only one of them. Thus, only two grid-controlled tubes T_1 and T_2 are provided, the other two being normal gas-filled rectifying tubes T_3 and T_4 . "Vertical" control is applied to thyratrons T_1 and T_2 . Alternating grid voltages lagging by 90° and derived from

the phase network consisting of C_3 , R_7 and C_2 , R_8 are applied to these tubes. Superimposed on these voltages is a variable direct voltage supplied by a bridge circuit consisting of resistors R_{11} , R_{12} and R_{13} and the high-vacuum amplifier tube V . If the control voltage V_c applied to the first grid of V is made more negative, the internal resistance of V increases, making the control grids of T_1 , T_2 more positive. As a consequence, the output voltage V_o of the rectifier is increased.

The main disadvantage of transformerless rectifying circuits is that the nominal output voltage always bears a fixed relation to the supply voltage, which in turn is predetermined. Fortunately, however, in the most important case of a three-phase bridge rectifying circuit operated from the 380 volt power line, the direct output voltage is somewhat more than 440 volts, which is a quite convenient value for such purposes as driving d.c. shunt wound motors. For this reason this particular rectifying circuit is very often employed, especially for large outputs where the elimination of the transformer results in a substantial saving in cost. Moreover, the three-phase bridge-connected rectifying circuit produces a very small ripple voltage, of the same order as that in a six-phase rectifier.

Large Rectifiers for Electronic Motor Control Systems

The circuit of a large-power three-phase rectifier which is particularly suitable for use where a counter electromotive force is present in the load circuit, for

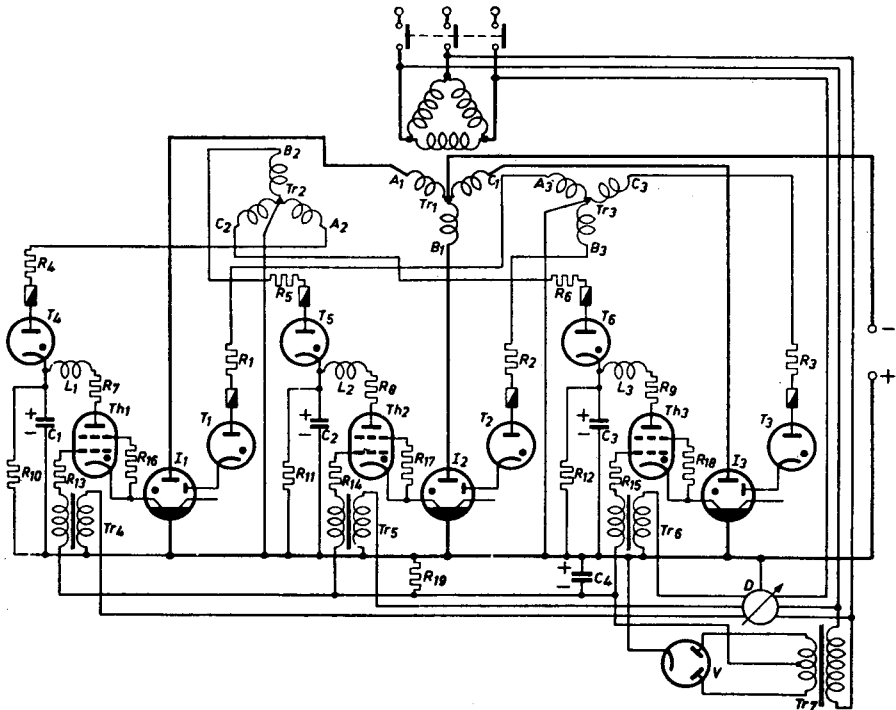


Fig. 13-25. Controlled three-phase rectifying circuit using ignitrons equipped with auxiliary anode.

instance in electronic motor control systems, is shown in fig. 13-25.

Special rectifier ignitrons I_1 , I_2 , I_3 are used which are equipped with a spare ignitor and an auxiliary anode (see fig. 5-3). The mercury pool cathode of this type of tube can operate at very large currents, and the steel jacketing renders ignitrons very suitable for heavy industrial service under unfavourable conditions. For the sake of simplicity all protection circuits and relays for the preheating time of the thyratrons etc., the cooling water flow control system and the filament transformers are omitted from fig. 13-25.

In addition to the power transformer Tr_1 , two smaller three-phase transformers Tr_2 and Tr_3 are provided, and supply the voltages for the auxiliary circuits. Tr_2 produces three voltages which are phase-shifted by about 180 degrees with respect to the power transformer voltages.

The ignition circuit of ignitron I_1 will first be explained. The ignition circuits of I_2 and I_3 are identical with that of I_1 . The voltage appearing at winding A_2 of transformer Tr_2 is rectified by tube T_4 so that capacitor C_1 is charged during the negative half cycles of the anode voltage to between 400 and 600 volts with the polarity indicated. This voltage is applied to the anode of thyatron Th_1 . When Th_1 is ignited a current surge flows via the ignitor of I_1 , causing this tube to ignite. The current flows for less than one millisecond, and the thyatron Th_1 then ceases to pass current, due to the inductor L_1 which tends to maintain the current, thus partially charging C_1 with inverse polarity. In the following negative half cycle of the anode voltage, however, capacitor C_1 is charged again by T_4 and is ready for ignition in the next positive half cycle.

To ensure that the thyatron ignites with the desired firing angle, a "horizontal" control circuit with phase-shifted impulses is provided from the secondary winding of the peaking transformer Tr_4 inserted in the control grid circuit of Th_1 .

The transformer primary is connected to a phase-shifting transformer *) D which produces three voltages which are phase-shifted by an adjustable angle with respect to the three-phase input voltage. Instead of the phase-shifting transformer, suitable RC -networks for the generation of phase-shifted voltages may be used. The control grid receives an additional negative bias of about 40 volts which is supplied by a small rectifier consisting of the transformer Tr_7 and the high-vacuum tube V . The rectifier unit Type 1289 (see fig. 3-13) is equally suitable for this purpose.

At the commencement of each positive half cycle of the anode voltage the auxiliary anode of I_1 also receives a positive voltage which is supplied by winding A_3 of transformer Tr_3 via tube T_1 . As soon as the ignition of I_1 is initiated by Th_1 a discharge from cathode to auxiliary anode occurs which is maintained until the end of the half cycle. Thus sufficient electrons and ions are present in the tube to permit a current of something less than one ampere to flow to

*) This is an electromechanical device similar to a three phase motor or generator, including three stator windings and three rotor windings connected to three collector rings on the shaft. The stator windings being connected to the three-phase line, a three phase output voltage will be produced in the rotor windings which is shifted in phase by a certain angle dependent on the position of the rotor shaft. The phase-shifting transformer thus acts like three RC phase networks in order to obtain three control voltages which are shifted in phase with respect to the line voltages by a variable angle.

the main anode as soon as the supply voltage is only a few volts higher than the arc voltage. This arrangement is necessary because in rectifier circuits, which use normally constructed ignitrons where the thyratrons are ignited by the anode voltage of the ignitrons, a minimum anode voltage of about 200 volts is required for the operation of the ignitors, and this voltage may not always be available in circuits containing a counter electromotive force. Moreover, ignitron tubes of normal construction as used in welding equipment require

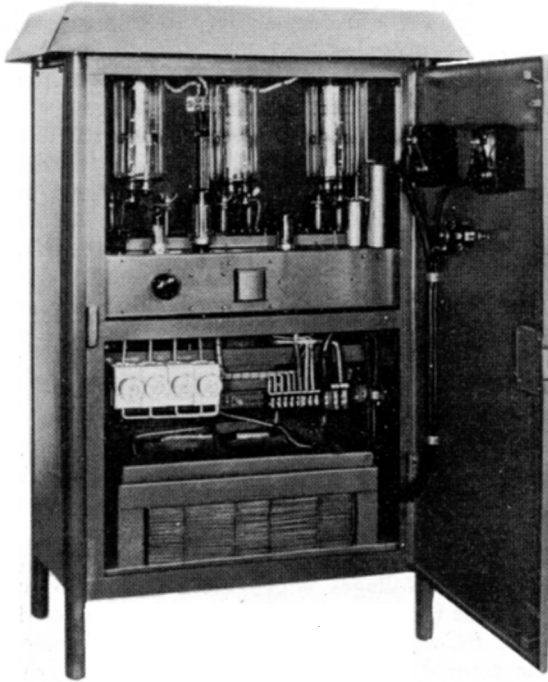


Fig. 13-26. Electronically controlled rectifier with constant output voltage for bookkeeping machines.

a certain minimum anode current (about 10 amperes) for maintaining the main discharge. Thus they tend to misfire if the rectifier works under approximately no-load conditions. For this reason also, a circuit such as that in fig. 13-25 using ignitrons with auxiliary anodes should be employed in applications where the load varies considerably, for instance in electronic motor control equipment. The rectifier tubes T_1 to T_6 pass only small average currents and may be therefore comparatively inexpensive.

If three PL 5555 rectifier ignitron tubes are used in the circuit of fig. 13-25, an average rectified output current of 600 amperes at a maximum output voltage of approximately 300 volts is delivered and can be regulated practically down to zero.

Fig. 13-26 shows an electronically controlled rectifier for bookkeeping machines the output voltage of which is held automatically constant within

the limits of $\pm 0,5\%$. The output power is 110 volts 40 amperes. Fig. 13-27 shows a three-phase rectifier of smaller output, the output voltage of which is similarly held constant by electronic control circuitry.

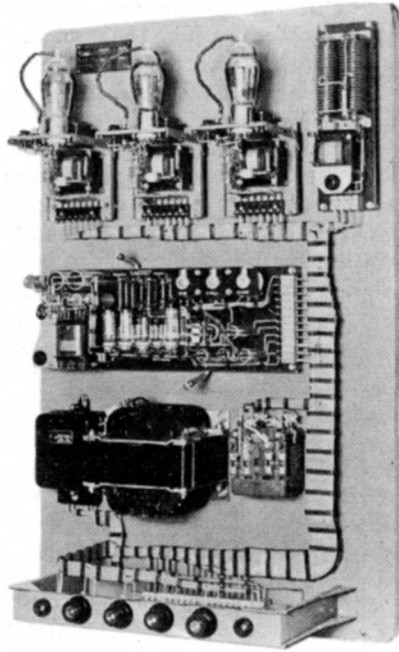


Fig. 13-27. Three-phase rectifier delivering constant output voltage by electronic control.

14. ELECTRONIC DIMMING OF LAMPS

Another example of the growing use of electronic tubes in a wide field of applications is the dimming of lamps, for example in stage lighting and the illumination of auditoriums in theatres, cinemas and the like. Hitherto, variable series resistors have usually been employed for dimming incandescent lamps, as shown in fig. 14-1. However, this method is uneconomic if heavy loads

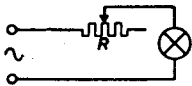


Fig. 14-1. Dimming an incandescent lamp by means of a variable resistor.

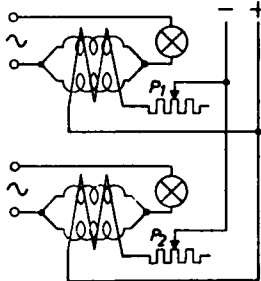


Fig. 14-2. Dimming by means of saturable core reactors.

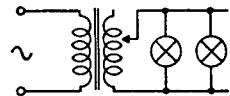


Fig. 14-3. Dimming by means of a variable voltage transformer.

have to be controlled, because of the losses in the resistor; saturable-core reactors are therefore often preferred (see fig. 14-2). By adjusting the pre-magnetisation the permeability and thus the a.c. resistance of the reactor may be varied. The range of adjustment is not very large, however, and furthermore a rather poor power factor has to be accepted. Another method of dimming is to use a variable voltage transformer (fig. 14-3) which will be preferred when large loads have to be controlled; but the controlling process then requires considerable mechanical equipment or a special servo motor.

For dimming incandescent lamps one of the methods described above may be used, but they are all quite unsuitable for controlling fluorescent lamps. Fluorescent lighting makes use of a gas discharge, the ignition voltage of which is comparatively high, so that dimming over a wide range by varying the supply voltage becomes difficult. However, a solution is found by using two inverse-parallel connected thyratrons, the ignition points of which are phase shifted.

A typical circuit is shown in fig. 14-4. In this arrangement an alternating current, the value of which can be varied by grid control, flows through the tubes and the load. The latter consists of the fluorescent lamp in series with a choke. The presence of the choke is essential for the correct operation of the circuit, since its inductance causes a certain delay of the lamp current, and thus of the voltage across the lamp, with respect to the mains voltage. Since the power factor of the choke is approximately 0.5, the voltage across the lamp is phase-delayed by

Fig. 14-4. Dimming a fluorescent lamp by means of two inverse-parallel connected thyratrons.

about 60 degrees with respect to the mains voltage.

As will be seen from fig. 14-5, in these circumstances the ignition voltage of the lamp is smaller than the instantaneous value of the mains voltage, so that ignition can take place without difficulty. When the thyratrons are ignited with a firing angle of 60 degrees, the full current will still flow through the lamp. When the firing angle is increased by means of grid control up to 135 degrees, the lamp current falls almost to zero (see fig. 14-5). Nevertheless, ignition of the lamp can still occur under this condition, since the instantaneous value of the mains supply voltage for $\varphi = 135^\circ$ is still about 220 volts if a mains voltage of 220 volt r.m.s. is assumed. Moreover, the voltage $V_{\max} \cdot \sin \varphi$ which, after ignition of a thyatron, is suddenly applied to the series connection of lamp and choke, excites an oscillating circuit consisting of the choke L and the capacitance C_1 appearing between the ignition electrodes and between the filament transformer windings connected thereto. Thus a momentary voltage surge appears across the lamp, and facilitates ignition.

Finally, the ignition voltage of the lamp can be reduced by applying a longi-

tudinal metallic strip, painted along the glass tube but not reaching to the ignition electrodes. A voltage difference occurring between these electrodes will now be concentrated at the small gaps between the ignition electrodes and the ends

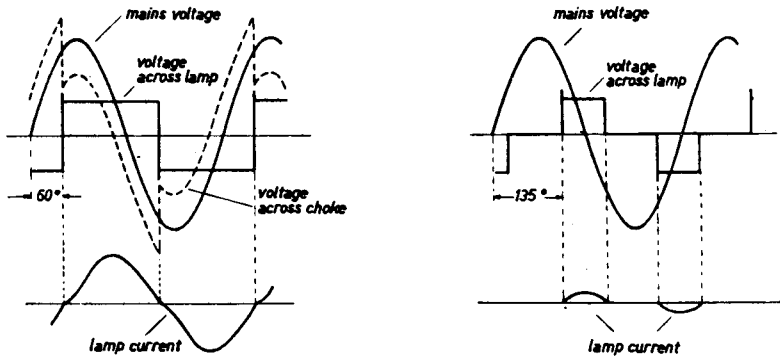


Fig. 14-5. Voltages and currents in the circuit of fig. 14-4 for different firing angles.

of the strip. The field strength near the electrodes is thereby increased, and the ignition voltage is consequently reduced.

Fig. 14-6 shows several fluorescent lamps, one of them being provided with the longitudinal strip. If the strip is earthed by a thin metal band some 3 mm wide placed round the glass tube at the non-earthed side of the lamp (see fig. 14-4), the ignition voltage now appears across the capacitance C_2 formed by the strip and the non-earthed electrode and the gap between them. By this method ignition of the lamp is facilitated so that the lamp will operate at currents down to about one milliamperes without flickering. Thus very gradual dimming is possible over the whole range down to the point at which the current may be interrupted without the spectators being aware.

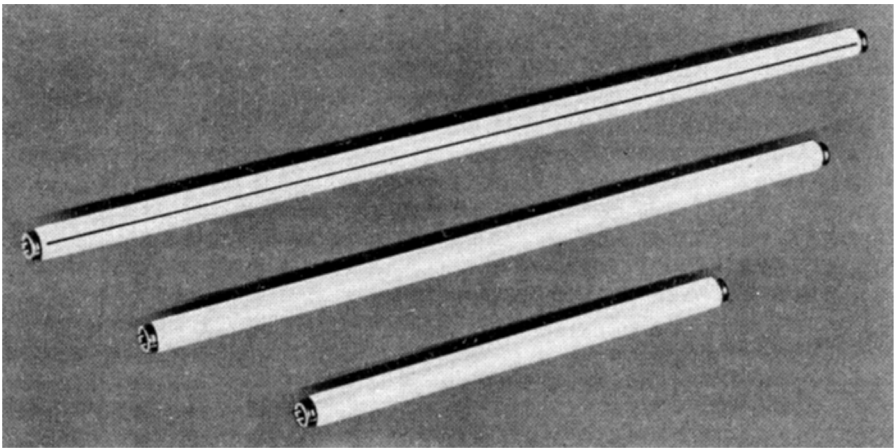


Fig. 14-6. Fluorescent lamps, one of them equipped with longitudinal strip.

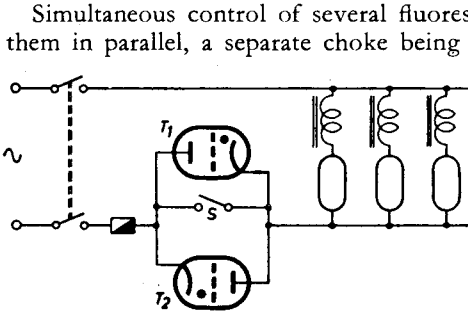


Fig. 14-7. Simultaneous dimming of several fluorescent lamps.

fig. 14-7. The thermal starter switch normally used is omitted, but the filaments of both electrodes of each lamp are fed from separate filament transformers in order to maintain the necessary temperature of the electrodes at reduced lamp current. Since the electrodes are self-heated by ionic bombardment at full current, it is recommended that the filament heating be reduced as the lamp current is increased.

It should be pointed out that a dimming device using two inverse-parallel connected thyristors is suitable not only for fluorescent lamps, but also for high-voltage neon illumination. In this case, however, the device must be connected in the primary circuit of the high-voltage transformer.

Electronic circuit for dimming fluorescent lamps

In fig. 14-8 is illustrated the complete circuit of an electronic dimming apparatus. The two thyristors obtain a negative control-grid bias from two

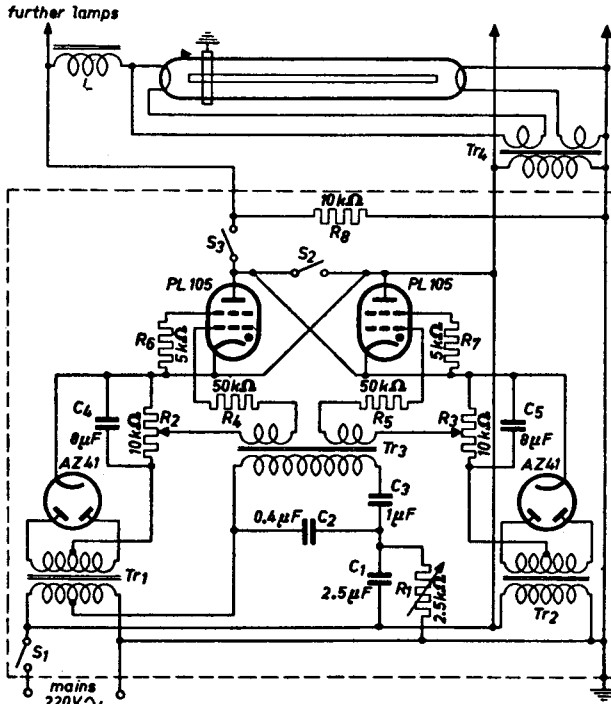


Fig. 14-8. Complete circuit of an electronic dimming device.

rectifiers, each containing an AZ 41 rectifying tube. Ignition is effected by phase-shifted impulses of about 80 volts, which are produced by a peaking transformer Type 84590. The primary of this transformer is connected to a phase-shifting network consisting of the primary winding of one mains transformer, the adjustable resistor R_1 , and the capacitors C_1 , C_2 and C_3 . The fluorescent lamps are connected as usual in series with a choke to the dimming device, but no power-factor correction capacitor is provided.

The electrodes of the lamp are heated by a filament transformer Tr_4 which has two separate secondary windings. The device is switched on by switch S_1 , switch S_2 being at first closed. After the thyratrons and lamps have been pre-

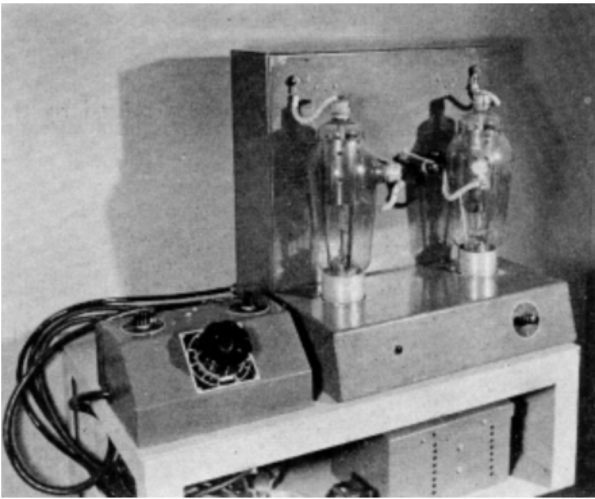


Fig. 14-9. Electronic dimming equipment for an output power of 3.2 kVA.

heated, switch S_3 is closed, causing the lamps to ignite. Switch S_2 may then be opened and dimming of the lamps can be controlled by means of the adjustable resistor R_1 . If full illumination is required it is recommended that S_2 be closed again in order to avoid passing the full lamp current through the thyratrons needlessly.

Electronic dimming devices are used in theatres, cinemas and concert halls; in lecture rooms and auditoriums where the lecture is illustrated by screen projection; in hospitals and sanatoria where X -ray negatives have to be examined; in photographic darkrooms and so forth. Examples of typical equipments are illustrated in fig. 14-9 and 14-10 which show apparatus for loads of 3.2 kVA and 6.4 kVA respectively.

The oscillograms reproduced in fig. 14-11 show the curves of lamp current and light intensity of four fluorescent lamps. The lamp current was adjusted to 1200, 600, 300, 150 and 75 milliamperes by an electronic dimming device of the type shown in fig. 14-8.

Not only fluorescent lamps but also incandescent lamps may be dimmed

by the method described above. Because, however, of the low cold resistance of incandescent lamps, due precautions with regard to current limitation must be taken.

The maximum number of lamps which may be connected to a dimming device of the type described is governed by the maximum average anode current rating I_a of the thyratrons employed. The r.m.s. value of the alternating current through a pair of inverse-parallel connected thyratrons is:

$$I = \frac{\pi}{\sqrt{2}} \cdot I_a, \quad (14.1)$$

and the apparent output power:

$$P = \frac{\pi I_a V}{\sqrt{2}}, \quad (14.2)$$

where V is the supply voltage. The power losses caused by the arc voltage drop of the tubes are neglected.

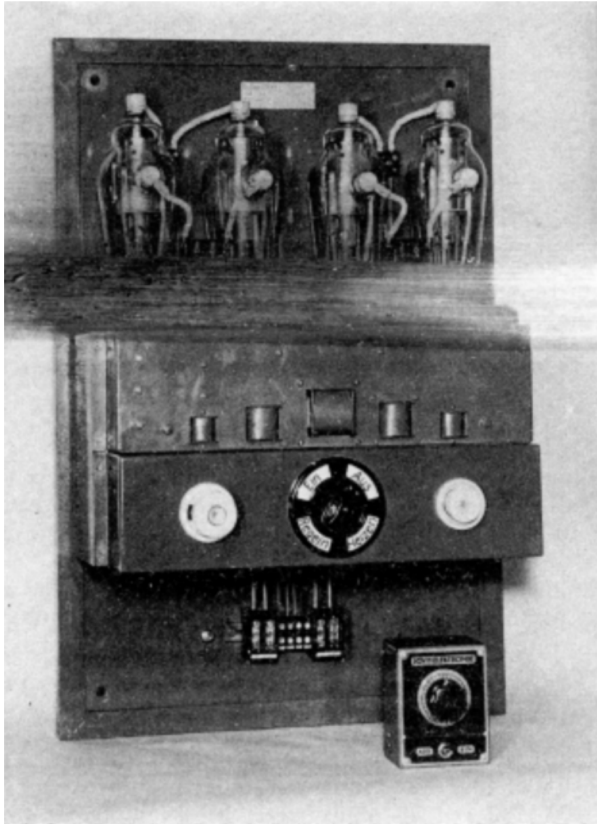


Fig. 14-10. Electronic dimming device for 6.4 kVA output power.

The useful output power is then:

$$P_o = \frac{\pi I_a V}{\sqrt{2}} \cdot \cos \varphi. \quad (14.3)$$

If the power consumption of one fluorescent lamp is p , and the power losses in the choke q , the number of lamps which may be connected is:

$$n = \frac{\pi I_a V}{(p + q) \sqrt{2}} \cdot \cos \varphi. \quad (14.4)$$

For example, if:

$$\begin{aligned} I_a &= 6.4 \text{ amperes,} \\ V &= 220 \text{ volts,} \\ \cos \varphi &= 0.5, \\ p &= 40 \text{ watts,} \\ q &= 9 \text{ watts,} \end{aligned}$$

then the number of lamps which may be connected is:

$$n = \frac{3.14 \times 6.4 \times 220 \times 0.5}{49 \times 1.41} = 32 \text{ lamps.}$$

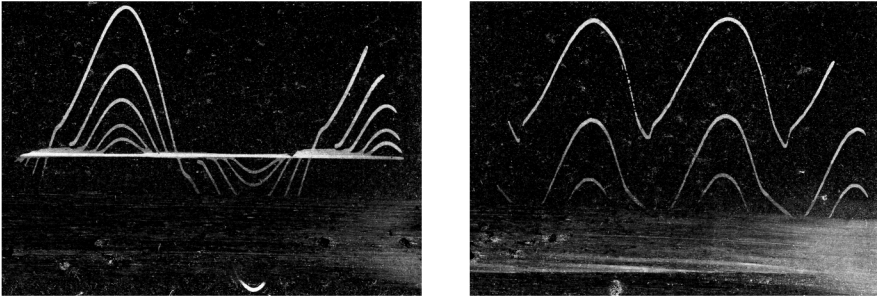


Fig. 14-11. Lamp current (a) and light intensity (b) plotted as functions of the firing angle, obtained by a dimming device of the type shown in fig. 14-8.

As might be expected in view of the greatly distorted wave-form of the current flowing through the lamps, there is some risk of interference with radio reception in the neighbourhood of the dimming apparatus. This interference can reach the receiver in two ways, namely:

- (1) via the mains,
- (2) by radiation from the wiring and lamps.

The first-mentioned is the more serious, especially in the case of d.c./a.c. receivers. However, this form of interference can be easily suppressed by means of the mains filter shown in fig. 14-12 which is suitable for dimming devices of the type illustrated in fig. 14-8. In addition to this filter, a capacitor should be shunted across the thyratrons. In general it will be advisable to mount the filter in the same enclosure as the dimming apparatus. The enclosure must be efficiently earthed by some other earth connection than that for the radio set.

Interference radiated from the wiring and the lamps must be suppressed

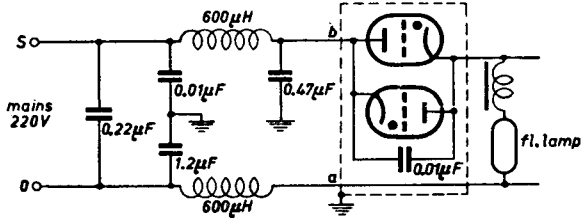


Fig. 14-12. Radio-interference suppression circuit.

by screening the whole of the wiring by running it in metal conduit. It is not advisable to install the radio set in the same room as the dimming apparatus. A screened aerial lead-in wire should be used in all cases.

Installation for Keeping the Lighting Constant

One of the essential advantages of electronics is the possibility of controlling an installation in such a way that hardly any control power is required. Advantage can be taken of this feature for controlling the illumination level, e.g. for keeping the illumination in workshops, laboratories, etc. constant at dusk.

Fig. 14-13 shows a circuit which is suited for this purpose. The illumination is checked by a photocell, which should be mounted at a suitable point. This photocell controls a d.c. amplifier equipped with a pentode E 80 F in grounded anode circuit, its load resistance R_2 being included in the cathode circuit. The

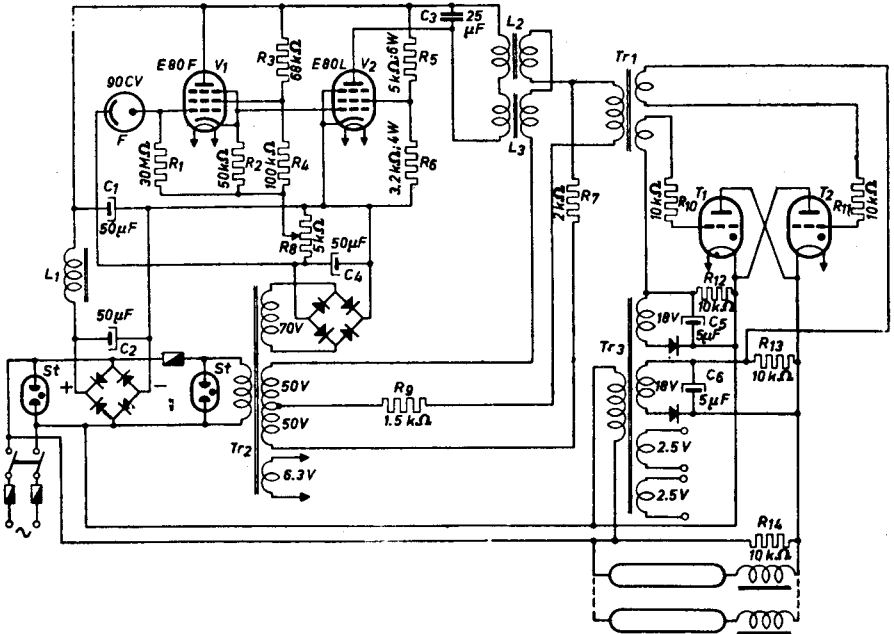


Fig. 14-13. Circuit of an installation for automatically controlling the illumination level of fluorescent lamps, depending on the daylight. T_1 and T_2 are two PL 5557 thyratrons connected in inverse-parallel.

voltage variations produced across this resistor are in phase with the variations of the input voltage; since, however, the two voltages have opposite effects on the grid circuit of tube V_1 , the negative feedback obtained is so heavy that the gain of this tube becomes smaller than unity. This is obviously a disadvantage, but the high input resistance of the circuit is in this way changed into a low input resistance, and the heavy negative feedback moreover ensures an extremely high stability.

The output voltage of tube V_1 is applied to the E 80 L output pentode V_2 . The d.c. windings of two saturable-core reactors L_2 and L_3 are included in the anode circuit of this tube, so that these reactors can be premagnetised. Their a.c. windings are connected in series in opposed directions

to prevent alternating voltages being induced in the d.c. circuit (cf. Chapter 13, p. 147). The saturable-core reactors form part of a phase-shifting network consisting of resistors and inductors. The peaking transformer Tr_1 is connected to the output of this network, so that, at the transformer secondary, voltage pulses are produced the phase of which can be varied with respect to that of the supply voltage by varying the premagnetising current and thereby the inductance

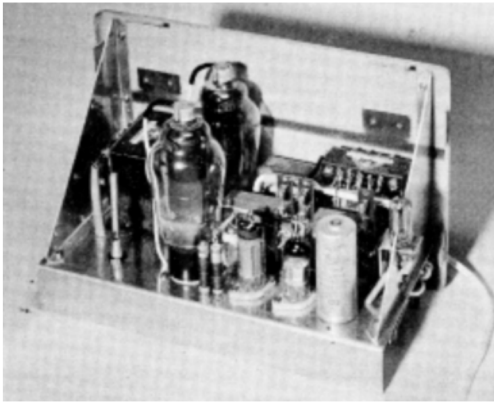


Fig. 14-15. Experimental apparatus according to the circuit shown in fig. 14-13.

results in an increase of the current flowing through the lamps according to the graph of fig. 14-14.

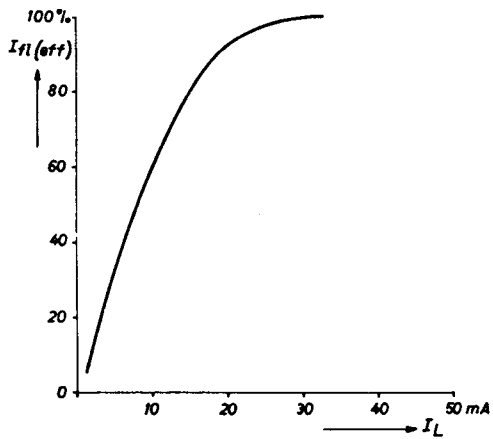


Fig. 14-14. Current I_{fl} flowing through the lamps as a function of the premagnetising current I_L .

of the reactors. In this way the current I_{fl} flowing through the fluorescent lamps is controlled according to the premagnetising current I_L of the reactors, as shown in the graph of fig. 14-14.

When the daylight decreases at dusk the current flowing through the photocell F also decreases, so that the grid of V_1 becomes less negative. The cathode current of this tube then increases, so that the grid of V_2 also becomes less negative, and its anode current — and hence also the premagnetising current through the reactors — increases. This

Fig. 14-15 depicts the chassis of an installation constructed on these lines. The saturable-core reactors L_2 and L_3 are provided with cores consisting of special transformer laminations containing 50% nickel. The d.c. winding consists of 5000 turns of copper wire with a diameter of 0.15 mm and the a.c. winding of 1300 turns of copper wire with a diameter of 0.12 mm. Transformer Tr_1 has a core of mu-metal; its primary consists of 900 turns of copper wire with a diameter of 0.1 mm and its secondary of 2×1500 turns of copper wire having the same diameter.

Electronic stage lighting circuit

For stage lighting in theatres a comparatively large number of different lighting circuits is necessary, which have to be controlled individually or in groups. The methods employed hitherto, namely variable resistors and auto-transformers, require a large and costly control board. Furthermore, if rheostats are used a large amount of electrical energy is wasted as heat. The advantages of electronic dimming methods are therefore of special interest in this class of installation.

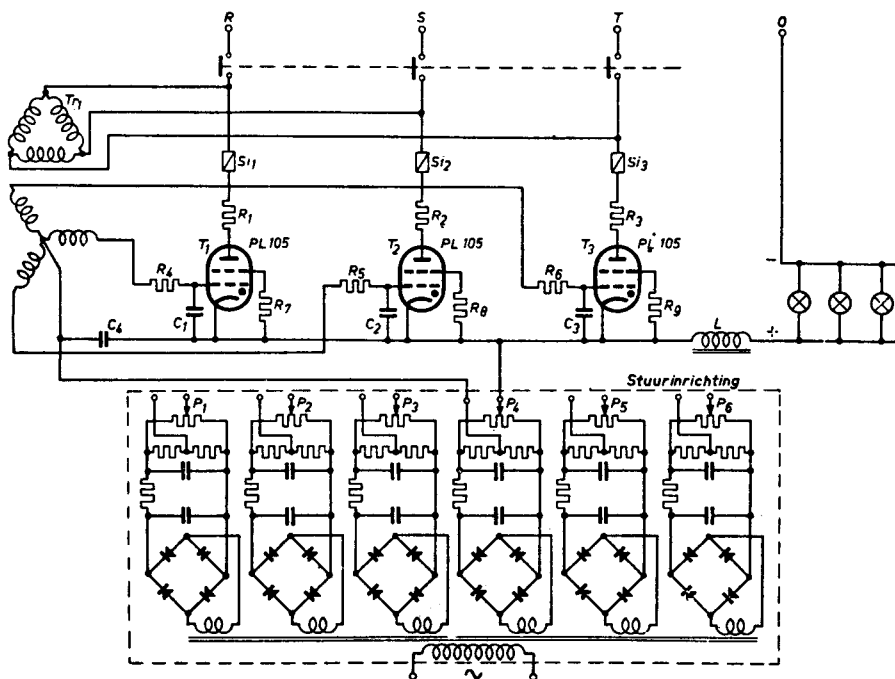


Fig. 14-16. Simplified three-phase rectifying circuit for dimming stage lighting.

Since the spot-lights and floods used for stage lighting are generally fed from a d.c. supply, a controlled two-phase or three-phase rectifier can be provided for each lighting circuit to convert the mains supply to a d.c. supply, the voltage of which is variable.

Fig. 14-16 shows the basic circuit of a suitable three-phase rectifier with

control panel. Rectification is performed by three thyratrons T_1 , T_2 and T_3 which are fed directly from the mains, a power transformer being omitted in order to reduce cost. It is true that a d.c. component then flows in the neutral conductor, but as the larger theatres often possess their own power station or mains power transformer, this may not be objectionable. If, however, this arrangement is considered undesirable, a power transformer must be provided unless a three-phase full-wave rectifying circuit is used which avoids imposing a d.c. load on the mains. Furthermore, as mentioned in an earlier chapter, the latter circuit has the advantage of a lower ripple voltage, which is of the same order as that in a six-phase rectifier.

The tubes T_1 , T_2 and T_3 are grid-controlled by voltages consisting of a variable positive or negative direct voltage and alternating voltages which are phase-

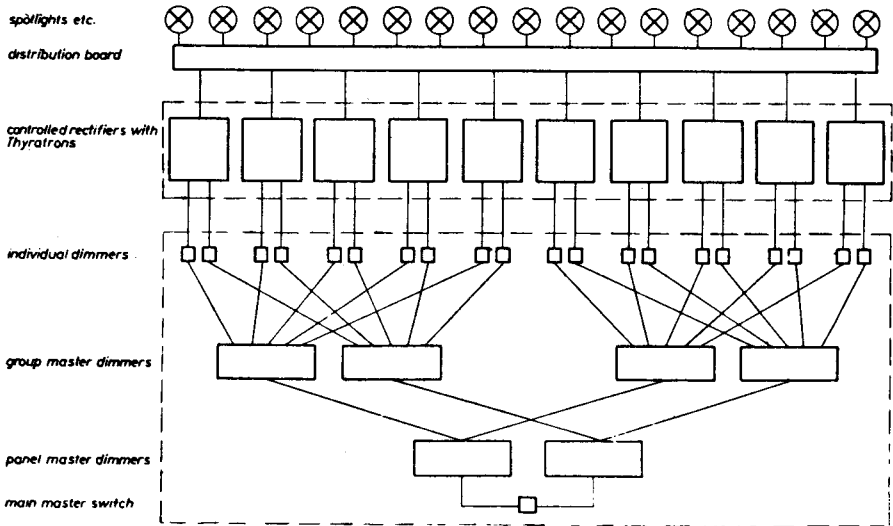


Fig. 14-17. Block diagram of a complete electronic stage lighting control.

agged by 90° behind the corresponding anode voltages. These grid voltages are produced by transformer Tr_1 . The variable direct voltage is supplied by a rectifier included in the control panel. Several potentiometers, P_1 to P_6 , are provided and serve for controlling the rectifier circuit shown, and also further similar lighting circuits.

A block diagram of a complete system of electronic control for stage lighting is shown in fig. 14-17. The flood- and spot-lights are connected to a distribution board at which connections to the individual rectifier circuits can be made. Each rectifier circuit can be controlled by its individual dimmer. A number of circuits form a group, controlled by a "group master-dimmer", and all groups can be dimmed simultaneously by a "panel master-dimmer". All these controls are duplicated on a second panel so that while one panel is in use the switches and dimmers of the second panel may be pre-set for the next scene. For the next scene the lighting is transferred to the second panel by the main master switch, which changes over all or some of the circuits to the new settings.

Since all the control switches and dimmers carry only small currents, they can be of miniature type so that, for instance, two complete control panels for a 48-way installation can easily be mounted on a control desk measuring only three feet square. This permits control from a remote point, for instance from a position in front of the stage, where an unobstructed view of the scene can be obtained.

15. SPEED AND TEMPERATURE CONTROL

Electronic control of operating speeds, process temperatures and similar factors contribute substantially to uniformity and improved quality of the product. In many of these devices photocells play an important part as, for instance, in automatic register control devices. A typical example of register control is paper positioning in wrapping machines and other paper-processing equipment. The paper for the final product, such as paper bags, boxes etc. is usually supplied in reels, printed with a trade name or other matter at regular intervals by some cheap rotary printing process. This continuous paper must be cut into suitable lengths and it is obvious that the printed matter should always occupy the same position on the cut pieces.

It is therefore necessary that the cutting frequency be accurately related to the speed at which the paper passes through the machine. It is true that a slight error of, say, $\frac{1}{16}$ " may be of little consequence for a single piece, but since the errors are cumulative this would result in a total error of as much as 6" after 100 cuts.

Mere synchronisation of the cutting frequency and the speed of travel does not, however, ensure satisfactory operation because it does not take into account the uncontrollable expansion or contraction of the paper due to temperature and humidity variations. A regulating device is therefore required which detects and subsequently corrects the positive or negative errors of every cut. The obvious solution is to use for this purpose a photocell-controlled device which responds to register marks printed on the margin of the paper together with the trade name.

Paper Register Control

Fig. 15-1 shows the block diagram of such a device. The travelling paper is provided with register marks M at regular intervals, indicating the position at which the paper should be cut. The width of the marks should be roughly

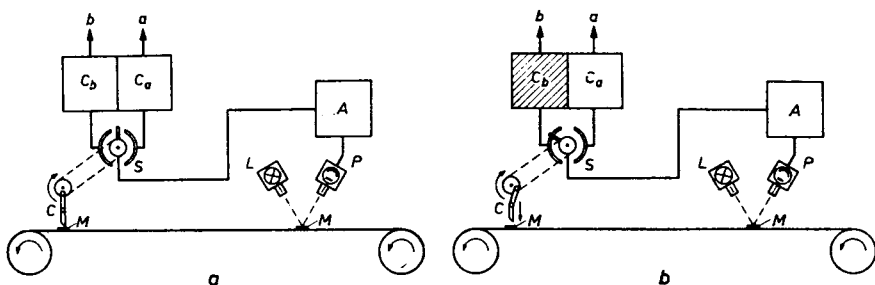


Fig. 15-1. Block diagram of a regulating device for cutting continuous paper.

mission gear. When, on the other hand, the cutting frequency is too high, the signal is fed to the controlling unit C_a which temporarily accelerates the paper speed by means of a relay and a servo-motor connected to a . Errors are thus corrected before the cumulative error has reached an inadmissibly high value.

Fig. 15-2 shows the circuit of the electronic part of the regulating device. The direct current required is supplied by a Type 1701 gas-filled rectifying tube T_1 . The gas-filled photocell F , Type 3546 is d.c. fed, so as to respond to variations in light intensity at every instant. (If the photocell were a.c. fed no signal would be produced by the register marks during the half cycles when the photocell anode is negative.)

When the photocell is excited by the reflection of the beam of light from the bright paper surface a photo-electric current will flow which gives rise to a voltage drop across the resistor R_6 , so that the capacitor C_4 is charged with the polarity indicated in fig. 15-2. During the short interruptions of the beam of light by the register marks this capacitor will produce a negative voltage at the control grid of the EF 40 pentode V . This temporarily reduces the anode current of the EF 40 as a result of which a positive voltage pulse occurs at its anode and is transmitted via the capacitor C_2 to the control grid of one or the other of the Type PL 5557 thyratrons, Th_1 or Th_2 , depending on the position of the rotary switch S . If the paper speed is too high, thyatron Th_2 will ignite; and if too low, thyatron Th_1 will ignite. At the correct speed the switch S is in its neutral position during the passage of the register mark so that neither of the thyratrons will ignite.

Once a thyatron has ignited, it will remain conductive because it is d.c. fed. The relay Rel_2 (or Rel_1) will then be energised, so that the braking (or accelerating) circuit connected to b (or a) is closed. In the meantime, however, a second contact of this relay is opened, to disconnect the capacitor C_3 from the positive supply voltage. This capacitor had previously been charged as indicated in fig. 15-2, so that the PL2D21 thyatron Th_3 was ignited as a result of the positive voltage applied to its first grid. As soon as C_3 is disconnected from the supply voltage it discharges via the resistors R_{14} and R_{15} and the cathode-to-grid path of the thyatron Th_3 , and since a fixed negative grid bias of about -2 V derived from R_4 is applied via the resistor R_{12} to the second grid of this tube, it will extinguish after a short interval. As a result Rel_3 will no longer be energised, so that it will drop out and the thyatron PL 5557 also extinguishes. The relay Rel_2 (or Rel_1) then drops out and the braking (or accelerating) circuit is interrupted. The capacitor C_3 is then again charged via R_{13} , so that the thyatron Th_3 ignites again. In this way the duration of the braking or accelerating action is determined and can be adjusted by means of the variable resistor R_{15} .

An Alternative Register Control

The device as described, using mechanical relays, does not, however, work satisfactorily for very high cutting speeds. It is possible, of course, to increase the maximum speed of this register control device by printing the register mark opposite only every second, third or fourth etc. trade mark, but finally a limit will be reached. In this case a device which does not employ mechanical relays must be used.

A suitable circuit is given in fig. 15-3. Instead of a rotating switch, two photo-

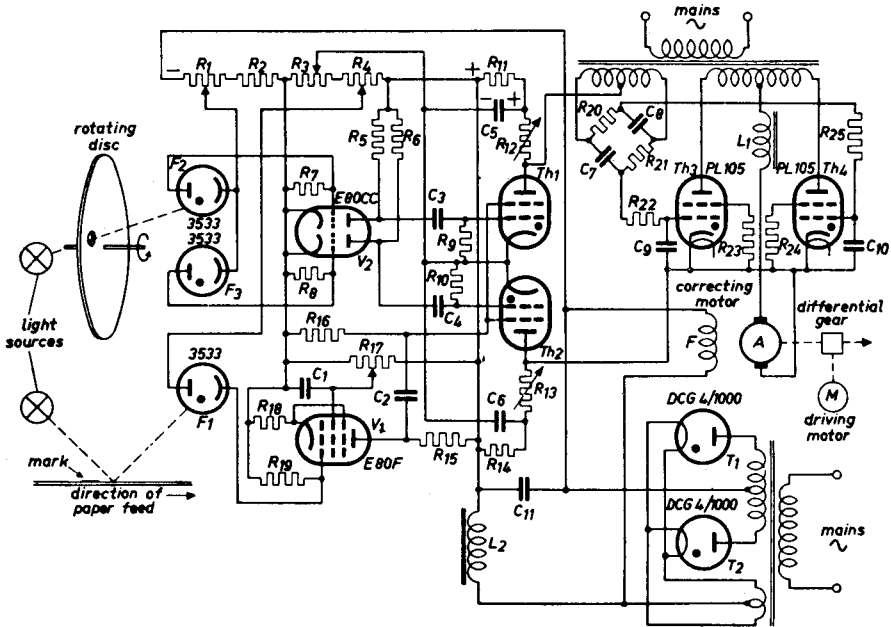


Fig. 15-3. Electronic regulating device using no mechanical relays.

cells F_2 and F_3 are provided which are illuminated by a light source if a hole in a disc rotating synchronously with the cutting frequency lets the beam pass through. The photocells are mounted so that the one (F_2) is illuminated just before a cut is made, and the second (F_3) just after the cut. At the instant when the cut is being made neither cell is illuminated. At that instant, however, the light falling on photo-cell F_1 is reduced by the dark register mark, provided the travelling paper is at the correct position for cutting. If this is not the case, the reduction of the light falling on F_1 will coincide with the illumination of F_2 or F_3 according to whether the paper speed is too high or too low.

This is illustrated in fig. 15-4. The paper is driven by a motor working at constant speed. An auxiliary "correcting" motor, which is coupled to the main motor by a differential gear, runs at a normal speed if the speed of the paper is correct. It is a d.c. shunt wound motor whose field is fed by a rectifier equipped with tubes T_1 , T_2 . The armature is supplied

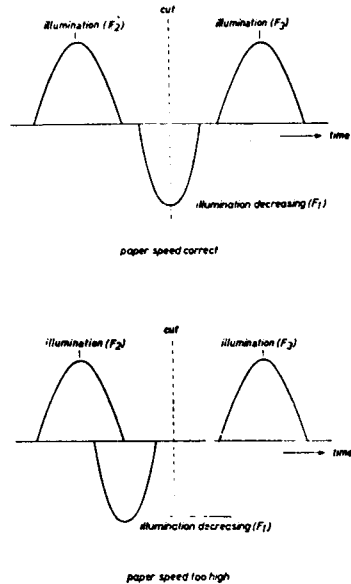


Fig. 15-4. Position of voltage impulses for correct paper speed, and for the case when the paper speed is too high.

at normal voltage by a second rectifier equipped with thyratrons Th_3 , Th_4 . To these tubes are applied alternating control grid voltages which lag by 90 degrees behind the anode voltages; they are produced by the phase-shifting network consisting of C_8 , R_{20} , and C_7 , R_{21} . As long as the controlling thyratrons Th_1 and Th_2 (Type PL2D21) are non-conducting, no voltage difference exists between their anodes and consequently no further voltage appears in the control grid circuits of thyratrons Th_3 and Th_4 (Type PL 105). These tubes will obviously then ignite with approximately 90 degrees phase delay (see Chapter 3 "Vertical Control") so that the output voltage of the rectifier is reduced, causing the motor to run at normal speed.

Each time the photocell F_2 or F_3 is illuminated a negative voltage appears at the corresponding control grid of the double triode V_2 , causing a decrease of the anode current of that particular section. Since this is only a momentary decrease, a positive voltage impulse arises at the anode which is applied via capacitor C_3 or C_4 to the first grid of thyatron Th_1 or Th_2 . However, this does not result in ignition of either thyatron, since their second grids receive also a negative bias which can be preadjusted by means of potentiometer R_2 .

Photocell F_1 is normally illuminated so that a positive voltage drop appears across resistor R_{19} and tube V_1 passes current. If, however, the light falling on F_1 is reduced for a moment by the dark register mark, the voltage drop across R_{19} is correspondingly reduced, and the sudden decrease of the anode current flowing through V_1 produces a positive voltage impulse at the anode which is applied via capacitor C_2 to the second grids of thyratrons Th_1 and Th_2 . But this impulse alone is also not sufficient to make either of these tubes ignite.

If now the paper speed is correct, positive impulses are applied *successively* to the grids of thyratrons Th_1 and Th_2 when the register mark passes by, so none of these tubes ignites. Matters are different, however, if the paper speed is, for instance, slightly too high. In this case the impulses caused by F_2 and F_1 will partially or completely coincide, as is shown in the lower part of fig. 15-4. As a result the thyatron Th_1 will ignite, and capacitor C_5 , which had been charged with the polarity indicated, can discharge through Th_1 and R_{12} .

Since resistor R_{11} is comparatively large, recharging of C_5 takes a greater time than discharging, so the thyatron will finally extinguish again when the voltage across C_5 has become smaller than the arc voltage. As long as Th_1 is ignited its anode potential corresponds to the arc voltage, and between the anodes of both thyratrons Th_1 and Th_2 appears a voltage difference which is applied as a negative potential to the control grids of thyratrons Th_3 and Th_4 causing the output voltage of the rectifier to drop almost to zero for a short time. The speed of the correcting motor therefore decreases, thus reducing the paper speed and re-establishing its synchronism with the cutting frequency.

The duration of the speed decrease and thus the amount of correction can be varied by means of R_{12} over a certain range. If the paper speed is too low, the impulse caused by F_1 will coincide with that transmitted by F_3 , causing thyatron Th_2 to fire. Now the voltage difference appearing between the anodes of Th_1 and Th_2 is applied to the control grids of thyratrons Th_3 , Th_4 as a positive potential, thus causing the rectifier output voltage to rise to its full value. The speed of the

correcting motor is correspondingly increased, again restoring the synchronism with the cutting frequency.

Lateral Register Control

Another type of register control frequently used is for checking the position of a strip of paper or fabric fed through a processing machine. Lateral displacements of the strip passing through the machine are counteracted by a servo-motor which rotates, in a clockwise or anticlockwise direction, driving gearing which leads the paper back to the correct position.

It is possible to use as detecting elements miniature switches operated by the travelling material, and switching the servo-motor on or off by means of contactors. This method has, however, serious drawbacks. First, to avoid continuous operation of the contactors, which would soon lead to a breakdown, it would be necessary to introduce a fairly large "dead zone" which obviously reduces the accuracy of the regulating system. Second, the material is often so delicate (thin tissue paper, for example) that it is not mechanically strong enough to operate switches. Finally — and this will usually be the decisive factor — contactors can only switch the motor on or off so that it either operates at full speed or stops, the speed thus being independent of the magnitude of the error to be counteracted. This makes a further increase of the dead zone necessary to avoid the risk of hunting.

In the device described below the detecting element again consists of a photocell, and the quantity of light falling on its cathode determines both the *direction* and the *speed* of rotation of the motor for correcting lateral displacement of the material.

When the normal quantity of light falls on the photocell the motor remains at rest. As soon as the quantity of light increases the motor starts to run in a clockwise direction, its speed increasing approximately linearly with the intensity of light. Similarly, the speed of the motor when running in an anti-clockwise direction increases with the reduction of the quantity of light. When the photocell is receiving the maximum quantity of light or is in complete darkness the motor thus runs at full speed in a clockwise or anti-clockwise direction, whilst its speed decreases as the average intensity of light is approached.

Fig. 15-5 shows the circuit of this device. The field coil FW of the d.c. shunt-wound motor is fed from the rectifier formed by the transformer Tr_3 and the gas-filled rectifying tube T_1 (1701). The direct voltage supplied by this rectifier is also used, after having been smoothed by the filter L_2C_3 , for feeding the gas-filled photocell Type 3554 (or 3533) and the EF 40 pentode V_2 which serves as a pre-amplifier.

The armature A of the motor is fed from the a.c. mains via the choke L_1 and the two PL 5559 thyratrons Th_1 and Th_2 connected in anti-parallel. When both thyratrons are blocked or only pass small, equal currents, the armature remains at rest, but as soon as the phase of the control voltage applied to one of these thyratrons is shifted, so that its current increases, the motor will rotate at the corresponding speed in one direction or the other. Since as a rule the current drain of servo-motors is fairly small, two PL 5559 thyratrons, which will pass a mean anode current of 2.5 amperes, suffice.

Special attention should be paid to the control circuit of these tubes. The cathodes are at different potentials so that the tubes must be controlled by shifting

the phase of a voltage peak applied to the control grid with respect to that of the alternating anode voltage. The transformer Tr_4 supplies two alternating voltages in anti-phase, thus providing the required negative grid bias during the half cycles when the anodes of the corresponding tubes are positive. The

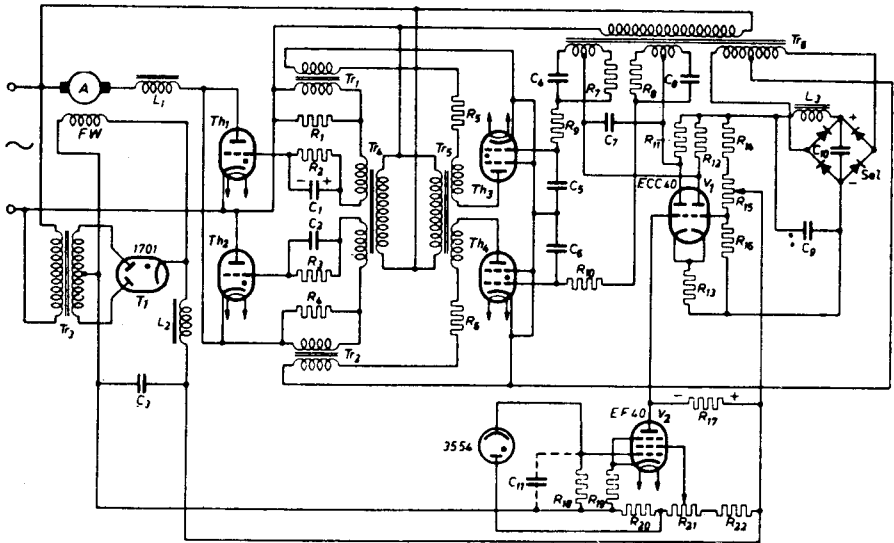


Fig. 15-5. Side register control circuit.

- $R_1 = 100 \text{ k}\Omega$
- $R_2 = 50 \text{ k}\Omega$
- $R_3 = 50 \text{ k}\Omega$
- $R_4 = 100 \text{ k}\Omega$
- $R_5 = 40 \text{ k}\Omega$
- $R_6 = 40 \text{ k}\Omega$
- $R_7 = 15 \text{ k}\Omega$
- $R_8 = 15 \text{ k}\Omega$
- $R_9 = 50 \text{ k}\Omega$
- $R_{10} = 50 \text{ k}\Omega$
- $R_{11} = 25 \text{ k}\Omega$

- $R_{12} = 25 \text{ k}\Omega$
- $R_{13} = 4.4 \text{ k}\Omega$
- $R_{14} = 70 \text{ k}\Omega$
- $R_{15} = 50 \text{ k}\Omega$
- $R_{16} = 40 \text{ k}\Omega$
- $R_{17} = 10 \text{ k}\Omega$
- $R_{18} = 2 \text{ M}\Omega$
- $R_{19} = 500 \Omega$
- $R_{20} = 8 \text{ k}\Omega$
- $R_{21} = 10 \text{ k}\Omega$
- $R_{22} = 4 \text{ k}\Omega$

- $C_1 = 1 \mu\text{F}$
- $C_2 = 1 \mu\text{F}$
- $C_3 = 16 \mu\text{F}$
- $C_4 = 0.2 \mu\text{F}$
- $C_5 = 0.05 \mu\text{F}$
- $C_6 = 0.05 \mu\text{F}$
- $C_7 = 0.01 \mu\text{F}$
- $C_8 = 0.2 \mu\text{F}$
- $C_9 = 8 \mu\text{F}$
- $C_{10} = 8 \mu\text{F}$
- $C_{11} = 1000 \text{ pF}$

- $L_1 = 5 \text{ H}, 2 \text{ A}$
- $L_2 = 10 \text{ H}, 60 \text{ mA}$
- $L_3 = 10 \text{ H}, 60 \text{ mA}$

Sel = selenium rectifier for 220 V, 20 A.

$Tr_1 = Tr_2 =$ primary: 3500 turns 0,12 mm \varnothing ;
secondary: 3500 turns 0,12 mm \varnothing .

$Tr_3 =$ primary: mains voltage;
secondary: $2 \times 285 \text{ V}, 0,3 \text{ A},$
 $6,3 \text{ V}, 0,2 \text{ A},$
 $1,8 \text{ V}, 1,8 \text{ A},$
 $5 \text{ V}, 4,5 \text{ A},$
 $5 \text{ V}, 4,5 \text{ A}.$

$Tr_4 =$ primary: mains voltage;
secondary: 12 V, 0,01 A,
12 V, 0,01 A.

$Tr_5 =$ primary: mains voltage;
secondary: 110 V; 0,1 A,
110 V, 0,1 A,
6,3 V, 1,2 A.

$Tr_6 =$ primary: mains voltage;
secondary: $2 \times 15, 0,02 \text{ A},$
 $2 \times 15 \text{ V}, 0,02 \text{ A},$
 $2 \times 90 \text{ V}, 0,03 \text{ A},$
 $6,3 \text{ V}, 0,6 \text{ A}.$

capacitors C_1 and C_2 , shunted across the grid-protecting resistors R_2 and R_3 , are charged as indicated in fig. 15-5 during the positive half cycles of the grid bias and this negative voltage is added to the negative alternating grid voltage during the following half cycle, which safeguards the thyratrons from random ignition. The igniting peaks superimposed on this grid bias are supplied by the transformers Tr_1 and Tr_2 , the primaries of which are fed by the two thyratrons Th_3 and Th_4 (PL2D21).

The cathodes of the latter tubes are interconnected, so that they can be controlled by means of a sinusoidal voltage superimposed on a varying (negative or positive) direct voltage (vertical control). The sinusoidal grid voltages must be delayed by 90° with respect to the anode voltages. This is achieved by means of the phase-shifting networks C_4R_7 and C_8R_8 . The varying direct voltage is supplied by a special rectifier comprising the four selenium cells *Sel* and the ECC 40 double triode V_1 .

Assuming the grid voltage of both triode sections of the ECC 40 to be identical, the anodes of both triodes will then also be at the same potential. By suitable choice of the working points of these tubes this potential may be made negative with respect to the centre tap of the right-hand secondary of the transformer Tr_c . Both thyratrons Th_3 and Th_4 are then almost completely blocked, i.e. current is passed only during very short intervals of each half cycle. Since the small currents then flowing through Th_1 and Th_2 are identical the motor shaft will not run in either direction but vibrate slightly, which is conducive to rapid starting of the motor.

If the grid of the left ECC 40 section is made more negative its current will decrease so that the voltage at its anode will rise, but the resulting reduction of the voltage drop across the cathode resistor R_{13} will decrease the negative grid voltage of the other triode section, so that the anode current of the latter increases and the voltage at its anode decreases. The increased voltage at the anode of the left ECC 40 section will cause the ignition point of thyatron Th_4 to be advanced, while Th_3 is completely blocked.

Since these thyratrons are a.c. fed by means of transformer Tr_5 , current will then flow through Th_4 during part of each cycle, giving rise to voltage peaks across the secondary of the transformer Tr_2 which ignite thyatron Th_2 .

The latter is also a.c. fed, so that the mean value of the current flowing through this tube and the armature of the servo-motor is determined by the instant at which Th_2 is ignited in each cycle.

When the negative grid voltage of the left ECC 40 section is further increased the ignition point of the thyratrons Th_4 and Th_2 is obviously advanced, so that the mean value of the current flowing through the armature is also increased.

If the grid of the left ECC 40 section is made less negative than that of the other section the reverse will obviously occur, the ignition point of thyratrons Th_3 and Th_1 being advanced.

The grid of the right ECC 40 section is given a fixed potential by means of the voltage divider R_{14} , R_{15} , R_{16} , whilst the potential of the other grid is determined by the difference between the voltage taken from R_{15} and the voltage drop across the load resistor R_{17} of the EF 40 pre-amplifying tube V_2 . The potentiometer R_{15} is so adjusted that the two grids of the ECC 40 tube have the same potential when the quantity of light impinging on the

photocell and hence the current flowing through R_{17} are of their average value.

The capacitor C_{11} is required to introduce a time delay in the circuit and its value should be determined experimentally.

The circuit has the advantage that the motor is braked very efficaciously if the illumination of the photocell is changed correspondingly. This is because of the kinetic energy fed back to the mains in form of electric energy during the braking process by means of one thyatron which is operated as inverter during this period.

This may be illustrated by fig. 15-6. In the upper part (a) the shape of output voltage is shown, thyatron Th_1 being ignited with firing angle φ_1 and thyatron Th_2 being non-conducting. The motor then runs in one direction producing a counter electromotive force E which is in opposition to the positive half cycles of the mains voltage (with respect to Th_1). A current I therefore flows and supplies the motor with electrical energy.

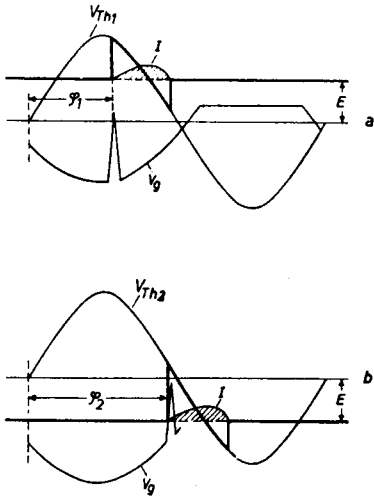


Fig. 15-6. Rectifier and inverter action of tubes Th_1 and Th_2 .



Fig. 15-7. Experimental equipment corresponding to the circuit of fig. 15-5.

Assuming now that a medium illumination of the photocell takes place so that both thyratrons receive igniting peaks with approximately 180 degrees phase delay during each cycle. Obviously Th_1 will be now completely blocked, but Th_2 is able to ignite with a firing angle φ_2 , since the counter electromotive force still present augments the anode voltage of this tube (fig. 15-6b). A current therefore flows, but now in the opposite direction, and since the polarity of E remains unchanged, electrical energy is fed back into the mains. The kinetic energy of the motor is, of course, reduced by the same amount, thus producing a very effective braking action.

The construction of an experimental set according to fig. 15-5 is shown in fig. 15-7.

Variable High-Speed Drive

Variable high-speed drives are often required for certain tools, such as grinding machines etc. The speeds concerned are in the order of 20 000 to 120 000 revolutions per minute. A suitable method of obtaining such high speeds is the use of a.c. motors which are supplied by an alternating voltage of correspondingly variable high frequency. Three-phase asynchronous motors have been developed for speeds up to 120 000 r.p.m. which are fed by a supply the frequency of which is variable between 1000 and 2000 cycles per second. For this three-phase supply a special electronic device is used, the circuit of which is described below.

The variable frequency is generated by a so-called RC oscillator the operation of which may be explained by fig. 15-8. Every oscillator consists of two main parts, an amplifying element and a frequency-determining element. The former is generally an amplifier tube the output voltage of which is partially fed-back in correct phase to the input (see Chapter 1). The frequency-determining element is introduced to obtain the correct phase relation between the output and input voltages at one given frequency. Usually an oscillating circuit consisting of a capacitance and an inductance is provided, but combinations of capacitances and ohmic resistors can also be used. In this case a network consisting of the capacitor C_1 and resistor R_1 connected in series, and capacitor C_2 shunted by resistor R_2 is employed.

High-vacuum tubes V_1 and V_2 produce alternating voltages which are phase-shifted by 180 degrees with respect to their input voltages. It therefore follows that the output voltage V_o appearing across the anode resistor R_{a2} of the second amplifier tube V_2 is in turn in phase with the input voltage V_i , which is applied to the grid of the first tube V_1 .

The network divides the output voltage V_o into two parts V' and V'' , V' appearing across C_2 and R_2 connected in parallel and representing the input voltage V_i . Obviously oscillations will be generated if the gain produced by V_1 and V_2 is equal to the attenuation caused by the network. Moreover,

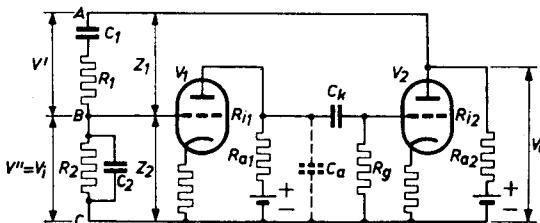


Fig. 15-8. RC oscillator circuit.

V'' must be in phase with V' , which is the case for one certain frequency. The impedance of C_1 and R_1 connected in series is

$$Z_1 = R_1 + \frac{1}{j\omega C_1}, \tag{15.1}$$

while the impedance of C_2 and R_2 connected in parallel is

$$Z_2 = \frac{\frac{R_2}{j\omega C_2}}{1 + \frac{R_2}{j\omega C_2}}. \tag{15.2}$$

Obviously the voltages across these impedances are in phase if the quotient Z_1/Z_2 is real. As can easily be shown, in these circumstances the following condition is fulfilled:

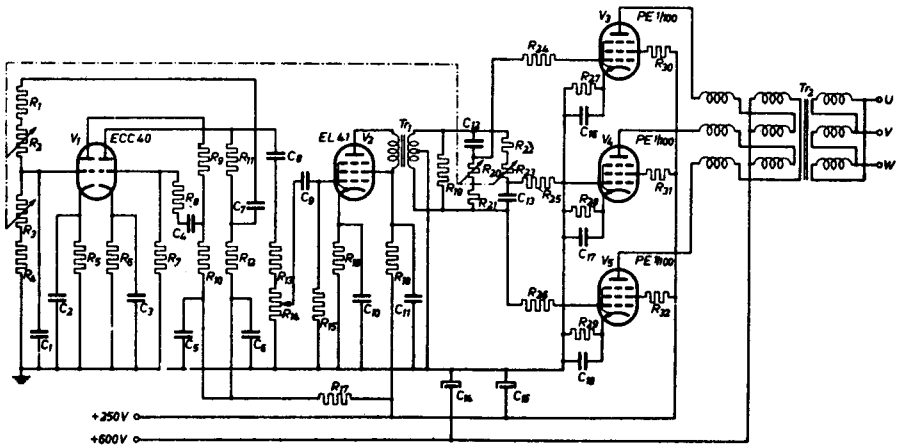


Fig. 15-9. Complete circuit of a three-phase variable-frequency generator.

$R_1 = 68 \text{ k}\Omega$	$R_{18} = 1 \text{ k}\Omega$	$C_2 = 25 \text{ }\mu\text{F}$
$R_2 = 0.1 \text{ M}\Omega$	$R_{19} = 2.2 \text{ k}\Omega$	$C_3 = 25 \text{ }\mu\text{F}$
$R_3 = 0.1 \text{ M}\Omega$	$R_{20} = 10 \text{ k}\Omega$	$C_4 = 0.01 \text{ }\mu\text{F}$
$R_4 = 68 \text{ k}\Omega$	$R_{21} = 10 \text{ k}\Omega$	$C_5 = 0.5 \text{ }\mu\text{F}$
$R_5 = 3.3 \text{ k}\Omega$	$R_{22} = 10 \text{ k}\Omega$	$C_6 = 0.5 \text{ }\mu\text{F}$
$R_6 = 1 \text{ k}\Omega$	$R_{23} = 10 \text{ k}\Omega$	$C_7 = 820 \text{ pF}$
$R_7 = 0.1 \text{ M}\Omega$	$R_{24} = 1 \text{ k}\Omega$	$C_8 = 0.01 \text{ }\mu\text{F}$
$R_8 = 0.82 \text{ M}\Omega$	$R_{25} = 1 \text{ k}\Omega$	$C_9 = 0.01 \text{ }\mu\text{F}$
$R_9 = 82 \text{ k}\Omega$	$R_{26} = 1 \text{ k}\Omega$	$C_{10} = 25 \text{ }\mu\text{F}$
$R_{10} = 22 \text{ k}\Omega$	$R_{27} = 230 \text{ }\Omega$	$C_{11} = 50 \text{ }\mu\text{F}$
$R_{11} = 2.7 \text{ k}\Omega$	$R_{28} = 230 \text{ }\Omega$	$C_{12} = 0.01 \text{ }\mu\text{F}$
$R_{12} = 2.7 \text{ k}\Omega$	$R_{29} = 230 \text{ }\Omega$	$C_{13} = 3900 \text{ pF}$
$R_{13} = 1 \text{ M}\Omega$	$R_{30} = 100 \text{ }\Omega$	$C_{14} = 16 \text{ }\mu\text{F}$
$R_{14} = 1 \text{ M}\Omega$	$R_{31} = 100 \text{ }\Omega$	$C_{15} = 16 \text{ }\mu\text{F}$
$R_{15} = 0.68 \text{ M}\Omega$	$R_{32} = 100 \text{ }\Omega$	$C_{16} = 50 \text{ }\mu\text{F}$
$R_{16} = 170 \text{ }\Omega$		$C_{17} = 50 \text{ }\mu\text{F}$
$R_{17} = 18 \text{ k}\Omega$	$C_1 = 820 \text{ pF}$	$C_{18} = 50 \text{ }\mu\text{F}$

$$jR_1R_2 - \frac{j}{\omega^2 C_1 C_2} = 0, \quad (15.3)$$

or

$$R_1R_2 = \frac{1}{\omega^2 C_1 C_2}. \quad (15.4)$$

Thus the resonant frequency of the RC -oscillator is:

$$\omega_0 = \frac{1}{\sqrt{R_1 R_2 C_1 C_2}}. \quad (15.5)$$

The complete circuit of a three-phase variable frequency generator is given in fig. 15-9. It consists of a frequency-determining oscillator stage, a phase-shifting circuit and a three-phase power stage. An RC oscillator circuit is used, the frequency of which is adjustable within a range of 1000 to 2000 c/s, the output voltage remaining constant.

A double triode V_1 (ECC 40) is used. In order to excite the three-phase power stage three voltages, phase-shifted about 120 degrees with respect to each other, are derived from the oscillator output voltage. For this purpose an amplifier stage using tube V_2 (EL 41) is provided, the anode circuit of which contains the primary winding of transformer Tr_1 . The secondary winding of Tr_1 forms together with C_{12} , $R_{20} + R_{21}$, and C_{13} , $R_{23} + R_{22}$ two phase networks producing two alternating voltages phase-shifted about 120 degrees with respect to each other. The third voltage is taken immediately from the transformer secondary.

Since the frequency of the oscillator is to be variable, the resistors of the phase determining networks R_{20} and R_{23} have to be altered simultaneously with the frequency determining resistors R_2 and R_3 of the RC -oscillator in order to secure the 120° phase shifting of the three output voltages throughout the frequency range. For this reason the four resistors are ganged.

The power stage is formed by three transmitting pentodes V_3 , V_4 , V_5 (PE 1/100). Their anode circuits contain the primary windings of the output transformer Tr_2 . The secondary windings are connected to the motor. In order to compensate slight non-symmetries of the primary circuits the transformer primary windings are connected in zig-zag.

The device is designed for an output power of approximately 100 VA at an output voltage of 3×220 volts.

To obtain a sinusoidal output voltage the three power tubes are operated in Class A. The necessary grid bias is produced by cathode resistors R_{27} , R_{28}

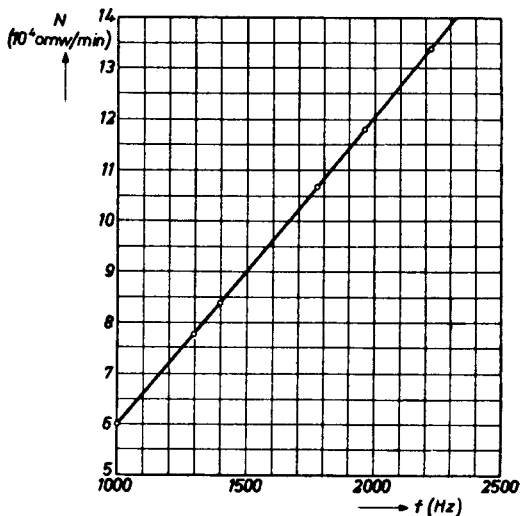


Fig. 15-10. Diagram showing speed of the motor against the generator frequency.

and R_{29} . All supply voltages are delivered by built-in power packs not shown in the circuit diagram. The direct anode supply voltages of the PE 1/100 tubes are 600 volts, the other direct supply voltages 250 volts.

Fig. 15-10 is a graph presenting the speed of the motor as a function of the generator frequency under no-load conditions. Fig. 15-11 illustrates the construction of an experimental set.

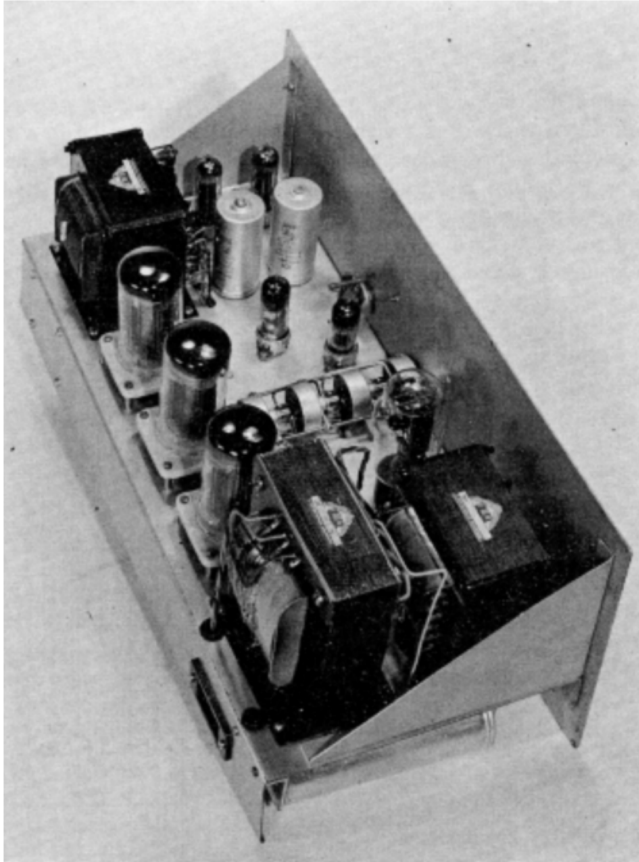


Fig. 15-11. Experimental model of the three-phase generator.

Temperature Control

A frequently recurring problem is the temperature control of furnaces, chemical liquids, drying rooms etc. The heat will be generated by electrical energy, the temperature being controlled by hand or automatically. For control by hand variable heating rheostats or sometimes variable series resistors, or, for larger powers, auto-transformers have hitherto been used. Series resistors have the disadvantage that a large amount of energy is wasted in heating the resistor; variable rheostats do not permit very accurate regulation and often

require large and complex operating mechanisms. Auto-transformers are somewhat expensive, and moreover, their comparatively large bulk and weight are often serious disadvantages.

By electronic means, however, control can be achieved without involving serious power loss, and at the same time offering the advantage of very easy control. Remote control by tiny switches, potentiometers etc. is also possible. The weight and the dimensions of the electronic device are only small, thus it can easily be installed. Below are given some typical examples of electronic heat control.

The basic circuit of a simple device to be operated by hand, suitable for powers up to some kilowatts, is shown in fig. 15-12. The main circuit contains the load which is assumed as a transformer with heating element in the secondary circuit. Two inverse-parallel connected thyratrons T_1 and T_2 are employed. The useful power which can be delivered by such a circuit is given by

$$P_o = \frac{\pi}{\sqrt{2}} \cdot I_a V \cdot \cos \varphi, \quad (15.6)$$

I_a being the average anode current per tube, V the mains supply voltage (r.m.s. value) and $\cos \varphi$ the power factor of the load.

Both tubes are ignited alternately in each cycle by an alternating voltage appearing in the grid circuits of the tubes. This voltage is produced by the secondary windings of transformer Tr_2 ; it can be phase-shifted with respect to the anode voltage by means of a phase network consisting of Tr_3 , R_5 and the inductor L . It is therefore a horizontal control, the firing angle being adjustable by means of variable resistor R_5 within the range of nearly zero to 180 degrees. Adjustment of the current flowing through the circuit from the maximum value down to practically zero is thus possible. The variable resistor R_5 , of course, can be installed at any convenient position for remote control.

Often it is required to control loads of considerably higher powers up to several hundred kVA. In these cases ignitron tubes connected in a similar circuit are well suitable. Fig. 15-13 shows the basic circuit, the auxiliary circuitry for preheating the thyratrons and rectifying tubes, the water flow control etc. being omitted. A furnace load is assumed the power factor of which is nearly unity, i.e. the voltage and current waves are practically in phase.

In this case, however, the firing circuit used in fig. 5-8 cannot be employed

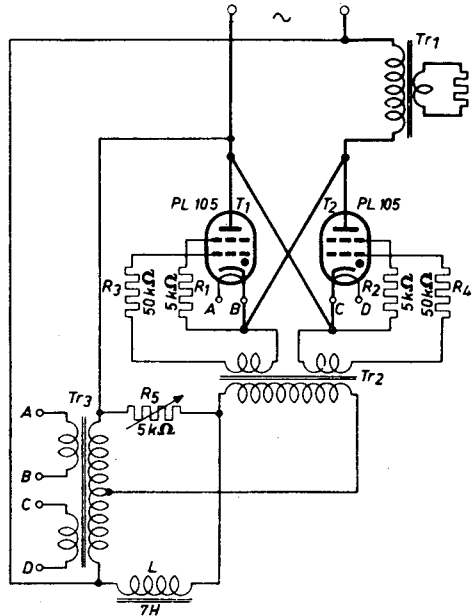


Fig. 15-12. Basic circuit of a simple hand-operated temperature control device for small powers.

because at the beginning of each current half wave the instantaneous value of the anode voltage would not be large enough to ensure safe operation of the ignitor. Another firing circuit is therefore provided using the discharge of a capacitor which delivers enough electrical energy to start the ignitor, independent

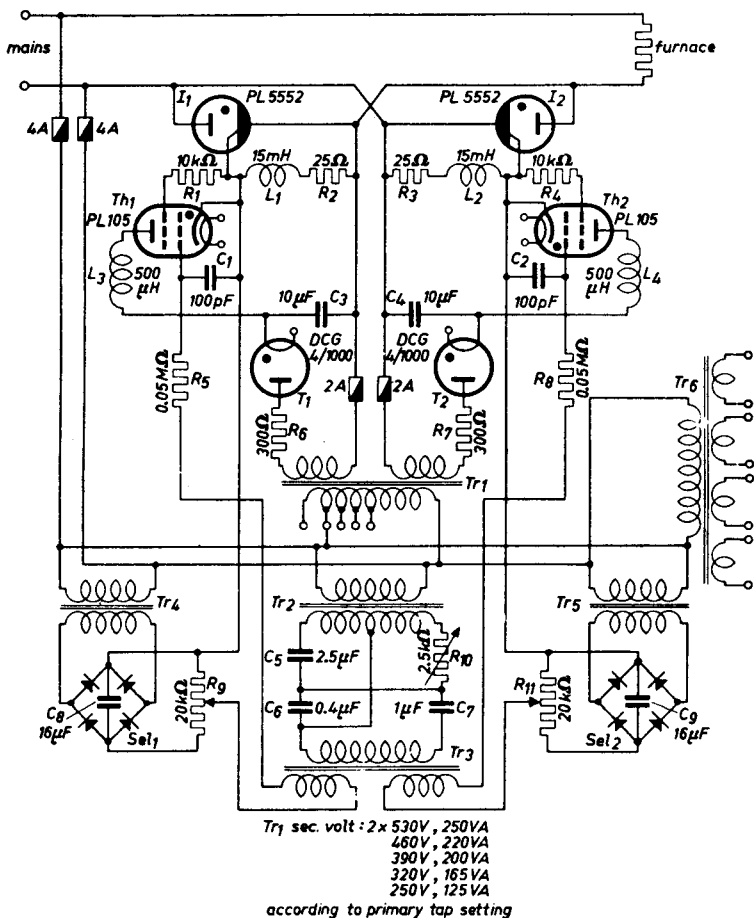


Fig. 15-13. Circuit of a hand-operated temperature control device for larger powers.

of the instantaneous value of the voltage at the main anode, i.e. independent of the firing angle.

Capacitor C_3 is charged during the negative half cycle of the alternating voltage via the rectifying tube T_1 up to a potential of, say, 500 volts. This voltage appears across the series connection of the thyatron Th_1 and the ignitor of the ignitron I_1 . When Th_1 is fired by a positive impulse applied to its first grid, the capacitor C_3 rapidly discharges through Th_1 and the ignitor, starting the main discharge, even if the instantaneous value of the anode voltage of I_1 is

as low as 20 or 30 volts. The inductance L_3 tries to maintain the discharge current, thus causing a short-time negative charge of C_3 which is sufficient to extinguish the thyatron Tb_1 . In the following half cycle, I_2 is fired correspondingly, by discharge of C_4 , and C_3 is recharged to its original potential. The thyratrons Tb_1 and Tb_2 are horizontally controlled, the positive peaks being produced by peaking transformer Tr_3 .

The variable resistor R_{10} permits variation of the firing angle and thus regulation of the current flowing through the ignitron tubes and the load. Transformer Tr_1 has a number of taps on its primary winding in order to permit the amount of electrical energy stored in the capacitors C_3 and C_4 to be adjusted for correct firing of the ignitrons.

Temperature Regulation

Another problem frequently arising is the stabilising of a temperature within two predetermined limits. The circuit of an automatic device for this purpose is given in fig. 15-14.

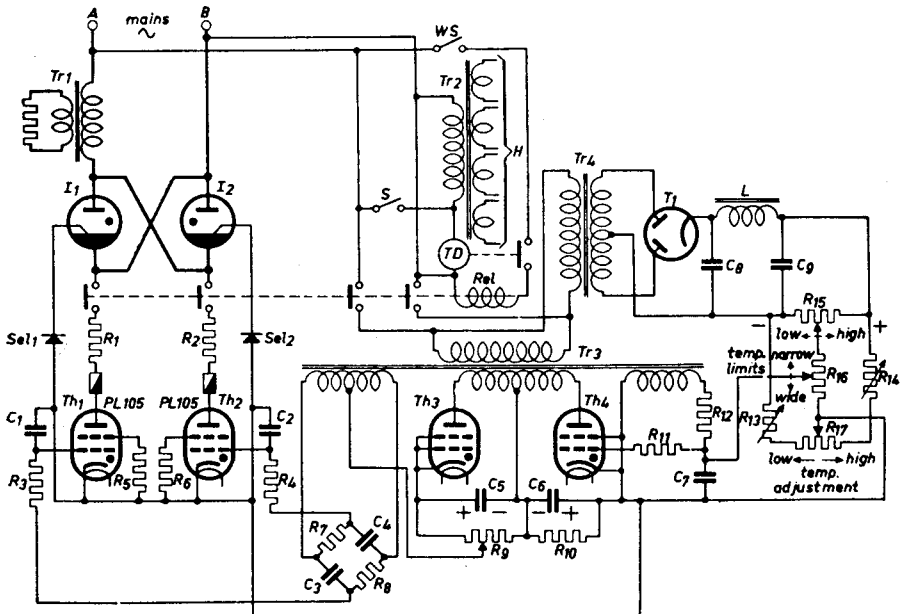


Fig. 15-14. Automatic temperature regulating circuit.

The temperature detecting element, which is not shown, operates the slider of the variable resistor R_{15} and responds to the output voltage of a thermocouple mounted in the furnace, via an interconnected amplifier stage. The position of the slider of R_{15} therefore indicates the actual temperature of the furnace. It may be assumed that the slider is moved to the positive side if the temperature rises. The desired temperature is pre-set by the setting of potentiometer R_{17} , the dial of which is calibrated in temperature degrees.

Assuming for the moment that this value is considerably higher than the actual temperature, a voltage difference then arises across resistor R_{16} of such polarity that the control grid of the PL2D21 thyatron Th_4 becomes negative with respect to the cathode. In addition to this negative direct voltage an alternating voltage is applied to the grid which appears across the capacitor C_7 . If the value of C_7 is chosen so that the a.c. resistance of C_7 is small compared with R_{12} , the voltage across C_7 is phase-lagged by nearly 90 degrees behind the anode voltage as illustrated in fig. 15-15. Obviously this constitutes a vertical control for influencing the firing angle of Th_4 , but for the moment this tube is completely blocked since its control grid is highly negative because of the voltage drop across R_{16} . Consequently no voltage appears across capacitor C_6 .

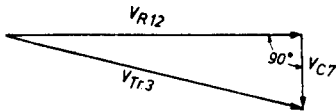


Fig. 15-15. Vector diagram corresponding to fig. 15-14.

However, thyatron Th_3 (PL2D21) connected as diode, can ignite and its anode current causes a voltage drop across resistor R_9 , charging C_5 with the polarity indicated. A part of this voltage is applied to the grid circuit of thyratrons Th_1 and Th_2 with positive sign. These tubes alternately cause ignition of the ignitrons I_1 , I_2 similarly as in the circuit shown in

fig. 5-7; in this case, however, they are so connected that their cathodes are at the same potential, thus making the use of vertical control possible.

Assuming that terminal A of the mains is positive, then ignitron I_1 is able to pass current if its ignitor is operated. If the contacts of relay Rel have been closed after expiration of the preheating time, the anode of Th_2 receives a positive voltage. When this tube is ignited a current surge flows from A via Th_2 , the selenium rectifier $Se1_1$ and the ignitor of I_1 to B , starting the main discharge of I_1 . Selenium rectifier $Se1_2$ prevents part of the current from flowing through the ignitor of I_2 in the reverse direction, thus avoiding damaging of the igniting rod.

In a corresponding manner ignition of I_2 occurs in the following half cycle of the mains voltage. Tubes Th_1 and Th_2 receive alternating grid voltages, phase-lagged by 90 degrees behind the anode voltages by means of the phase-shifting networks R_7 , C_4 and R_8 , C_3 ; moreover, in the case under consideration, a positive direct voltage is present, as already shown. Both tubes will therefore ignite in each cycle without phase delay, permitting the full heating current to flow through the ignitron tubes.

The furnace temperature is thereby gradually increased, decreasing correspondingly the voltage drop across resistor R_{16} . Shortly before reaching the desired temperature value the vertical control of Th_4 begins to become effective, allowing this tube to ignite at first with large firing angle which, however, will be gradually reduced. A slowly increasing voltage drop arises therefore across R_{10} charging capacitor C_6 . This voltage is in opposition to the voltage taken from variable resistor R_6 , the slider of which is set so that the values of both voltages are equal when the desired temperature is reached. No direct voltage is then present in the grid circuits of Th_1 and Th_2 , and these tubes ignite with approximately 90 degrees phase delay, causing a reduced current to flow through the ignitrons which should be sufficient to replace the heat losses arising in the furnace.

However, if this current is too large, the temperature continues to rise, and

the current further decreases in the manner described, until a point is reached somewhat above the desired temperature level, when Tb_4 ignites without phase delay. The voltage across C_6 now being much larger than the voltage across R_9 , thyratrons Tb_1 , Tb_2 are completely blocked. Consequently, the temperature falls and the tubes begin to pass current again, the operation continuing until finally a condition of equilibrium is reached.

Resistor R_{16} serves for adjusting the limits between which the furnace temperature may swing. Obviously this range will become narrower if the voltage difference taken from R_{16} is increased. Resistors R_{13} and R_{14} serve for calibration of the dial of R_{17} .

S is the main switch, TD a time relay which is needed in order to provide the preheating time of the thyratrons. Contact WS is the water flow switch which is closed if the flow of cooling water for the ignitrons is sufficient.

Fig. 15-16 shows the construction of an electronically controlled temperature regulating device suitable for heating powers up to 1.5 kW.



Fig. 15-16. Electronic temperature regulating device for heating powers up to 1.5 kW.

16. ELECTRONIC CONTROL OF RESISTANCE WELDING

One of the most important fields of application for electronics is the automatic control of resistance welding machines. It has already been shown in Chapter 5 that the primary circuit of a welding transformer can be closed or opened by means of two inverse-parallel connected ignitron tubes instead of by electro-mechanical contactors. Particularly in large power welding machines the introduction of electronically operated devices is essential since the risk of burning of the contacts of mechanical switches reduces their useful life and reliability and thus results in deterioration of quality of the final product. Another very serious disadvantage of many mechanical contactors is that it is not possible to ensure that the flow of welding current always commences at a point in the cycle where the instantaneous value of the current is zero.

This is illustrated by fig. 16-1. If the welding current flow is started somewhat earlier or later than the point corresponding to the angle of lag of the welding transformer, large transient currents occur which cause not only burning of the welding electrodes or the workpiece, but will also in some cases blow the mains fuses. This danger can be completely avoided if ignitron contactors are used, since it is then possible, by correct adjustment of the firing angle, to start the current flow exactly at the right moment (synchronous timing). Moreover, the heat supplied to the workpiece can easily be controlled by gradual increase

or reduction of the firing angle. Since the energy converted into heat during a weld is

$$E = I^2 R t, \tag{16.1}$$

E depends not only on the weld time t , but also on the current I .

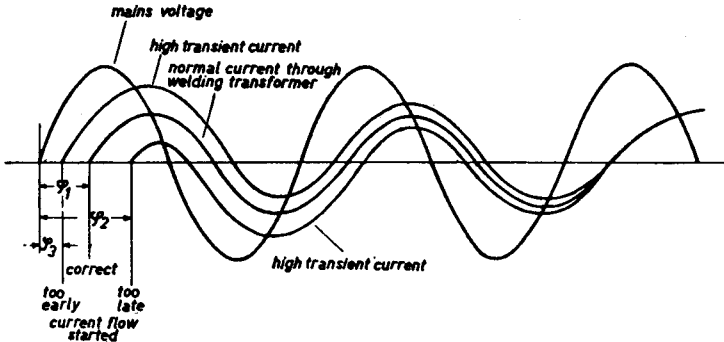


Fig. 16-1. Transient currents arising when current flow is started before or after the power factor angle of the welding transformer.

Fig. 16-2 shows how the current flowing through the ignitron tubes and the primary of the welding transformer can be adjusted almost down to zero by increasing the firing angle of the tubes. A similar effect cannot be achieved if mechanical contactors are used: if the contact should be closed at a later time than that corresponding to the $\cos \varphi$ the current would indeed decrease in this half cycle, but it would also increase greatly in the following half cycle (see fig. 16-1). Furthermore it must be borne in mind that the heavy current flowing through an ignitron tube always continues until the end of the half cycle, becoming gradually zero. If, however, the current is interrupted by means of a

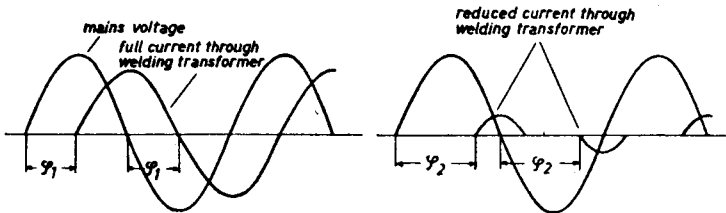


Fig. 16-2. Controlling the current by varying the firing angle of the ignitron tubes (heat control).

mechanical contactor, this will usually occur within a half cycle, thus causing a considerable voltage surge across the transformer primary which might seriously damage the equipment.

Simple Electronic Welding Contactor

Fig. 16-3 shows the circuit of a very simple electronic welding contactor for small powers, in which no provision has been made to control the current by

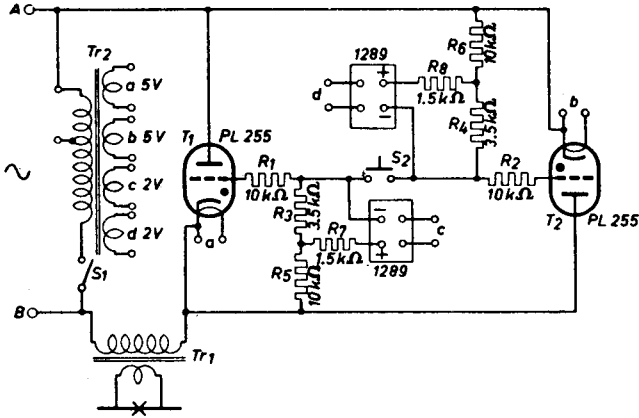


Fig. 16-3. Simple electronic contactor.

varying the firing angle. This apparatus is equipped with two thyristors T_1 and T_2 (PL 255), which can supply a continuous current with an r.m.s. value of 27 A. By means of two auxiliary rectifiers (Type No. 1289), a negative bias of approximately 40 V is applied to the grid of these tubes, so that they remain extinguished as long as the pedal switch S_2 remains open. Upon closure of S_2 , however, a voltage divider is formed by R_6 , R_4 , R_3 and R_5 . It will be assumed that point A of the mains is positive at this instant. A positive voltage, which is equal to the voltage drop across R_3 , R_5 and exceeds the negative voltage

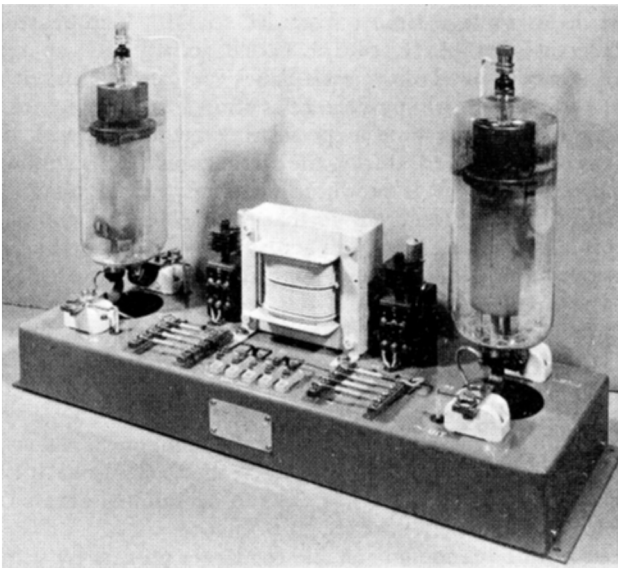


Fig. 16-4. Example of an electronic contactor according to fig. 16-3.

supplied by the rectifier 1289, is then applied to the grid of tube T_1 , so that the latter is ignited. During the following half cycle of the mains voltage T_2 is ignited in a similar way, and so-on, until switch S_2 is opened again. The current flowing through S_2 has a value of a few milliamps only, so that there is no question of the contacts being damaged due to arcing.

Fig. 16-4 shows an example of such an electronic contactor. The two tubes are supported exclusively by their filament connections. The chassis is provided with an aperture below each tube to ensure a good air circulation.

It will be clear that this electronic contactor is not only suitable for controlling spot welders, but can be used in all cases where a high switching frequency is required, for example for intermittently switching neon sign boards on and off.

Electronic Control of Small Spot Welders

Fig. 16-5 shows the circuit of a different apparatus for controlling small spot welders. Two thyratrons T_1 and T_2 connected in anti-parallel are incorporated in the primary circuit of the welding transformer Tr_1 . Thyratrons type PL 105 are used for this purpose; when they are connected in anti-parallel, they can supply a current with an r.m.s. value of 55 A at a duty cycle of less than 10%.

In the control grid circuits of these tubes the secondary windings of a peaking transformer Tr_3 (Type No. 84590) and two direct voltage sources for providing the negative grid bias are included. The primary of Tr_3 is connected to the output of a push-pull amplifier equipped with two tubes E 80 L. The control voltage for these tubes is taken from the resistors R_5 and R_6 shunted across the output of a phase-shifting network. Tr_4 is a normal push-pull output transformer for a power of 4 W, having a primary impedance of $2 \times 7 \text{ k}\Omega$ at 50 c/s and a transformer ratio of $2 \times 2 : 1.5$. The variable resistor R_{17} of the phase-shifting network renders it possible to shift the control voltage with respect to the mains voltage almost from 10° to 170° . The phase angle of the pulses at the secondaries of Tr_3 can be varied accordingly, so that the point at which the thyratrons are ignited each half cycle can be delayed. In this way it is possible to adjust the welding current as shown in the diagram of fig. 16-2.

For adjusting the welding time a special timer equipped with the thyatron T_3 (PL2D21) has been provided. During the off time this tube is conducting, so that a voltage drop of 50 to 70 V is produced across R_{13} ; this voltage drop is used for biasing the grids of V_1 and V_2 ; so that these tubes are completely cut off. A voltage drop is, moreover, produced across $R_{15} + R_{16}$, so that C_{10} is charged, and a negative control grid voltage is applied to the thyatron T_3 ; this tube is, however, not extinguished, since it is d.c. fed. For extinguishing this tube its anode circuit must be interrupted during a few milliseconds by means of the pedal switch S_3 , after which C_{10} will be discharged via R_{15} and R_{16} . When the voltage across C_{10} has dropped to approximately 2 V, T_3 will re-ignite, the welding time thus being terminated. The duration of this time can be varied within certain limits by means of R_{15} . With the component values indicated in the circuit diagram, welding times ranging from 0.02 sec to 0.12 sec can be obtained. By means of S_2 the capacitor C_{11} can be shunted across C_{10} to extend the range from approximately 0.12 sec to 0.75 sec.

The apparatus can be rendered suitable for larger powers by using thyratrons PL 255 instead of thyratrons PL 105. The maximum permissible current flowing through the tubes connected in anti-parallel then has an r.m.s. value of 110 A

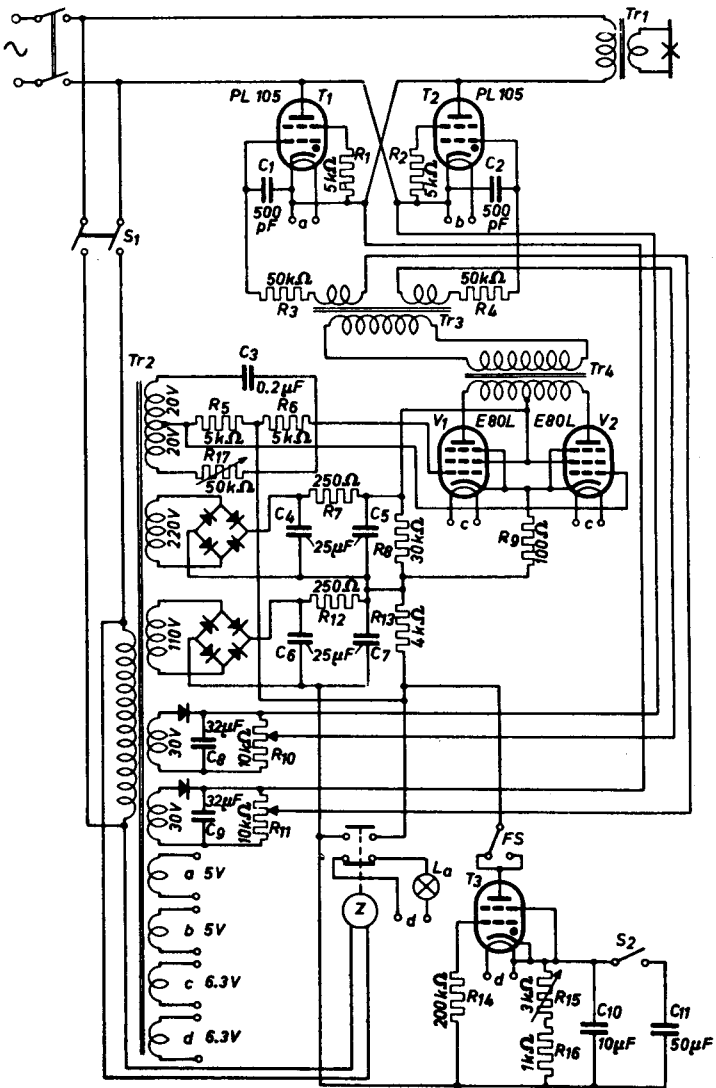


Fig. 16-5. Electronic control device for small spot welders.

at a maximum duty cycle of 10%, as shown by the characteristics of fig. 16-6. For a given adjustment the duty cycle is given by the relation:

$$\text{duty cycle} = \frac{T}{t_0} \cdot 100\%, \quad (16.2)$$

in which T represents the duration of the welding time and t_0 the integration time of the tubes employed. For the tubes PL 57, PL 105 and PL 255 this time is 5 sec at a duty cycle from 0 to 50%, and 15 sec at a duty cycle from 50 to 100%.

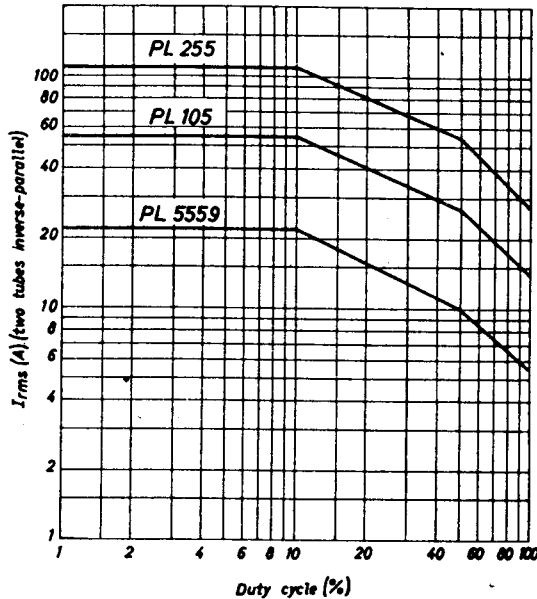


Fig. 16-6. Power characteristics (welding current as a function of the duty cycle) for different types of thyratron in anti-parallel (two tubes).

Consequently, a current with an r.m.s. value of for example 110 A can be supplied by two tubes PL 255 at a maximum welding time of 0.5 sec followed by an off time of at least 4.5 sec. With PL 105 thyratrons a current with an r.m.s. value of 55 A is permissible during only 0.5 sec at a maximum duty cycle of 10%.

Four-Time Spot Welding Control

As already mentioned in Chapter 12, in welding machines it is not only the actual weld time which is of importance, but also the "squeeze" time required for positioning of the workpiece and pressing the electrodes together. After a weld has been made the electrodes still remain closed for a while in order to allow the molten metal to cool down; this time interval is the "hold" time. Finally, the "off" time is required in order to re-open the electrodes and remove the workpiece. These four successive time intervals must be repeated periodically.

For this purpose an electronic device, the circuit of which is illustrated in

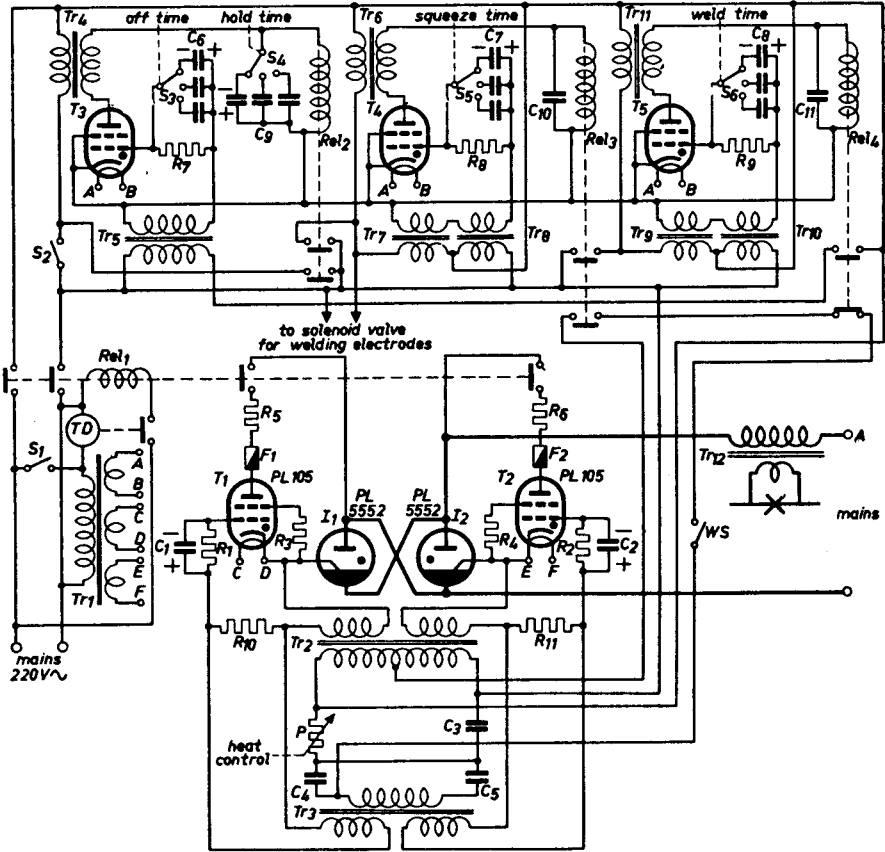


Fig. 16-7. Electronic timing control circuit suitable for welding machines.

fig. 16-7, can be used. The device is switched on by switch S_1 , and thyristors T_1 to T_5 are preheated. After the expiration of the preheating time the time relay TD closes its contact, energising relay Rel_1 . The device is now ready for operation, provided sufficient water for cooling of the ignitrons is flowing so that contact WS of the water flow switch is closed.

The welding current flows from point P through the welding transformer Tr_{12} and the two inverse-parallel connected ignitrons I_1 and I_2 back to point Q . Ignitrons I_1 and I_2 are ignited alternately by the thyristors T_1 and T_2 , a short-time current surge of approximately 40 amperes flowing for about 100 microseconds through each thyristor and the ignitor for the corresponding ignitron. This will happen each time the thyristor is fired by a positive voltage impulse occurring in the grid circuit.

As soon as the ignitron is ignited, i.e. the main discharge between the liquid mercury pool cathode and the anode has started, the voltage between these electrodes falls to the value of the arc voltage (about 14 volts). At this moment

the anode of the thyatron receives a still lower potential and the tube is extinguished. If, however, the ignition of the ignitron does not occur for some reason, the full igniting current would flow continuously during the whole half cycle through the thyatron, causing damage to the tube. Fuses are therefore connected in the anode leads of the thyatron tubes which are sluggish enough to withstand the short-time current surges, but will blow if this current continues to flow.

The grid circuits of the tubes T_1 and T_2 are similar, so that it will suffice to describe only one of them, for instance that of T_1 . The grid circuit consists of the protective resistor R_1 which is by-passed by capacitor C_1 , a secondary winding of transformer Tr_2 producing a voltage which is phase-shifted about 180 degrees with respect to the anode voltage, and a secondary winding of peaking transformer Tr_3 . During the negative half cycle of the anode voltage a positive grid current flows due to the voltage produced by Tr_2 , and the voltage drop arising across R_1 charges capacitor C_1 with the polarity shown. In the following half cycle this voltage is added to the (now negative) voltage half cycle produced by Tr_2 , so that tube T_1 does not pass current unless a positive voltage peak is produced by Tr_3 . The primary winding of Tr_3 is connected via contacts of Rel_3 , Rel_4 and WS to a phase-shifting network of the type described in an earlier chapter. The phase shift is obtained by means of variable resistor P , and the firing angle of the ignitrons I_1 , I_2 is varied correspondingly.

The small PL 2D 21 thyratrons T_3 , T_4 , T_5 , which form part of the timing circuit, are for the moment extinguished. Transformers Tr_8 and Tr_{10} produce alternating voltages, which charge capacitors C_7 and C_8 with the polarity shown via the grid-to-cathode path of tubes T_4 and T_5 . If operating switch S_2 is now closed, tube T_3 ignites since there is no negative bias present in its grid circuit. Due to the anode current flowing through Rel_2 , capacitor C_9 is charged as shown. Rel_2 closes its contacts, by-passing S_2 and operating the solenoid valve for closing the welding electrodes.

At the same time tube T_4 receives anode voltage via transformer Tr_6 , and Tr_7 is connected to the mains, so that an additional alternating voltage appears in the grid circuit of T_4 the magnitude of which is equal to the voltage produced by Tr_8 , but is phase-shifted by 180 degrees. The two voltages therefore cancel each other, and only the negative voltage across C_7 is now present in the grid circuit, preventing T_4 from firing. However, C_7 discharges slowly through R_8 , thus producing the squeeze time. Selector switch S_5 allows capacitors of different values to be connected in the circuit, thus varying the squeeze time.

As soon as C_7 has discharged, thyatron T_4 ignites and energises Rel_3 to close the primary circuit of Tr_3 , and the weld time begins. Simultaneously T_5 receives anode voltage via Tr_{11} , and the voltage produced by Tr_{10} is cancelled by an equal voltage phase-shifted by 180 degrees which is generated by transformer Tr_9 . Thus in the grid circuit of T_5 only the negative voltage across C_8 appears which prevents T_5 from firing. The value of C_8 can be varied by selector switch S_6 . This capacitor discharges through R_9 and produces the weld time.

Subsequently T_5 ignites, energising relay Rel_4 , and the weld time ceases. At the same time transformer Tr_5 is switched on which produces a voltage in the grid circuit of T_3 , phase shifted about 180 degrees with respect to the anode voltage. Capacitor C_6 is charged via the grid-to-cathode path of the tube with the shown polarity. T_3 is thus immediately extinguished; relay Rel_2 does not

drop out, however, until capacitor C_9 , the value of which can be chosen by selector switch S_4 , has discharged. Thus the hold time is produced.

Finally Rel_2 drops out, operating the solenoid valve for opening the welding electrodes. Simultaneously T_4 is extinguished, relay Rel_3 drops out, and T_5 becomes non-conducting. Consequently Rel_4 drops out, disconnecting T_5 . Capacitor C_6 , the value of which can be preset by selector switch S_3 , now can discharge through R_7 , producing the off time.

If operating switch S_2 is still closed tube T_3 ignites again after the discharge

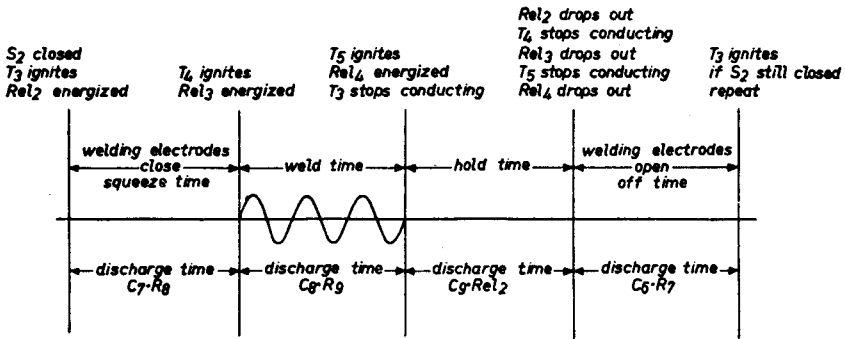


Fig. 16-8. Analysis of a complete welding cycle showing the four time intervals.

of C_6 , and the whole cycle repeats once more; otherwise the timer now remains inoperative. Thus, by momentarily closing S_2 a single time sequence can be started, whilst the cycle repeats continuously as long as S_2 is closed. The complete cycle comprising the four time intervals is illustrated in fig. 16-8. Instead of the ignitron tubes PL 5552 shown, other types of larger size can be inserted without changing the value of any other component.

Control of Seam Welding

Seam welding demands a very high degree of accuracy in the operation of the welding machines, and mechanical contactors can seldom meet these requirements. In addition, the mains voltage in industrial works is often subject to large fluctuations, especially when several welding machines are in operation simultaneously, and these fluctuations may seriously affect the quality of the weld. All these troubles, however, are eliminated if ignitron contactors, which are controlled automatically by electronic timing devices, are used in seam welding machines.

In fig. 16-9 the circuit of an automatic electronic welding controller is shown. The device includes an electronic timer which controls weld time and off time, both of which can be adjusted independently. A hand-operated circuit is also provided for gradual heat control by phase shifting the firing points of the ignitron tubes, and this device also serves for automatic compensation of mains voltage fluctuations. A special circuit prevents the risk of only one ignitron firing, which otherwise might occur if the firing angle is chosen too small. All processes are controlled by electronic means, and no mechanical relays or contactors are needed with the exception of a time relay for preheating the thyratrons etc.

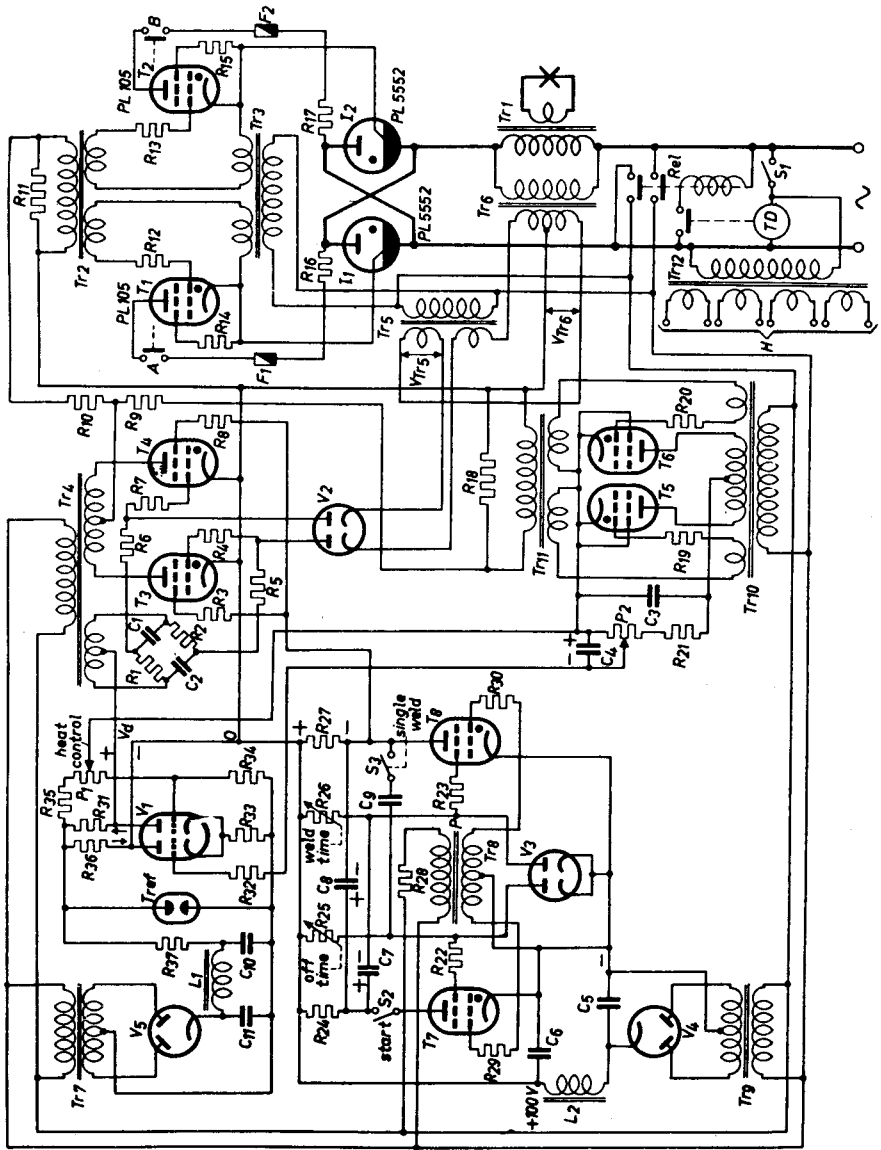


Fig. 16-9 Automatic control circuit for seam welding machines.

The main circuit contains the welding transformer Tr_1 and the inverse-parallel connected ignitrons I_1 and I_2 . For igniting these tubes two thyratrons T_1 and T_2 are provided. When switch S_1 is closed transformer Tr_{12} is connected to the mains, thus preheating all tubes. For the sake of simplicity, only one control filament transformer is shown in fig. 16-9; in practice it is usual to provide several separate filament transformers. After expiration of the preheating time the time relay TD closes its contact, energising relay Rel which connects the primaries of transformers Tr_3 , Tr_4 , Tr_5 , Tr_7 , Tr_8 , Tr_9 and Tr_{10} to the mains.

At the same time contacts A and B , also operated by Rel , are closed. As soon as thyatron T_1 or T_2 is ignited at any point within a half cycle the corresponding ignitron tube also ignites. The secondary windings of transformer Tr_3 are included in the grid circuits of T_1 and T_2 and produce voltages which are phase-shifted about 180 degrees with respect to the anode voltages of T_1 and T_2 . Moreover, the grid circuits contain the secondary windings of peaking transformer Tr_2 .

The anode and grid voltages of tubes T_1 and T_2 thus resulting are shown in fig. 16-10. Obviously this constitutes a horizontal control with phase-shifted impulses.

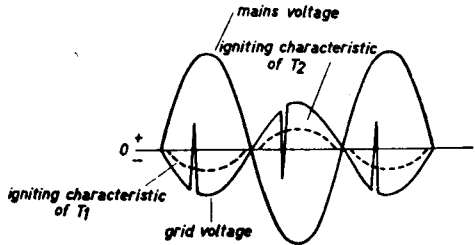


Fig. 16-10. Grid voltages of thyratrons T_1 and T_2 .

The primary winding of Tr_2 is connected to the output of a two-phase rectifier circuit consisting of transformer Tr_4 and the two small inert gas-filled tetrode thyratrons T_3 and T_4 . Obviously each time T_3 or T_4 ignites, a current impulse flows through the primary of Tr_2 , causing a voltage impulse in the secondary windings which appears in the grid circuits of T_1 and T_2 . If the ignition of T_2 and T_4 is caused to lag in phase by a certain angle with respect to the mains voltage, ignition of T_1 and T_2 and thus of I_1 and I_2 occurs with the same phase delay. The firing angle of tubes T_3 and T_4 is adjusted by vertical control; an alternating voltage phase-lagged by 90 degrees is applied to their control grids which is taken from a phase-shifting network consisting of R_1C_1 and R_2C_2 . Moreover a variable direct grid voltage V_d is superimposed, which may be adjusted between positive and negative values, thus causing variation of the firing angle over a range of nearly 0 to 180 degrees.

The variable voltage V_d appears between the anodes of the high-vacuum double triode V_1 . This tube is connected in a direct voltage amplifier circuit known as a "long-tailed pair". The anodes are supplied by a direct voltage which is held constant by a voltage reference tube T_{ref} . Two equal resistors R_{30} , R_{31} are inserted in the anode leads. The cathodes are connected to the negative H.T. terminal via a common resistor R_{33} , and the control grid of the right-hand triode system receives a certain potential from a voltage divider formed by resistors R_{35} , P_1 and R_{34} .

Assuming for the moment that the control grid of the left-hand triode system has the same potential, then equal currents flow through both systems, and there will be no voltage difference between the anodes. If the potential of the

left-hand grid is raised the anode current of the left-hand triode increases, and the anode potential decreases. Simultaneously the current flowing through the common cathode resistor is increased, causing decrease of the actual potential of the right-hand grid. This in turn causes increase of the potential at the right-hand anode, and a voltage difference with the shown polarity arises between the two anodes. Consequently the firing angle of the thyratrons T_3, T_4 is reduced. On the other hand, the ignition point of these tubes is delayed if the potential of the control grid of the left-hand triode system is lowered.

The transformer Tr_{10} together with the thyratrons T_5, T_6 form a rectifier circuit whose output voltage appearing across P_2, R_{21} is proportional to the mains voltage. A part of this output voltage is derived from P_2 and compared with a reference voltage taken from P_1 . The difference between these voltages is applied to the control grid of the left-hand triode system of the high-vacuum tube V_1 . If the mains voltage is reduced, for instance, by a momentary heavy load, the voltage derived from P_2 is equally reduced, and the potential of the left-hand control grid of V_1 is raised. This in turn causes a decrease of the firing angle of the thyratrons T_3, T_4 so that an increased welding current flows which compensates the lack of welding power caused by the reduced mains voltage. In this way uniformity of the welds is automatically secured, thus greatly contributing to the good performance of the welding machine.

By means of potentiometer P_1 the firing angle of the ignitrons and therefore the amount of heat applied to the workpiece can be adjusted. Potentiometer P_2 serves for adjusting the level of the automatic control, so that a certain firing angle is always maintained for compensating mains fluctuations if potentiometer P_1 is set for full heat.

Obviously the control of the mains voltage should be exercised only during the times of actual welding current flow. For this reason, tubes T_5, T_6 are grid-controlled in the same way as thyratrons T_1, T_2 by means of peaking transformer Tr_{11} , i.e. the output voltage of tubes T_5, T_6 decreases if the firing angle increases. Transformer Tr_{11} is excited similarly to transformer Tr_2 by the tubes T_3, T_4 . The auxiliary windings on transformer Tr_{10} serve for applying anti-phase alternating grid voltages to tubes T_5, T_6 ; they have a similar task to that of transformer Tr_3 (see fig. 16-10).

If two inverse-parallel connected ignitrons are ignited with a very small angle it may occur that the second ignitron fails to ignite. The reason is shown in fig. 16-11*a*. If the current flow of the first ignitron I_1 starts earlier than it should with respect to the power factor of the load, a transient current surge arises, the duration of which is more than 180 degrees. The ignitron ignited at point 1 therefore passes current until point 5. However, the firing impulse for the second ignitron I_2 occurs at point 2, but this tube is unable to ignite since only the arc voltage is present between anode and cathode until the first ignitron extinguishes.

This trouble can be eliminated by means of a special circuit which delays the firing impulse of the second ignitron until point 5, thus ensuring that the two ignitrons fire successively under all circumstances. For this purpose the control grid circuits of the thyratrons T_3 and T_4 include the secondary windings of transformers Tr_5 and Tr_6 as well as the diode systems of the high-vacuum tube V_2 . As shown in fig. 16-11*b* the voltage V_{Tr_5} is in phase with the anode voltage of tube T_4 . The voltage V_{Tr_6} is approximately twice the value of V_{Tr_5} .

and is in anti-phase to that voltage. Since the primary winding of Tr_6 is connected in parallel with the welding transformer, voltage V_{Tr_6} is present only if I_1 or I_2 passes current. The voltage appearing at the control grid of T_4 is equal to the potential of the right-hand anode of the rectifying tube V_2 and consists of the voltage V_d produced by tube V_1 and the superimposed alternating voltage produced by the RC phase-shifting network, which is phase-lagged

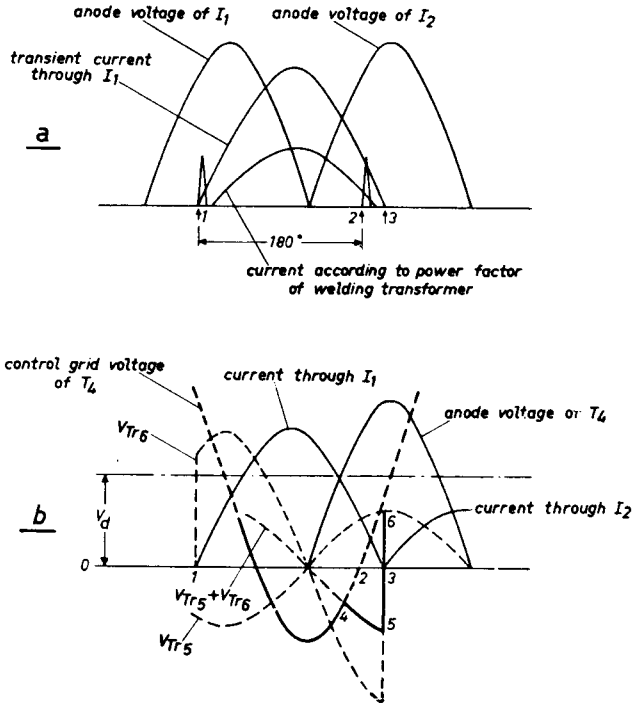


Fig. 16-11. Operation of the circuit for preventing the firing of one ignitron only.

by 90 degrees. The cathode potential of the right-hand diode system of tube V_2 is equal to the sum $V_{Tr_6} + V_{Tr_5}$, since I_1 has been assumed to be ignited. The right-hand diode system of V_2 becomes conductive as soon as the cathode potential becomes lower than that of its anode; obviously this is the case from point 4, and the potential of the control grid of tube T_4 now follows the curve $V_{Tr_6} - V_{Tr_5}$.

However, ignitron I_1 ceases to conduct when point 3 is reached, and the voltage V_{Tr_6} becomes zero at that instant. The control grid voltage of T_4 then swings from point 5 via point 3 to a positive value (point 6), crossing the zero axis and the igniting characteristic of T_4 (not shown). Consequently the firing impulse for I_2 will start at this instant. Thus the beginning of current flow through I_2 is shifted from point 2 to point 3, and alternate firing of ignitrons I_1 and I_2 is secured.

The electronic timer of this equipment consists of the thyratrons T_7 and T_8 ,

which are of the same type as thyratrons T_3 to T_6 , and the double diode V_3 . If switch S_3 is assumed to be closed, T_8 is obviously ignited since its anode and second grid receive positive voltages via resistors R_{27} and R_{26} , R_{23} . The right-hand diode system of V_3 ensures that point P does not reach a very high positive potential. The potential difference between point P and the positive supply voltage charges capacitor C_7 with the polarity indicated in the figure; similarly C_8 is charged since the anode of T_8 has a potential equal to the arc voltage of the tube.

Across resistor R_{27} a negative voltage of about 90 volts appears which is applied to the second grids of thyratrons T_3 and T_4 , blocking these tubes so that the ignitrons cannot ignite. If now switch S_2 is closed, thyatron T_7 ignites, causing simultaneously extinction of T_8 , since the voltage across C_8 reduces the potential of its anode for a short time below the value of the arc voltage.

Moreover, the voltage of C_7 is applied to the grid with negative polarity, preventing T_8 from firing again until C_7 has discharged through R_{24} and R_{26} . As during this time no voltage appears across R_{27} , thyratrons T_3 , T_4 are able to ignite, making a weld. The duration of the weld time can be adjusted by means of variable resistor R_{26} .

As soon as C_7 has discharged, T_8 ignites again, and since C_8 and C_9 have charged in the meantime with reverse polarity, T_7 is extinguished at that instant. C_9 discharges through R_{25} , R_{27} , producing the off time. Next T_7 ignites again, extinguishing T_8 ; the alternate firing and extinction of the two thyratrons continues until S_2 is opened.

It is necessary that the beginning of weld time and off time should coincide with the instants at which the mains voltage is zero. The timer is therefore synchronised by impulses produced by the peaking transformer Tr_8 . These impulses are applied to the first grids of T_7 , T_8 . In order to obtain the correct phase position of the impulses a resistor R_{28} is inserted in the primary circuit of Tr_8 .

If only single welds are to be made, switch S_3 is opened. If S_2 is closed, T_7 ignites and T_8 becomes non-conducting; after expiration of the weld time T_8 re-ignites. In practice T_7 is extinguished for a short time because of the charge of C_8 , but it will re-ignite immediately and remain in this condition. If, however, S_2 is opened again, T_7 becomes non-conducting, and C_7 and C_8 are charged with the shown polarity so that another weld time can be started by closing S_2 .

Obtaining an Even Number of Half Cycles during Welding

When operating resistance welding machines it is of importance that the weld time includes an even number of half cycles, that is to say that each weld time begins, for instance, with a positive half wave and ends with a negative one. The reason is that the last current half cycle of the preceding weld time causes remanent magnetisation of the welding transformer which may still be present at the beginning of the following weld time if the intermediate off time is short. If the following weld time begins with a current half cycle of the same polarity, an increased transient current occurs which causes sparking at the electrodes and may even blow the mains fuses. It is therefore necessary to design the electronic controlling device so that only an even number of welding half cycles occurs; moreover, care must be taken, of course, that the beginning of each weld time is started at a point corresponding to the phase delay of current

with respect to the mains voltage caused by the power factor of the welding transformer.

The basic circuit of an electronic welding timer for small welding machines which meets the above requirements is shown in fig. 16-12. If the welding current is not larger than some 50 amperes (r.m.s. value) thyatron tubes of sufficiently large size instead of ignitrons can be used, thus considerably simplifying the circuit. In the control grid circuit of tube T_1 the secondary winding of the peaking transformer Tr_4 is inserted, and ensures in-phase ignition. The primary winding of the transformer is connected to the mains supply via the variable

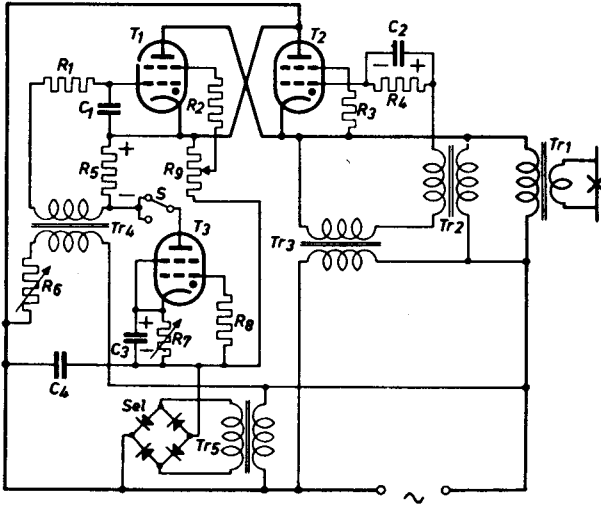


Fig. 16-12. Electronic welding timer producing even numbers of welding half-cycles.

resistor R_6 . This resistor, the value of which can be pre-set, serves for phase adjustment of the secondary impulses with respect to the mains voltage according to the power factor of the welding transformer.

Variable control of the heat applied to the workpiece by phase-shifting the ignition point of the thyatrons in each cycle is not provided in this circuit.

The control grid circuit of tube T_2 contains the protective resistor R_4 , bypassed by capacitor C_2 , and the secondary windings of transformers Tr_2 and Tr_3 . The latter produces an alternating voltage which is phase-shifted by 180 degrees with respect to the anode voltage of the thyatron.

If the influence of the voltage produced by Tr_2 is neglected, a voltage appears at the control grid of T_2 as shown in fig. 16-13b. During the negative half cycle of the anode voltage, capacitor C_2 is charged with the shown polarity via the grid-to-cathode path of T_2 . This voltage is added during the following half cycle to the voltage produced by Tr_3 , C_2 discharging again during this period slowly through R_4 . T_2 thus receives a negative control grid bias preventing the tube from firing.

The primary winding of transformer Tr_2 is connected in parallel to the welding transformer Tr_1 so that in the secondary winding of Tr_2 a voltage can be induced

only if tube T_1 is ignited, and a current half cycle has passed through the main circuit. This is illustrated by means of fig. 16-13*a* and *c*. If T_1 is ignited by a voltage impulse produced by Tr_4 a current half cycle flows, and across the primary winding of Tr_2 a voltage of the shown shape arises (see fig. 16-13*a*). At the secondary a voltage of similar form is present, and is superimposed upon the control grid voltage of T_2 shown in fig. 16-13*b*. As soon as the current flow through Tr_2 has ceased the secondary voltage decays, and a voltage impulse with reverse polarity is produced which effects the ignition of T_2 (see fig. 16-13*c*).

It is obvious that ignition of T_2 is possible only if T_1 was ignited in the preceding half cycle; on the other hand, ignition of T_2 always follows the ignition of T_1 (leading-trailing tube action). Thus it is ensured that the welding current can flow only during one or several *full* cycles. — It should be mentioned that the grid and anode voltage curves shown in fig. 16-13 are drawn to different scales.

A quite simple timing circuit is provided for determining the weld time.

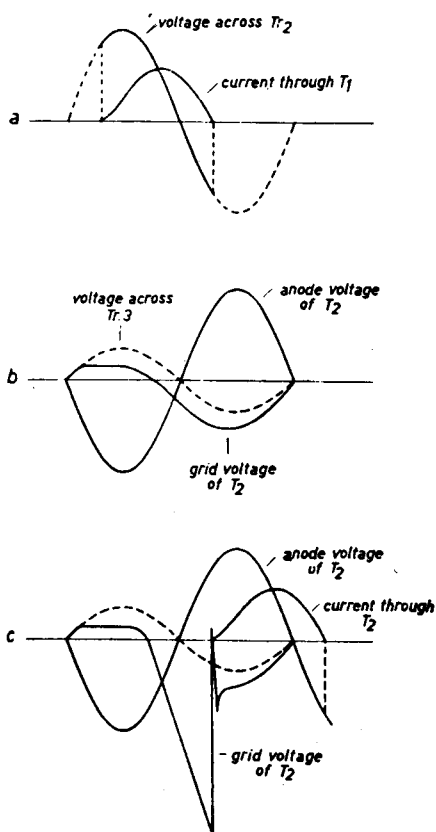


Fig. 16-13. Voltages and currents in the circuit of fig. 16-12.

The alternating voltage produced by transformer Tr_5 is rectified by selenium rectifier $Se1$, the positive pole being connected to the cathode of thyatron T_1 . The shield grid of this tube receives a small negative bias from potentiometer R_9 . When the device is switched on, the small inert-gas filled thyatron T_3 ignites, since no negative voltage is present at its control grid circuit. The anode current then flowing causes a voltage drop across resistor R_7 which charges capacitor C_3 with the polarity as indicated. Moreover, a voltage drop arises across resistor R_5 making the control grid of thyatron T_1 negative so that the positive peaks produced by Tr_4 are not able to ignite this tube.

To start the weld time, switch S is thrown over to its other position. This causes a very short interruption of the anode circuit of T_3 (some milliseconds only), but this will suffice to de-ionise the thyatron so that it will remain extinguished because of the voltage across capacitor C_3 . As a consequence, the voltage across R_5 decays, and no direct voltage is now present at the control grid circuit of T_1 . However, due to the negative bias of the shield grid, ignition of this tube occurs only if the next,

positive peak from Tr_4 appears in the grid circuit. Capacitor C_3 now can discharge slowly through resistor R_7 . By setting R_7 , the discharge time can be adjusted.

The negative grid bias of T_3 is equally reduced, until this tube finally ignites again, completing the weld time. This is shown graphically in fig. 16-14. At point 1 switch S is operated, causing the voltage across R_5 to be momentarily interrupted. The weld time starts, however, at point 2 where the next positive peak from Tr_4 occurs. At point 3 thyatron T_3 ignites again, but the weld time still continues up to point 4 because of the leading-trailing action of T_1 and T_2 .

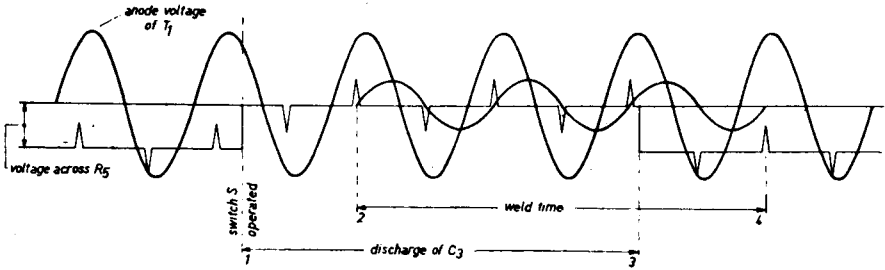


Fig. 16-14. Grid voltage of T_1 in fig. 16-12.

Thus, remanent magnetism of the welding transformer does not introduce any trouble when the next weld time is started.

The device described may be used in connection with a sequence timer producing the other time intervals necessary for operating the welding machine, i.e. squeeze time, hold time and off time. Such timers have already been described in Chapter 12.

Pulsation Welding

Synchronous timing and the avoidance of transient currents caused by remanent magnetism is of special importance for welding machines in which the weld time is broken up into several weld impulses, each consisting of a few cycles, and separated from each other by pauses of nearly the same length (pulsation welding, see fig. 16-15). Subdivision of the weld time in this way is sometimes useful in order to avoid overheating of the workpiece.

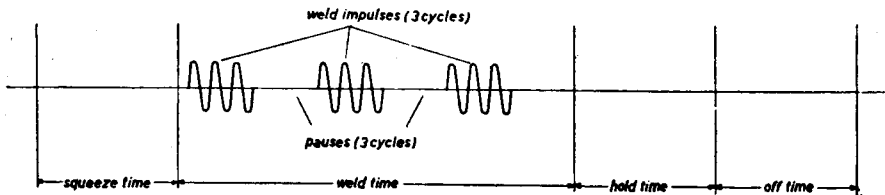


Fig. 16-15: Pulsation weld cycle.

The basic circuit of a device suitable for pulsation welds is illustrated in fig. 16-16. For the sake of simplicity all auxiliary circuits which are not essential for the functioning of the device are omitted in the figure. The main circuit contains, as usual, the welding transformer Tr_1 and two inverse-parallel connected ignitron tubes I_1 and I_2 which are ignited by thyratrons Tb_1 and Tb_2 in the well-known manner. The grid circuit of each of these tubes contains a secondary winding of transformers Tr_7 , Tr_8 and Tr_9 .

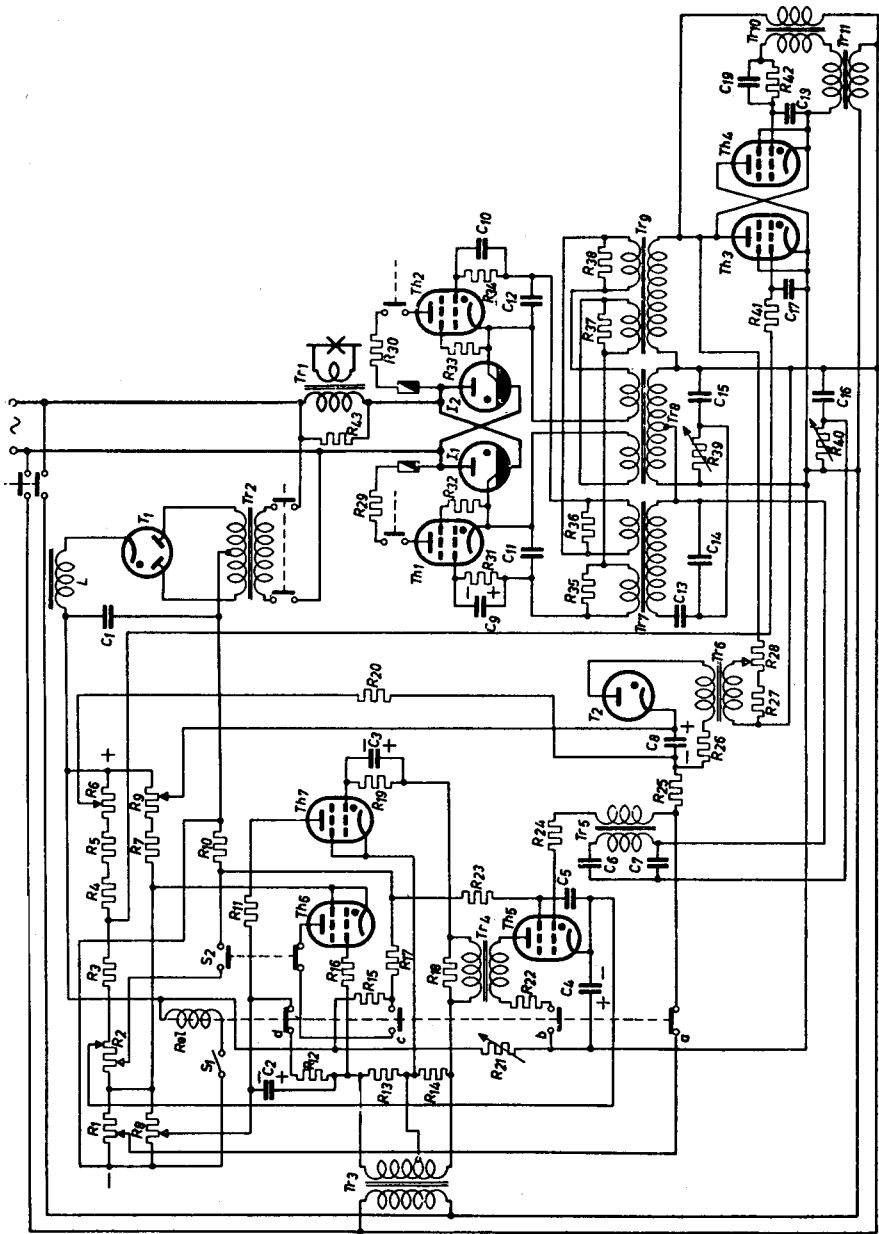


Fig. 16-16. Pulsation welding control circuit.

Since both grid circuits are identical it will suffice to explain the function of that of Th_1 . The voltage produced by Tr_8 is phase-shifted by 180 degrees with respect to the anode voltage so that capacitor C_9 is charged with the polarity as indicated during the negative half cycles of the anode voltage. During the following positive half cycle, therefore, a negative grid voltage arises of such a value that the positive voltage peak produced by peaking transformer Tr_7 does not suffice for igniting Th_1 (see fig. 16-17, curve *A*).

However, if transformer Tr_9 produces a voltage which is in phase with the anode voltage the negative grid voltage of Th_1 is reduced by this amount and the impulse produced by Tr_7 is increased so that ignition of tube Th_1 occurs (fig. 16-17, curve *B*). The same is true for tube Th_2 , so Th_1 and Th_2 ignite as soon as and as long as an alternating current flows through the primary of Tr_9 .

The firing angle of the tubes and thus the heat generated in the workpiece, can be controlled by phase-shifting the impulses produced by Tr_7 ; this is done by a phase network containing C_{15} and variable resistor R_{39} . The alternating current flowing through the primary winding of Tr_9 , the duration of which determines a weld, is controlled by thyratrons Th_3 and Th_4 acting as leading-trailing tubes in the same way as tubes T_1 and T_2 in fig. 16-12.

Th_4 receives an out-of-phase grid voltage via transformer Tr_{11} , preventing the tube from firing. If Th_3 is ignited, a current flows through Tr_{10} , and a voltage impulse in the secondary winding is generated at the end of current flow which in turn effects the succeeding ignition of tube Th_4 . In this way only even numbers of welding half cycles can occur, thus avoiding the undesired influence of remanent magnetism in the welding transformer.

For controlling the welding impulses a special timing circuit is provided, which ignites and extinguishes tube Th_3 at the desired frequency. A rectifier equipped with tube T_1 produces a direct voltage which is applied to the voltage divider R_1 to R_6 . Capacitor C_4 is charged with the shown polarity by a part of this voltage. The cathode of Th_3 thereby receives a positive potential, and this tube is blocked. If now switch S_1 is closed (which may be done, for instance, by a special sequence timer producing squeeze time, hold time and off time), relay *Rel* is energised, so that contacts *b* and *c* are closed, and contacts *a* and *d* are opened. The voltage across capacitor C_4 now appears with positive sign at the anode of thyatron Th_3 ; this tube, however, ignites only on the next positive voltage impulse produced by the peaking transformer Tr_5 , since capacitor C_4 still possesses a charge with the polarity as indicated.

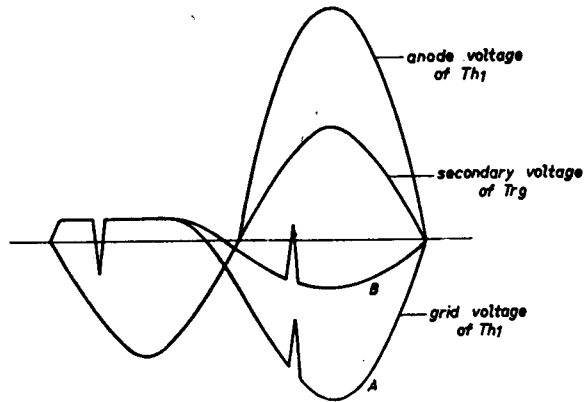


Fig. 16-17. Voltages in the circuit of fig. 16-12.

The phase delay of the impulse is pre-set by means of R_{40} according to the power factor of the welding transformer. After ignition of Th_3 , capacitor C_4 is immediately discharged, and the cathode of Th_3 receives for a short time the potential at the slider of R_2 . As a consequence, Th_3 is ignited, and the first weld impulse begins. C_4 now can charge again through the variable resistor R_{21} .

In order to avoid immediate ignition of Th_3 , transformer Tr_6 receives current by the alternate ignition of Th_3 and Th_4 . The voltage thereby produced in the secondary winding of Tr_6 is rectified by tube T_2 so that capacitor C_8 is charged with the shown polarity, applying a negative bias to the control grid of Th_5 .

If C_4 has re-charged so far that a positive voltage of a certain value is present at the cathode of Th_3 this tube becomes non-conducting, and the first weld impulse ends. Since no further charging of C_8 is now possible, this capacitor discharges through R_{20} , a part of R_6 and a part of R_9 , producing the pause between two weld impulses. This time interval may be controlled by adjusting R_9 . When C_8 has discharged so far that the impulses produced by Tr_5 become effective, Th_5 ignites again, thus introducing the second weld impulse.

A counting circuit is provided in order to limit the number of successive weld impulses to a number which is predetermined by adjustment of R_8 . At the beginning of each weld impulse a current surge flows through the primary winding of Tr_4 because of the discharge of C_4 . The voltage impulse arising at the secondary causes ignition of thyatron Th_7 , which is supplied with alternating anode voltage and an out-of-phase grid voltage by transformer Tr_3 . Obviously Th_7 can be conductive only for a part of a half cycle, and an anode current surge flows during this time interval, charging the capacitor C_2 to a certain voltage. At the beginning of the next weld impulse this process repeats, and C_2 receives another charge which doubles the voltage across it. Further weld impulses cause in this way a gradual increase of the voltage across C_2 until it reaches a value nearly equal to the voltage derived from R_8 . Thus the actual grid voltage of thyatron Th_6 has become nearly zero, and this tube ignites. The second grid of Th_5 has up to now received a potential by the voltage divider R_{10} , R_{17} and R_{15} slightly less than the cathode potential of this tube. On ignition of Th_6 , however, the potential of the second grid of Th_5 is lowered so much that Th_5 cannot re-ignite after expiration of the preceding weld impulse and the pause, thus finishing the weld time. Th_6 is extinguished, releasing Th_5 only when S_1 is opened and the relay has dropped out.

By operation of switch S_2 the anode circuit of Th_6 is interrupted, and the second grid of Th_5 receives a fixed potential which is slightly negative and is derived from a slider of R_2 . The counting circuit thus becomes ineffective, and the device may be used now for seam welding.

Fig. 16-18 shows two views of an electronic control equipment for welding machines, which is suitable for seam welding and can be fitted with an additional counting unit, as shown at the top of the left-hand view, for pulsation welding.

Three-Phase Welding Equipment

The devices hitherto described are in general suitable for single phase mains supply. However, this causes non-symmetrical loading of the three phase power line which may lead to trouble. For this reason, devices for handling larger powers are now designed so that three-phase mains supply may be used, i.e.

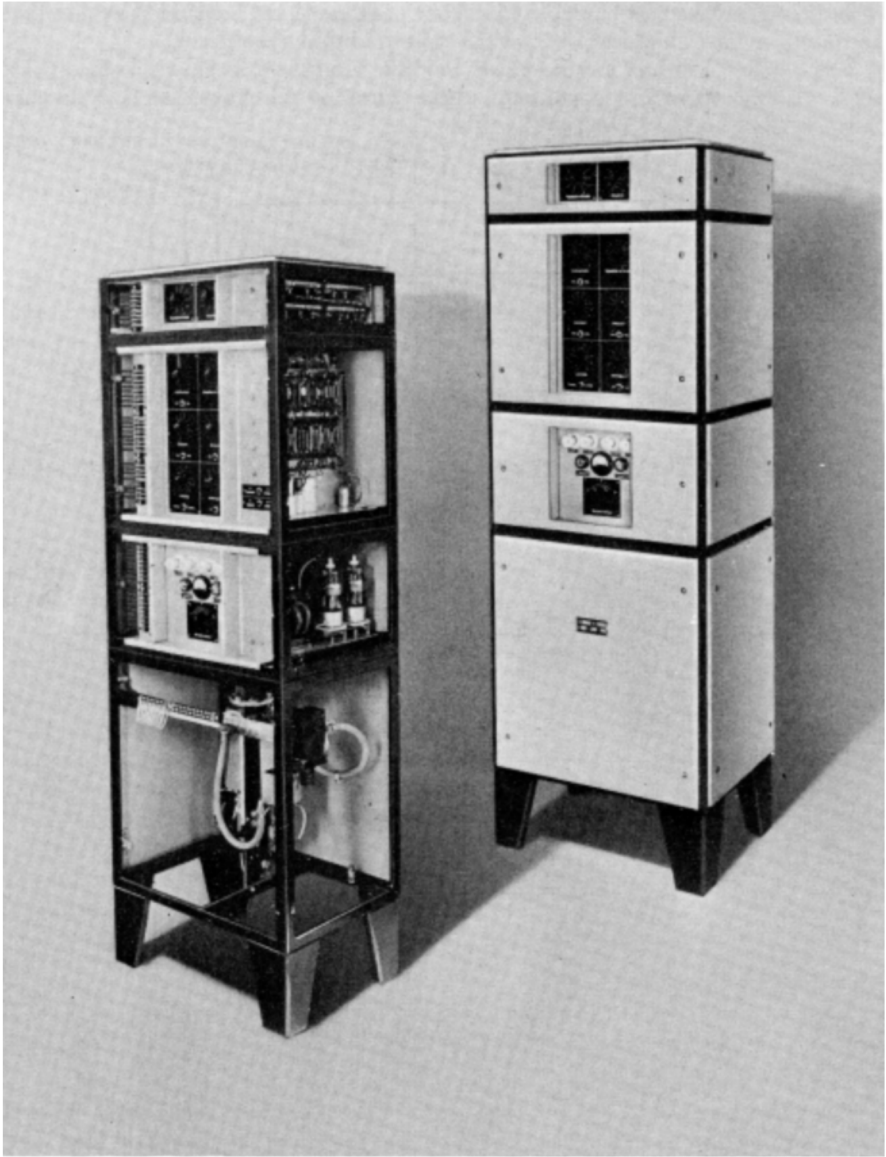


Fig. 16-18. Electronic control equipment for seam and pulsation welds.

the single-phase load is shared symmetrically on all three phases of the power line. Such devices operate on the well-known principle of a frequency changer. The three-phase current having a frequency of 50 or 60 c/s is changed by alternate rectification into single-phase current with reduced frequency.

The block diagram of a three-phase welding equipment is shown in fig. 16-19. The welding transformer contains three primary windings, each connected

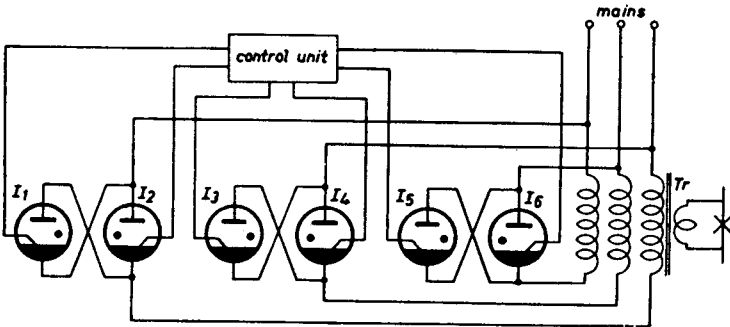


Fig. 16-19. Block diagram of a three-phase welding control.

to the three-phase mains via two inverse-parallel connected ignitron tubes. Thus six ignitrons are required in all. In fig. 16-20 the curves of output voltage (a) and output current (b) are plotted, the frequency being $13\frac{7}{11}$ c/s.

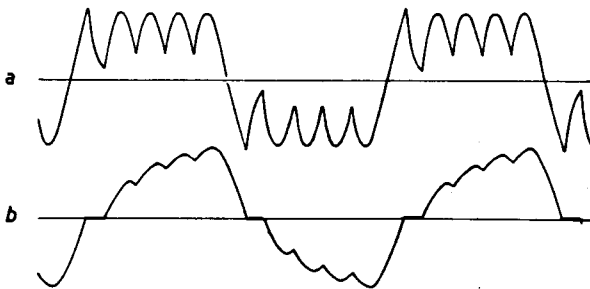


Fig. 16-20. Output voltage and current wave forms of a three-phase welding control system.

The advantages resulting from balanced three-phase welding may be well illustrated by the following table which shows the comparison of load conditions of two welding machines for equal useful power, designed for single-phase and three-phase supply:

	Single-phase supply 50 c/s	Three-phase supply 13 c/s
Useful power	100 kW	100 kW
Apparent power of the machine	250 kVA	118 kVA
Power factor	0.4	0.85
Load currents of the three phases	$\left. \begin{array}{l} R = S = 658 \text{ A} \\ T = 0 \end{array} \right\}$	$R = S = T = 200 \text{ A}$
Load of power line	$\left. \begin{array}{l} 250 \text{ kVA} \\ \text{not symmetrical} \end{array} \right\}$	132 kVA balanced

However, it should be borne in mind that the iron requirement of the welding transformer, due to the reduced frequency, is considerably higher than in the case of single-phase welding machines. As an example of the conventional



Fig. 16-21. Three-phase welding equipment.

construction, a photograph of a three-phase welding control equipment is reproduced in fig. 16-21.

Fig. 16-22 gives a survey of the various combinations of metals and alloys that can be spot-welded.

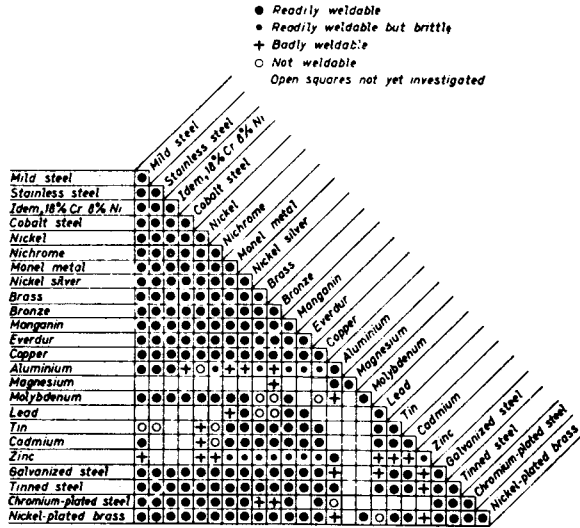


Fig. 16-22. Survey of the usability of the spot-welding process for various metals and alloys.

17. ELECTRONIC MOTOR CONTROL

A very important section of industrial electronics is electronic motor control, which often provides the solution for complex drive problems. Electronic control possesses so many advantages over other methods that its application is certain to be greatly extended in the near future.

The block diagram of an electronic motor control system is shown in fig. 17-1. It consists of a power transformer for connection to the single-phase or three-phase a.c. power line, and two rectifier units equipped with thyratrons for generating the variable-voltage d.c. supplies for the armature and for the field excitation of a d.c. shunt-wound motor. Sometimes the power transformer is omitted in order to save cost.

More or less automatic control of the rectifier output voltages is necessary in order to achieve the desired drive characteristics, and is exercised by what may be termed an "electronic brain" unit. In order to examine the possibilities of this system, the following simplified equations, which are valid for d.c. shunt-wound motors, will suffice. The motor speed n is proportional to the electromotive force E induced in the armature, and inversely proportional to the magnetic flux Φ of the field:

$$n = c_1 \cdot \frac{E}{\Phi}, \tag{17.1}$$

when c_1 is a constant depending upon the design of the motor and the chosen scale for E and Φ .

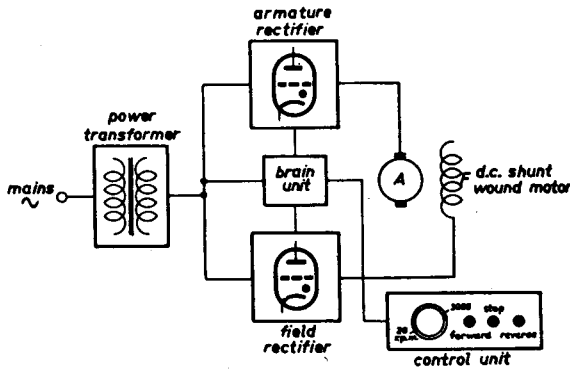


Fig. 17-1. Block diagram of an electronic motor control system.

Obviously the armature voltage V must be the sum of E and the voltage drop caused by the armature current I across the armature resistance R , so that

$$n = c_1 \cdot \frac{V - IR}{\Phi}. \quad (17.2)$$

The torque is:

$$T = c_2 I \Phi, \quad (17.3)$$

and the power output of the motor is:

$$P = nT = c_3 (VI - I^2 R). \quad (17.4)$$

Finally,

$$\Phi = c_4 i, \quad (17.5)$$

where i is the field excitation current.

It will be seen therefore, that control can be achieved in the following ways:

- (1) By reducing the field excitation the speed can be increased, the torque being correspondingly reduced but the power remaining constant. The maximum control range is usually about 1 : 3, but recent developments in motor design have increased this range up to 1 : 5.
- (2) By reducing the armature voltage the speed can be reduced from the rated value to a very low value, the power decreasing proportionally with the speed but the torque remaining constant.
- (3) By increasing the armature voltage the speed can be increased, the torque remaining constant. A total increase of speed of about 40% is possible.

The possibilities (1) and (2) are of particular practical importance. They are plotted graphically in fig. 17-2. Over the region of armature voltage control the output power is proportional to the instantaneous speed, up to the rated speed. Over the field control range the power remains constant, the torque

being reduced according to a hyperbolic curve described by equation (17.4). Depending upon the desired driving characteristics, it is thus possible to operate with variable armature voltage and constant field, or with constant armature voltage and variable field, or by variation of both armature and field. In this manner a motor control system can be designed for any desired purpose, and a number of additional features can be incorporated.

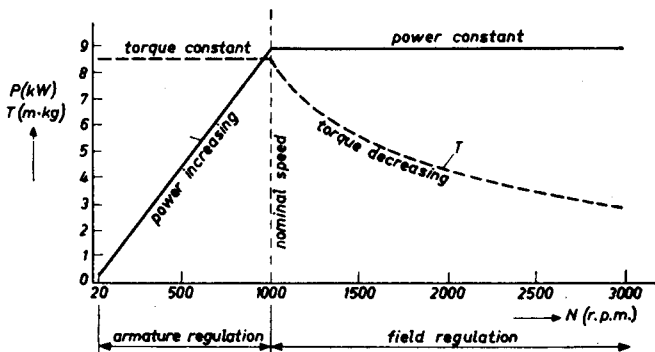


Fig. 17-2. Graph showing power and torque against speed for armature and field regulation.

In addition to speed control as already described, the following controls may be achieved:

- (1) The motor may be started with optimum torque and run up to a pre-determined speed in the clockwise or anti-clockwise direction.
- (2) A pre-determined speed may be held constant within narrow limits, independent of variations of load.
- (3) The speed may be changed according to a pre-set programme.
- (4) The motor can be rapidly brought to rest.
- (5) The motor can be rapidly reversed by push-button control.

Of particular importance is the fact that all the above-mentioned features may either be controlled automatically or operated by small, simple potentiometers, since all the controlling processes employ only small currents.

According to equation (17.2) the speed is directly proportional to the armature voltage (neglecting for the moment the voltage drop across the resistance of the armature). Thus the maximum range of speed control can be easily calculated if the rated value of the armature voltage is known and if it is kept in mind that a minimum armature voltage of about 9 volts is necessary to cover the potential differences across the carbon brushes and to ensure quiet running. Thus, for a 440-volt motor, a maximum speed range of 1 : 50 can be obtained; for a 220-volt motor the range is obviously only 1 : 25. The instantaneous speed can be calculated from the rated speed and the instantaneous output voltage of the armature rectifier. For example, assuming a three-phase rectifier with fully-conducting thyratrons, the output voltage will be:

$$V_o = 1.17 V_{tr}, \quad (17.6)$$

where V_{tr} is the r.m.s. value of the transformer voltage.

Assuming a firing angle of 30 degrees, the average output voltage is, according to equation (3.2):

$$V_o' = \frac{3 \cdot \sqrt{2}}{2\pi} \int_{60^\circ}^{180^\circ} \sin \varphi \, d\varphi \quad V_{tr} = 1.02 V_{tr} \quad (17.7)$$

If the rated speed is n_0 , the speed at a firing angle of 30 degrees will be:

$$n(\varphi = 30^\circ) = \frac{1.02}{1.17} \cdot n_0 \quad (17.8)$$

In the same manner the speed can also be calculated for larger firing angles, but it should be remembered that because of the inductive load presented by the armature, the time the thyristors are conducting is extended (see fig. 3-19), the transformer voltage swinging partly negative during each cycle. The actual

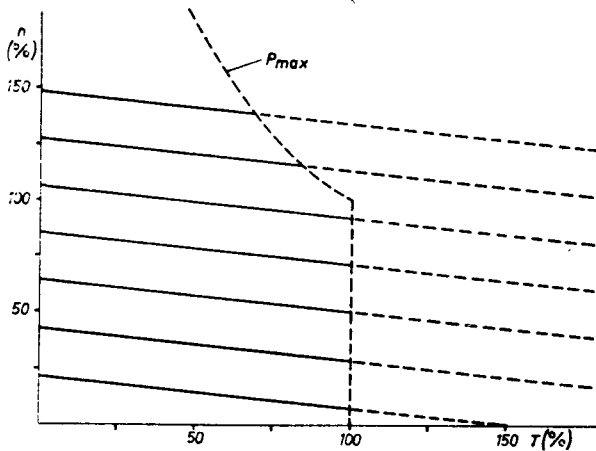


Fig. 17-3. Speed characteristics of a d.c. shunt wound motor.

output voltage is thus reduced. At very large firing angles, and if an electromotive force of suitable polarity is present in the circuit, current will flow almost entirely in the region of negative transformer voltage, thus feeding energy back into the mains. Consequently a very effective braking action is applied to the motor.

In most cases the voltage drop caused by the armature resistance cannot be neglected. The result of this voltage drop is that the speed, once adjusted by applying a certain armature voltage, is reduced when the motor load is increased. The speed/torque characteristic of a d.c. shunt-wound motor therefore shows a dropping trend as indicated in fig. 17-3. It is therefore necessary to take certain precautions, when starting the motor, to limit the armature current, as this would be excessive if the voltage corresponding to the desired speed at full load were applied immediately on starting. It is thus clear that two conditions must be fulfilled in order to obtain a satisfactory drive characteristic, namely:

- 1) The speed should remain independent of the load. This can be achieved by compensating for the voltage drop across the armature resistance in order to obtain a horizontal speed-torque characteristic.
- 2) The armature current should be automatically limited to an adjustable maximum value.

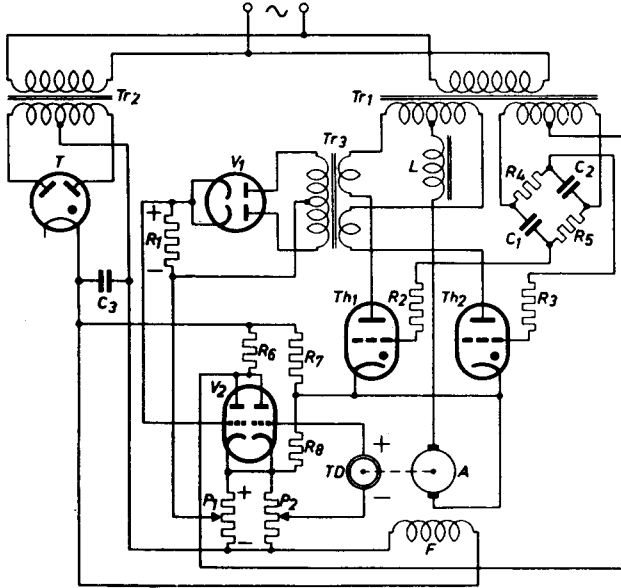


Fig. 17-4. Simplified full-wave motor-control system.

Both these demands can be met in a comparatively simple manner by means of electronic motor control, which not only operates instantaneously but utilises only small currents in the control circuits.

Fig. 17-4 is a greatly simplified basic circuit for a two-phase equipment designed for armature voltage control and containing also a circuit for automatic speed stabilisation. The armature rectifier is equipped with two thyratrons, Th_1 and Th_2 , and the field excitation current is taken from a rectifier using the gas-filled tube T . A smoothing reactor L is connected in series with the armature A . The grids of Th_1 and Th_2 receive alternating voltages which lag by 90 degrees and are superimposed on a variable direct voltage (vertical control). This direct voltage is derived from a bridge circuit comprising resistors R_7 , R_8 , R_6 and the internal resistance of the double triode V_2 . Obviously the bridge output voltage is positive if both triode sections of V_2 are blocked, and becomes negative if at least one section of V_2 is conducting. The right-hand section, which serves for speed stabilisation, receives a negative grid bias taken from potentiometer P_2 . This voltage is opposed by a direct voltage, proportional to the motor speed and delivered by the tachodynamo TD . If the speed is, for example, too high, the tachodynamo voltage exceeds the negative grid bias, so that the right-hand section of V_2 conducts. The firing angle of thyratrons Th_1 and Th_2 is thereby

increased, and the motor slows down until the desired speed is reached. If, on the other hand, the motor speed is too low, for example as the result of an increased load, the negative voltage taken from P_2 exceeds the positive voltage delivered by the tacho-dynamo, and the right-hand triode section is blocked. If the left-hand triode section also becomes blocked the ignition point of the thyratrons is advanced, allowing a larger current to flow through the armature corresponding to the heavier load, and the motor speeds up.

The left-hand section of the double triode V_2 serves for limiting the armature current to a maximum value which can be pre-set by means of the potentiometer P_1 . The voltage taken from P_1 is applied as negative bias to the grid of the left-hand triode section, blocking it for the moment. In the anode loads of thyratrons Tb_1 and Tb_2 are inserted the primary windings of a current transformer Tr_3 . In the secondary winding of this transformer, therefore, a voltage is induced which increases if the armature current increases. This voltage is rectified by tube V_1 and opposes the voltage taken from P_1 . As soon as the voltage supplied by Tr exceeds a certain value the left-hand triode section of V_2 begins to pass current, thus increasing the firing angle of the thyratrons. In this manner the armature current or the torque is prevented from exceeding a certain value predetermined by the setting of P_1 .

The great advantage of the current-limiting circuit is that the motor can be run up to speed without auxiliary starting devices — for example by push-button control. The desired speed can be pre-set, and the motor starts up with a torque corresponding to the pre-set limiting value of the armature current. Very smooth starting is thus achieved, which is of great advantage in many applications, for example in wire-drawing plants, spinning mills and cable factories etc. Moreover the automatic current-limiting feature completely safeguards the motor in the event of overload. By suitable design of the current limiting circuit it is possible to brake the motor to a standstill without the current increasing by more than a few percent of the pre-set limiting value.

Speed stabilisation by means of a tacho-dynamo has the advantage that the speed is very constant — the speed variation between no-load to full-load being less than one per cent. In many instances, however, such as machine-tool drives, such accurate speed stabilisation is not necessary, and a circuit can be used in which the armature voltage can be employed as a measure of the speed. The influence of the voltage drop across the armature resistance must then be eliminated by a special IR -compensating circuit. The drop in speed which would otherwise occur may be of no great importance at high speeds, but it cannot be neglected at low speeds when the armature voltage is comparable to the IR drop. Circuits employing speed stabilisation by armature voltage and IR compensation hold the speed constant to within a few per cent. Such a circuit will be fully described later.

Simple Single-Phase Motor Control Systems

In some applications, for example feed drives and servo mechanisms, constant torque is required over a range of speed adjustment which may be as great as 1 : 25. Since the output powers involved are but small, simple devices can be used which deliver constant field excitation and a variable armature voltage from a half-wave single-phase rectifier.

Such a circuit is shown in fig. 17-5, the motor being designed to operate at 220 volts and a power consumption of up to 80 watts. The field current is supplied by a selenium rectifier *Sel.* In order to reduce cost, a power transformer is not employed in this circuit. In the armature circuit is connected a PL 5557 thyatron which provides armature voltage regulation by grid control. The armature is, in fact, fed only by voltage pulses, and the current flowing through the armature is therefore not continuous but consists of pulses

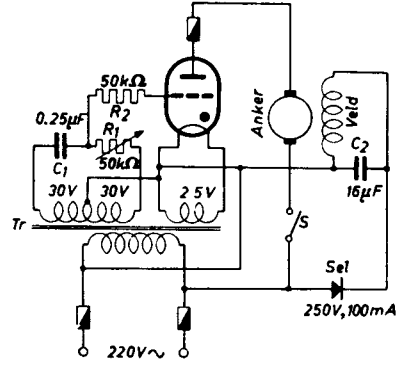
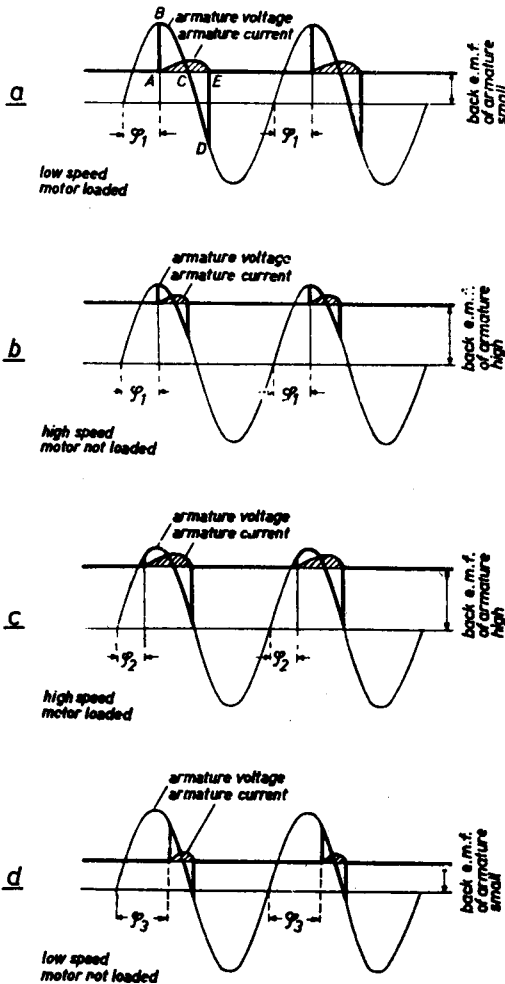


Fig. 17-5. Simple single-phase motor control circuit.



separated by intervals, the duration of which is dependent upon the firing angle of the thyatron and upon the back e.m.f. produced in the motor.

This is shown graphically in fig. 17-6, the arc voltage drop of the thyatron being neglected for the sake of simplicity. At (a) the firing angle φ_1 is about 90 degrees, the back e.m.f. being only small. Comparatively large current pulses then flow in the armature, i.e. the motor delivers power at low speed. The difference between the areas *ABC* and *CDE* is proportional to the voltage drop across the armature resistance caused by the current flowing in the armature. At (b) the firing angle is the same (φ_1) but a high back e.m.f. is assumed. The armature current pulses are small and may be sufficient only to provide for the internal losses of the motor when running at high speed. This corresponds to running under no-load conditions. If now the motor load increases, the speed will decrease and the armature current

Fig. 17-6. Wave forms of armature voltage and current at different speeds and load conditions.

will increase until conditions corresponding to (a) are reached. If, however, this drop in speed is undesirable, the increase of armature current caused by the load can be covered by advancing the ignition point of the thyatron (see φ_2 at (c)). The speed then remains constant. Finally, at (d) is shown how the firing angle must be increased, say to φ_3 , at low speed if the motor load assumed in (a) is removed and the speed is to remain constant.

Variation of the firing angle is achieved in fig. 17-5 by an alternating voltage of about $30 V_{rms}$, the phase of which can be varied with respect to that of the anode voltage of the thyatron (horizontal control). For this purpose a phase network comprising C_1 and R_1 is provided, the phase shift being adjusted by varying R_1 possibly by remote control. In order to minimise the total cost of the installation no armature current limiting circuit is included, the armature and thyatron being protected against overload by a fuse. Fig. 17-7 is an experimental control unit constructed on the lines of fig. 17-5.

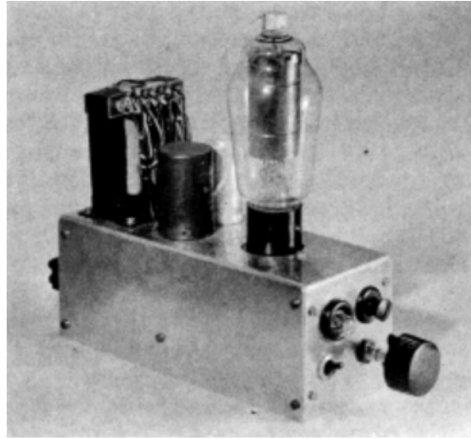


Fig. 17-7. Experimental equipment using the circuit of fig. 17-5.

A somewhat less simple device, incorporating speed stabilisation independent of load and a rapid braking circuit is illustrated in fig. 17-8. The field excitation

is again held constant and is supplied by a rectifier circuit employing the full-wave rectifier tube type 1701. Due to the field current a voltage drop of about 30 volts occurs across resistor R_5 so that a voltage can be taken from the slider of potentiometer R_3 and can be adjusted between -30 and $+220$ volts with respect to point b . This voltage is applied to the grid circuit of thyatron Th which controls the variable armature voltage. In addition, an alternating voltage of approximately $20 V_{rms}$, lagging by 90° , is produced by the phase network consisting of C_1 and R_1 , and is also applied to the grid of the thy-

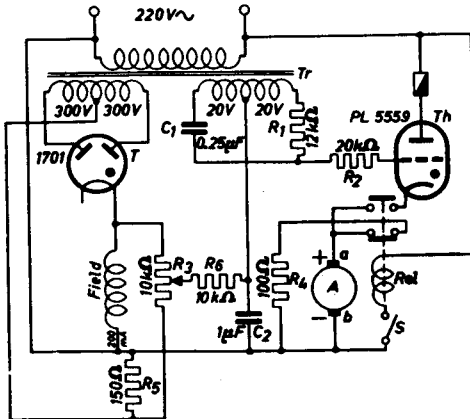


Fig. 17-8. Single-phase motor control with speed stabilising circuit and dynamic braking.

atron. The speed may be pre-set by adjustment of R_3 . If the voltage taken from R_3 is, for instance, 100 volts, the grid of Th becomes positive by this

amount as soon as switch S is closed and relay Rel is energised.

Thus the thyatron is fully conducting during alternate half cycles, and the motor starts to run. A back e.m.f. is therefore generated in the armature A and increases as the motor gathers speed, until finally a value of almost 100 volts is reached. Since this voltage is in opposition to the voltage taken from R_3 , any further increase of speed will cause the ignition point of Th to lag, and this in turn reduces the armature current until finally a speed is maintained which corresponds to the setting of R_3 , which may be calibrated accordingly.

If now the motor load increases, the speed and consequently the back e.m.f. will decrease by a small amount. The ignition point of the thyatron is then automatically advanced, and the armature current increases to correspond to the increased load, and further drop in speed is prevented. In this way speed stabilisation independent of load can be achieved to within a few percent, which will be quite sufficient in many applications.

As soon as switch S is opened, the relay drops out, and the armature is disconnected from the supply and connected across a braking resistor R_4 , where the energy generated by the motor — which now operates as a generator — is dissipated as heat, and the machine rapidly comes to rest.

This circuit can be used to control motors taking up to 400 watts if a PL 5559 thyatron is used. If a PL 105 thyatron is employed a motor taking up to 1.2 kW can be controlled.

Three-Phase Motor Control with Armature and Field Control

The complete circuit of a three-phase motor control system with combined armature and field control for a maximum output of 9 kW and incorporating speed stabilisation independent of load and IR compensation, automatic armature current limitation, quick braking and quick reverse, is given in fig. 17-9.

In order to obtain a wide range of speed control the device is designed for a 440-volt motor, so that in this circuit a power transformer is provided to step the mains voltage up to the required value. The secondary winding of this transformer is connected in zig-zag to eliminate premagnetisation by the unidirectional anode currents.

Three PL 105 thyatrons, Th_1 , Th_2 , Th_3 are provided for armature voltage rectification, so that a maximum average armature current of about 19 amperes can be supplied. Variation of the firing angle of these tubes is achieved by vertical control; the control grids receive alternating voltages of approximately 30 volts, phase lagged by 90 degrees with respect to the anode voltages, and taken from the secondary of transformer Tr_3 , the primary of which is not shown for the sake of simplicity.

The variable direct voltage required for vertical control appears across capacitor C_2 and is produced by rectification in both tubes V_1 and V_2 . If both sections of double triode V_2 are assumed to be blocked, then C_2 is charged by V_1 up to approximately 100 volts with the polarity indicated in fig. 17-9, and thyatrons Th_1 , Th_2 and Th_3 are fully conducting. However, if one or both sections of V_2 pass current, the charge of C_2 is reduced and may even change its polarity so that the firing angle of the thyatrons is increased correspondingly. Finally the thyatrons are completely blocked if one section of V_2 passes full current.

The left-hand section of V_2 , which serves for speed stabilisation independent of load, receives a grid voltage equal to the difference between an adjustable

reference voltage taken from R_{13} and stabilised by T_1 , T_2 , and half the armature voltage. As soon as the motor speed, and thus the armature voltage, slightly exceeds the value corresponding to the setting of R_{13} , the left-hand triode section of V_2 receives a positive grid bias, and consequently the firing angle of the thyratrons is increased until the speed has fallen to the predetermined value. If, however, the speed is lower than the pre-set value, due, for instance, to increased load, the voltage at the slider of R_{13} predominates, and the left-hand triode becomes less conductive, and the firing angle of the thyratrons is reduced. An increased armature current therefore flows to meet the increased load demand. It is true that the voltage drop across the armature resistance, which represents the difference between the armature voltage and the back e.m.f. of the armature, also increases.

Since the speed is directly proportional to the back e.m.f. and not to the armature voltage, the influence of the IR drop must be compensated by suitable measures. This is achieved by connecting the grid of the left-hand triode not directly to the common cathode lead of the thyratrons but to the slider of resistor R_{28} so that the left-hand triode receives an additional negative bias. The current through R_{28} is proportional to the armature current, as explained later, so that the left-hand triode becomes less conductive as the armature current increases. As a result the armature voltage is increased, thus compensating the speed drop which would otherwise occur at increased motor load. This compensation circuit is very effective, the speed variation from no-load to full-load conditions when the slider of R_{28} is set to the optimum position being less than one or two percent.

For automatic limitation of the armature current, special current transformers Tr_5 , Tr_6 and Tr_7 are included in the anode leads of the thyratrons, and the secondary voltages are rectified by small selenium rectifiers Sel_1 , Sel_2 , Sel_3 so that a rectified voltage of approximately 300 volts is produced at full-load armature current (approximately 19 amperes). A part of this voltage is applied via a voltage divider R_{26} , R_{27} to a bridge circuit consisting of resistors R_{28} , R_{29} , R_{32} and $R_{30} + R_{31}$. A voltage reference tube T_3 is connected across R_{30} , R_{31} . From R_{31} is taken the grid bias for the right-hand section of the double triode V_2 , the slider being adjusted so that the grid is about 30 to 40 volts more negative than the cathode as long as T_3 is not ignited. The right-hand triode is therefore blocked. The action of this part of the apparatus can be followed by first assuming that the slider of potentiometer R_{27} , which serves for adjustment of the armature current limitation, is turned to its extreme left-hand position.

If now the rectified voltage derived from the current transformers increases and exceeds a value of 300 volts, the voltage applied to the bridge circuit is somewhat greater than 150 volts, and the voltage across resistors R_{30} and R_{31} reaches the ignition voltage value of T_3 . This tube ignites, and the voltage across it falls to 85 volts so that the grid voltage of the right-hand section of V_2 becomes nearly zero. This triode section therefore passes current, with the result that the firing angle of the thyratrons is increased so that the armature current cannot increase further. If the slider of R_{27} is set at some other position, current limitation occurs at a lower value of armature current. For example, if the slider is moved to the extreme right, the armature current will be limited to approximately 10 amperes.

As in the circuit of fig. 17-8, rapid braking of the motor is achieved by dis-

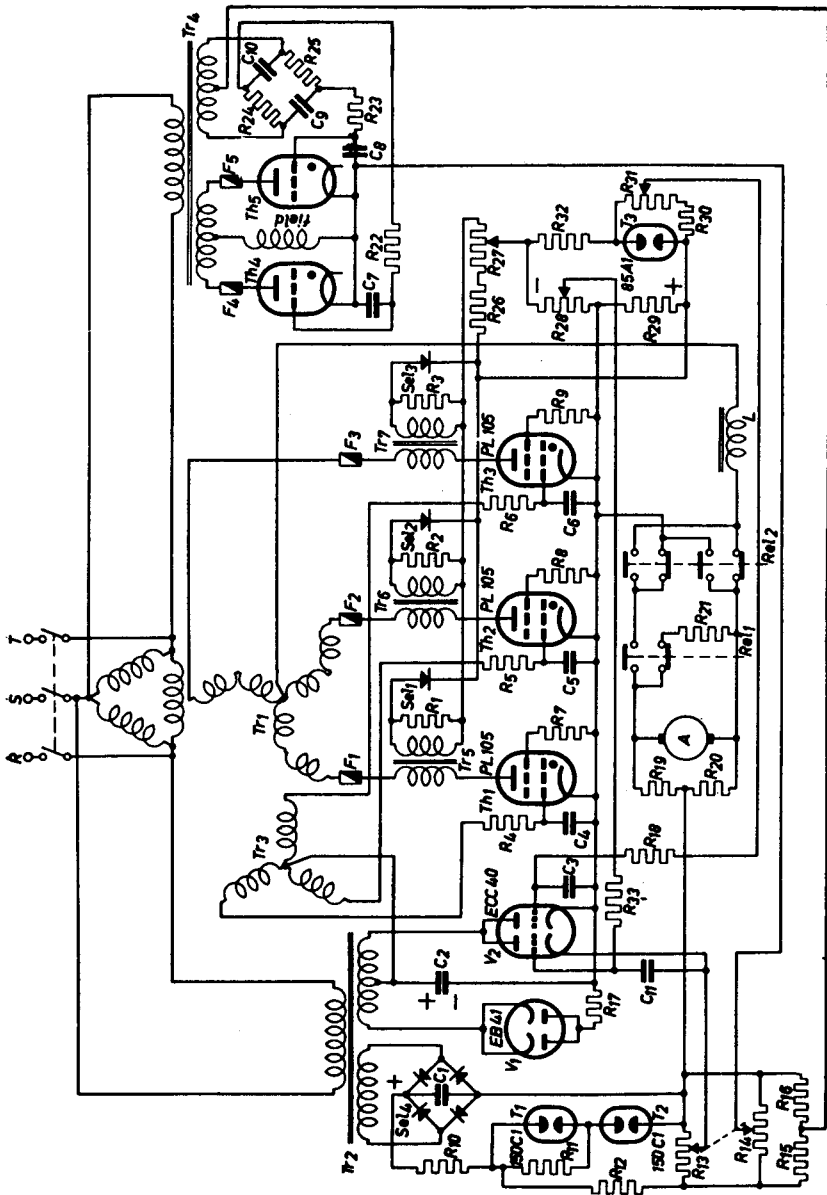


Fig. 17-9. Three-phase motor control circuit.

connecting the armature from the rectifier and connecting it in parallel with a braking resistor R_{21} . This is done by relay Rel_1 , the winding of which is omitted from the diagram for the sake of simplicity. To reverse the motor the armature connections are reversed by relay Rel_2 .

The field excitation of the motor is taken from a separate rectifier employing PL 5557 thyatrons Tb_4 and Tb_5 . The firing angle of these tubes is varied by vertical control. For this purpose RC phase-shifting networks R_{24} , C_{10} and R_{25} , C_9 are provided and the variable direct voltage is taken from a bridge network comprising potentiometer R_{14} and resistors R_{15} and R_{16} . Resistor R_{14} serves for controlling the field excitation. If desired it can be mechanically coupled to potentiometer R_{13} which serves for armature voltage control. With this arrangement a speed control range of approximately 1 : 150 is obtainable. By independent or simultaneous operation of the two potentiometers any special drive requirement can be met.

Instead of a single pair of potentiometers a number of control elements can be provided, each set for a different motor speed and direction of rotation. By means of a selector switch or push-buttons these control elements can be switched on successively so that a pre-determined programme is followed. The selector switch can be operated, for instance, by the carriage movements of an automatic machine tool.

Constant-Torque Drive

As a result of speed stabilisation with IR compensation and armature current limitation, a speed characteristic of the form shown in fig. 17-10 is obtained. The rate of decrease of speed at the upper limit value of torque can be made so high that the torque increases only little when the motor speed is reduced from the rated speed down to a standstill, i.e. the torque remains substantially constant. It is thus seen that a constant-torque drive can be obtained by adjusting the armature current limitation circuit for the desired torque (fig. 17-11).

Torque stabilisation is especially important for multi-motor drives such as those extensively used in rolling mills, wire drawing, and the textile industry etc. As is customary in this type of drive, a main motor determines the travel speed of the material, and it is essential that the other motors adapt themselves to the speed of the main motor in order to prevent undesired stretching or other

COMPONENT VALUES OF THE CIRCUIT OF FIG. 17-9

$R_1 = 10 \text{ k}\Omega$	$R_{15} = 15 \text{ k}\Omega$	$R_{29} = 20 \text{ k}\Omega$
$R_2 = 10 \text{ k}\Omega$	$R_{16} = 15 \text{ k}\Omega$	$R_{30} = 10 \text{ k}\Omega$
$R_3 = 10 \text{ k}\Omega$	$R_{17} = 20 \text{ k}\Omega$	$R_{31} = 50 \text{ k}\Omega$
$R_4 = 20 \text{ k}\Omega$	$R_{18} = 0.5 \text{ M}\Omega$	$R_{32} = 15 \text{ k}\Omega$
$R_5 = 20 \text{ k}\Omega$	$R_{19} = 10 \text{ k}\Omega$	$C_1 = 8 \text{ }\mu\text{F}$
$R_6 = 20 \text{ k}\Omega$	$R_{20} = 10 \text{ k}\Omega$	$C_2 = 2 \text{ }\mu\text{F}$
$R_7 = 5 \text{ k}\Omega$	$R_{21} = 20 \text{ }\Omega$	$C_3 = 0.1 \text{ }\mu\text{F}$
$R_8 = 5 \text{ k}\Omega$	$R_{22} = 20 \text{ k}\Omega$	$C_4 = 500 \text{ pF}$
$R_9 = 5 \text{ k}\Omega$	$R_{23} = 20 \text{ k}\Omega$	$C_5 = 500 \text{ pF}$
$R_{10} = 2 \text{ k}\Omega$	$R_{24} = 12 \text{ k}\Omega$	$C_6 = 500 \text{ pF}$
$R_{11} = 0.1 \text{ M}\Omega$	$R_{25} = 12 \text{ k}\Omega$	$C_7 = 500 \text{ pF}$
$R_{12} = 2.5 \text{ k}\Omega$	$R_{26} = 3 \text{ k}\Omega$	$C_8 = 500 \text{ pF}$
$R_{13} = 15 \text{ k}\Omega$	$R_{27} = 3 \text{ k}\Omega$	$C_9 = 0.25 \text{ }\mu\text{F}$
$R_{14} = 30 \text{ k}\Omega$	$R_{28} = 15 \text{ k}\Omega$	$C_{10} = 0.25 \text{ }\mu\text{F}$

straining of the material. This can be achieved in an elegant manner by an electronic motor control system, thus avoiding the elaborate mechanical linkage which would otherwise be necessary. The auxiliary motors are adjusted

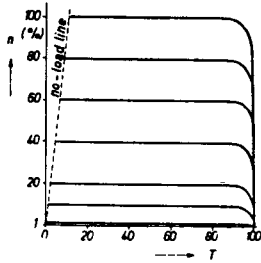


Fig. 17-10. Speed characteristics of a d.c. shunt wound motor with electronic speed stabilisation and armature current limitation.

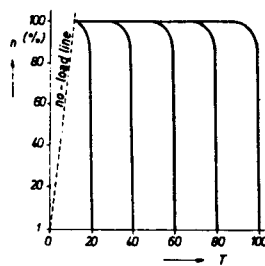


Fig. 17-11. Speed characteristics of a constant-torque drive.

for a torque corresponding to the permissible tension of the material, and this torque is then held constant. The speed will then automatically adjust itself to the required value. Thus the tension exerted on the material remains constant at all times independent of the operating speed, down to complete standstill.

Constant-Power Drive

A very common problem is concerned with spooling thread or wire, or rolling paper, cloth etc. Since the diameter of the roll increases steadily, the winding speed must decrease accordingly if the travel speed of the material remains constant. As the stress t on the material should remain constant, the torque T must increase proportionally to the radius r of the roll, since:

$$T = t \cdot r. \tag{17.9}$$

The power, however, which is the product of torque and speed, remains constant. In other words, the winding machine must be controlled in such a manner that the speed-torque characteristic is a hyperbola, i.e. a curve of constant power, as in fig. 17-12. This can easily be achieved by means of a pressure roll which, in effect, continuously measures the diameter of the wound roll or coil, the progressive displacement of the pressure roll with increasing spool diameter operating the potentiometer which adjusts the torque.

However, the pressure roll and its mechanical linkage can be dispensed with by regulating the drive for constant power. To achieve this, the output voltage of the armature rectifier is adjusted to a value corresponding to the desired mean

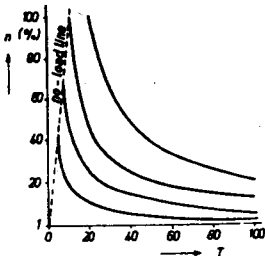


Fig. 17-12. Speed characteristics of a constant-power drive.

winding speed, i.e. the travel speed of the material, and a separate rectifier for the field excitation is controlled in such a way that the armature current, the nominal value of which is predetermined by the power demand, is held constant. This arrangement results in the field current increasing in proportion to the diameter of the winding roll, causing a corresponding increase of the torque according to equations (17.2) and (17.3), while the winding speed decreases in inverse proportion.

Since field control permits a speed variation range of the order of 1 : 3 to 1 : 5 only, this limitation applies equally to the ratio of the diameters of the empty and full winding roll, but in practice a larger ratio is seldom required.

Reversal and Braking

Reversal of the direction of rotation of a motor is easily achieved by reversing the armature supply by means of a contactor. Braking is accomplished in the simplest case by disconnecting the armature from the rectifier and connecting it in parallel with a resistor in which the kinetic energy of the armature is converted to heat (dynamic braking). A much more effective method of reversal and braking is as follows.

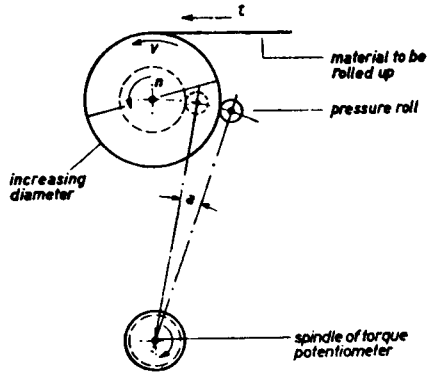


Fig. 17-13. Mechanical control of winding or spooling drives.

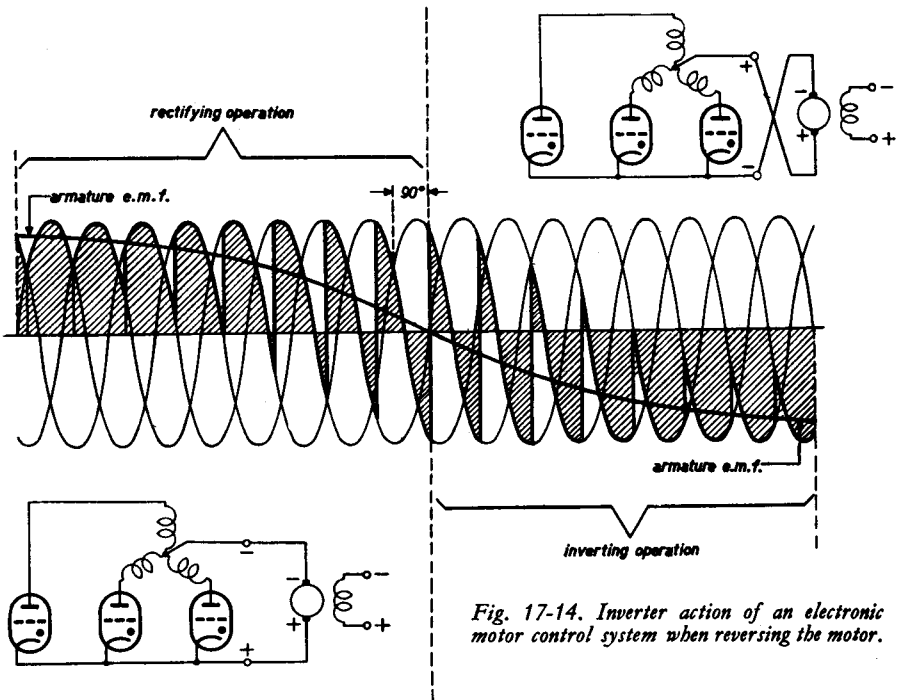


Fig. 17-14. Inverter action of an electronic motor control system when reversing the motor.

The thyratrons of the armature rectifier are first blocked so that no current flows in the armature; then the armature connections are reversed by a contactor. Since the e.m.f. of the armature is now applied to the armature rectifier circuit with reverse polarity, inverter action occurs in which the kinetic energy of the motor is converted into electrical energy and fed back into the power line (electrical braking).

This is illustrated in fig. 17-14 for a three-phase armature rectifier. The firing angle of the thyratrons is at first very large in order to avoid heavy short-circuit currents, but is gradually decreased with decreasing armature e.m.f. as indicated. At $\varphi = 90^\circ$ the voltage is zero and the motor is at rest. Further decrease of the firing angle causes the motor to start and to speed up in the reverse direction.

An improvement to this arrangement is to replace the contactor by a second set of thyratrons connected in inverse-parallel to the original tubes. A circuit of this type designed for single-phase half-wave rectification has already been described in Chapter 15 (fig. 15-5).

A somewhat different circuit, which is also intended for connection to a single-phase mains, is described below. Its simplicity renders this circuit particularly suitable for controlling small motors, such as are used for example to control the feed in automatic lathes and to advance the clutch of automatic milling machines.

In this circuit the thyratrons connected in anti-parallel are vertically controlled, but the variable direct grid voltage is derived by means of additional rectifiers from an alternating voltage, the amplitude of which is influenced by the pick-up.

By interconnecting isolating transformers, the grid circuits of the thyratrons can thus be separated, as required for tubes connected in anti-parallel.

As explained in Chapter 15, the condition is imposed on such a photoelectrically controlled installation that the two thyratrons connected in anti-parallel in the armature circuit of the d.c. shunt motor are extinguished at the mean illumination of the photocell, and that one of the thyratrons passes current during an increasing part of the half cycle as the illumination of the photocell increases or decreases, so that the rotational speed and direction of the motor are determined. When the illumination of the photocell differs from the average value, the direct grid voltage of one tube should therefore be such that it passes a larger current, whereas the direct grid voltage of the other tube should be rendered more negative or at least be maintained at the same value. In deriving

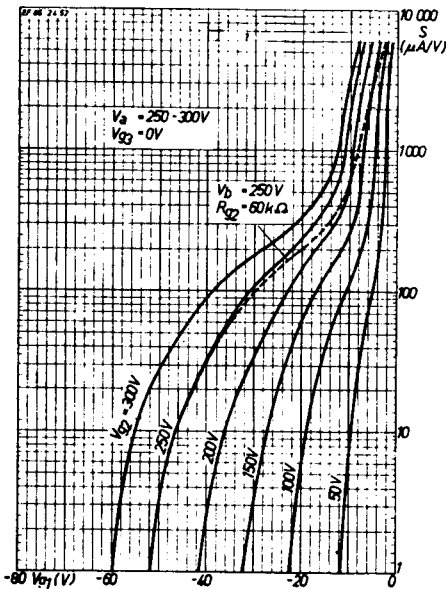


Fig. 17-15. Mutual conductance of a variable- μ tube as a function of its grid voltage with the screen-grid voltage as parameter.

be rendered more negative or at least be maintained at the same value. In deriving

these direct grid voltages from alternating voltages of corresponding amplitude, they should therefore vary with the illumination of the photocell.

This condition can be satisfied in a simple way by means of a circuit in which amplifying tubes with a variable mu are used. The control grid of such tubes consists of a helix with a variable pitch. In this way the tube is given a curved I_a/V_g characteristic the slope of which varies over a wide range. Fig. 17-15 represents the slope or "mutual conductance" of such a tube (EF 85) as a function of the

negative control grid voltage. This characteristic reveals that the mutual conductance can be varied between 1 and 6000 $\mu\text{A}/\text{V}$, which means that the amplification factor can be varied accordingly by shifting the working point. By applying a constant alternating voltage in series with a variable direct voltage to the control grid of such a variable-mu tube, the amplitude of the amplified alternating voltage will become a function of the direct grid voltage.

In the circuit of fig. 17-16 an alternating voltage $V_{g\sim}$ and a direct voltage are applied to the grid of tube T_1 ; this direct voltage consists of a constant negative voltage V_1 and a variable control voltage V_c . The voltage V_1 is so chosen that the average mutual conductance of the tube is obtained at the mean value of the control voltage V_c . At an increasing control voltage the working point is shifted to an area with a higher mutual conductance, which results in the alternating voltage across the primary of transformer Tr_1 being increased. Conversely, the tube will operate in an area with a smaller mutual conductance when the control voltage V_c is decreased, in which case the amplitude of the alternating voltage at transformer Tr_1 is reduced. At the secondary of Tr_1 an alternating voltage V_{o1} will thus be induced, the amplitude of which depends on the magnitude of V_c .

To obtain a second alternating voltage, the amplitude of which varies in the opposite direction, a second variable-mu tube, T_2 , is used. The grid voltage for this tube, derived from a voltage divider formed by the resistors R_2 , R_3 and R_4 and from the negative grid bias source V_2 , will have such a value at the mean control voltage V_c that the mutual conductance in the working point is equal to that of the tube T_1 . When the amplitudes of the alternating grid voltages $V_{g\sim}$ of both tubes are identical, the amplitudes of the alternating output voltages V_{o1} and V_{o2} will therefore also be identical. By increasing the control voltage V_c , the mean value of the direct anode current of tube T_1 , and hence the voltage drop across R_2 , is increased. As a consequence the direct grid voltage of T_2 is reduced, so that the amplification factor and the alternating output voltage V_{o2} are also reduced. Conversely, the alternating output voltage V_{o1} will de-

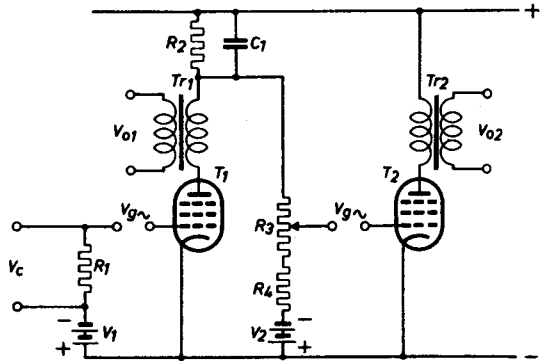


Fig. 17-16. Circuit with variable-mu tubes for controlling thyratrons.

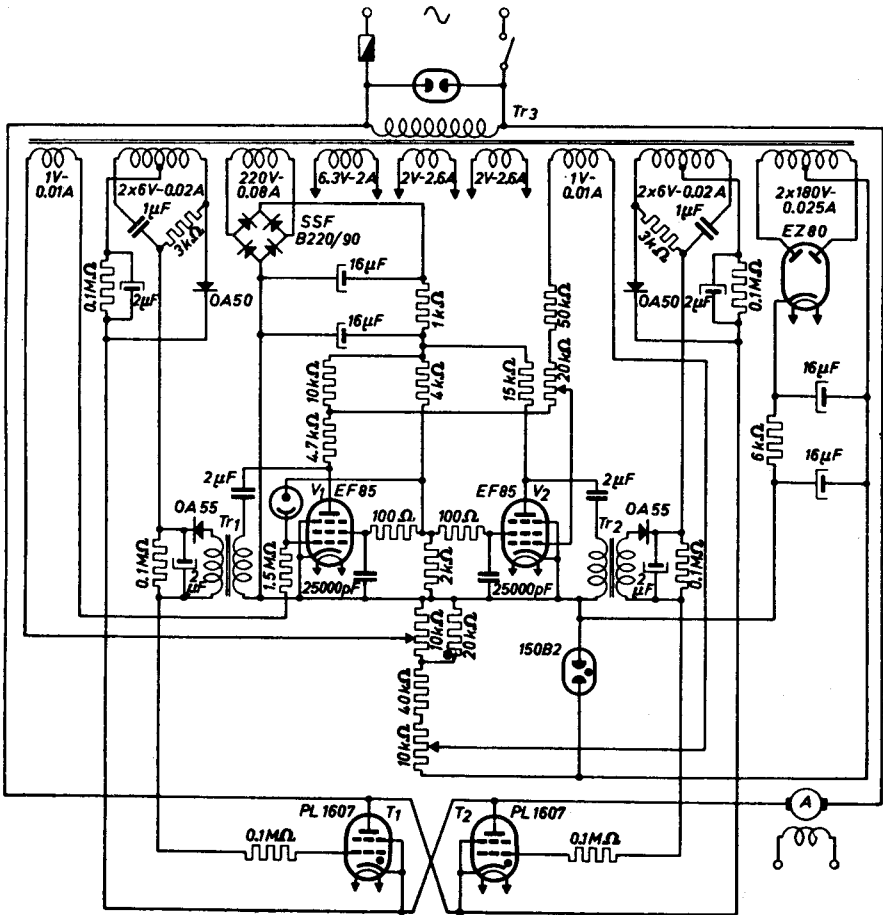


Fig. 17-17. Photoelectrically controlled rectifier for feeding and reversing motors.

crease and V_{o2} will increase when the control voltage V_c is reduced. It is thus seen that the amplitudes of V_{o1} and V_{o2} vary in opposite directions.

Fig. 17-17 shows the circuit of a photocell-controlled rectifier for feeding a motor and reversing its rotational direction; the motor is used for automatically advancing the clutch of an automatic milling machine. In this rectifier the thyatrons are controlled by means of the circuit described above.

The alternating voltages, having an r.m.s. value of 1 V, applied to the control grids of two variable- μ pentodes EF 85, are supplied by the corresponding transformer windings. The required negative grid biases are taken from a common voltage divider, fed by a rectifier which is stabilised by means of the tube 150 B2; the biases can be adjusted by means of the 10 k Ω potentiometers. The variable alternating voltages are supplied by the secondaries of the transformers Tr_1 and Tr_2 ; after having been rectified by the two germanium diodes OA 55, these voltages are fed to the grids of the two thyatrons PL 1607 connected in anti-parallel. To prevent the transformer cores from being premagnetised by

direct current, which would affect the voltage gain, the primary windings of Tr_1 and Tr_2 are separated from the anodes of the corresponding tubes by capacitors.

The voltages at the secondaries of Tr_1 and Tr_2 have been plotted in fig. 17-18 as functions of the control voltage applied to the first variable- μ pentode. At equilibrium the output voltages of the two transformers are approximately 14 V; a variation of ± 1.75 V in the control grid voltage of the first tube results in the output voltages of Tr_1 and Tr_2 being changed from 14 V to a maximum of 50 V and to a minimum of 2 V and 5 V respectively.

Since the direct voltages applied to the grids of the thyratrons must vary between positive and negative values for vertical control, the direct grid voltages derived from the variable alternating voltages are connected in series with opposed constant direct voltages the values of which are equal to half the obtainable maximum value. These

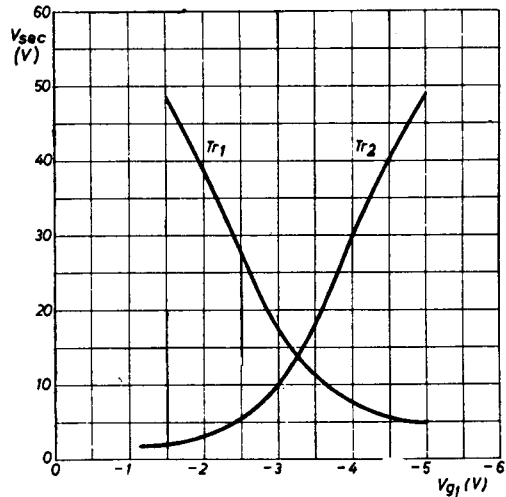


Fig. 17-18. Graph representing the control voltages obtained in the circuit of fig. 17-17 as a function of the grid voltage applied to the first variable- μ pentode.

compensating voltages are obtained by rectifying half the alternating voltages required for the phase-shifting networks by means of two germanium diodes OA 50. The alternating voltages in quadrature that are required for controlling the thyratrons PL 1607, are taken from the phase-shifting networks, which are connected in the customary way.

The constant field current of the motor is supplied by a simple metal rectifier, which has been omitted for the sake of simplicity.

The operation of an automatic

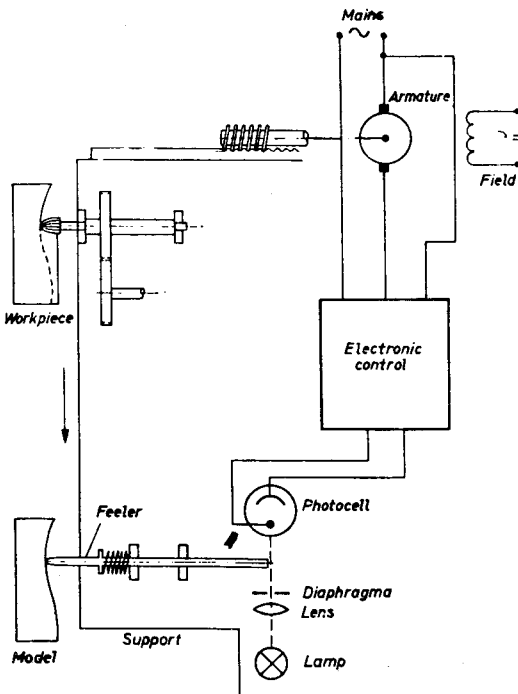


Fig. 17-19. Schematic representation of a device for controlling an automatic milling machine by means of a feeler.

milling machine that is electronically controlled by means of a feeler is shown schematically in fig. 17-19. The spindle of the cutter is mounted together with the feeler on the support, which is advanced in the direction of the arrow. When the resiliently mounted feeler is displaced from its idling position by the contours of the model, the illumination of the photocell increases or decreases, depending on the direction in which the feeler is moved; the motor then starts immediately in the corresponding direction, thus correcting the deviation. In contrast to the

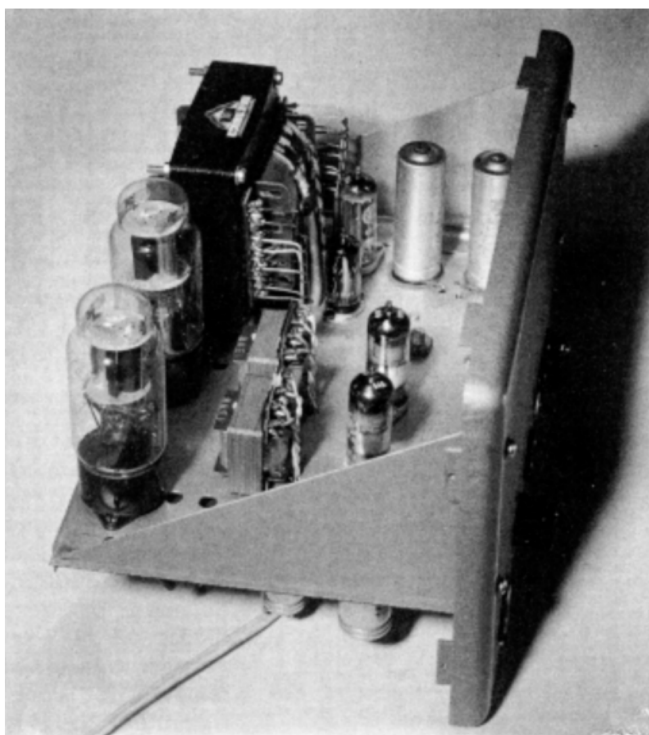


Fig. 17-20. Electronic equipment of the device represented in fig. 17-19.

customary control devices with a contact feeler, this installation operates continuously, and even sudden alterations in the form of the model are followed rapidly and without difficulty, so that a perfect copy of the workpiece is obtained.

Fig. 17-20 shows the electronic equipment of the installation, mounted on a separate chassis.

For large powers it will be preferable to design the rectifier for feeding and reversing motors in such a way that it can be connected to the three-phase mains. For supplying the armature current a rectifier with six thyratrons will then be required. Fig. 17-21 shows the complete circuit of such an installation, in which provision has also been made to limit the armature current.

The thyratrons T_4 to T_6 for supplying the armature current are fed by the

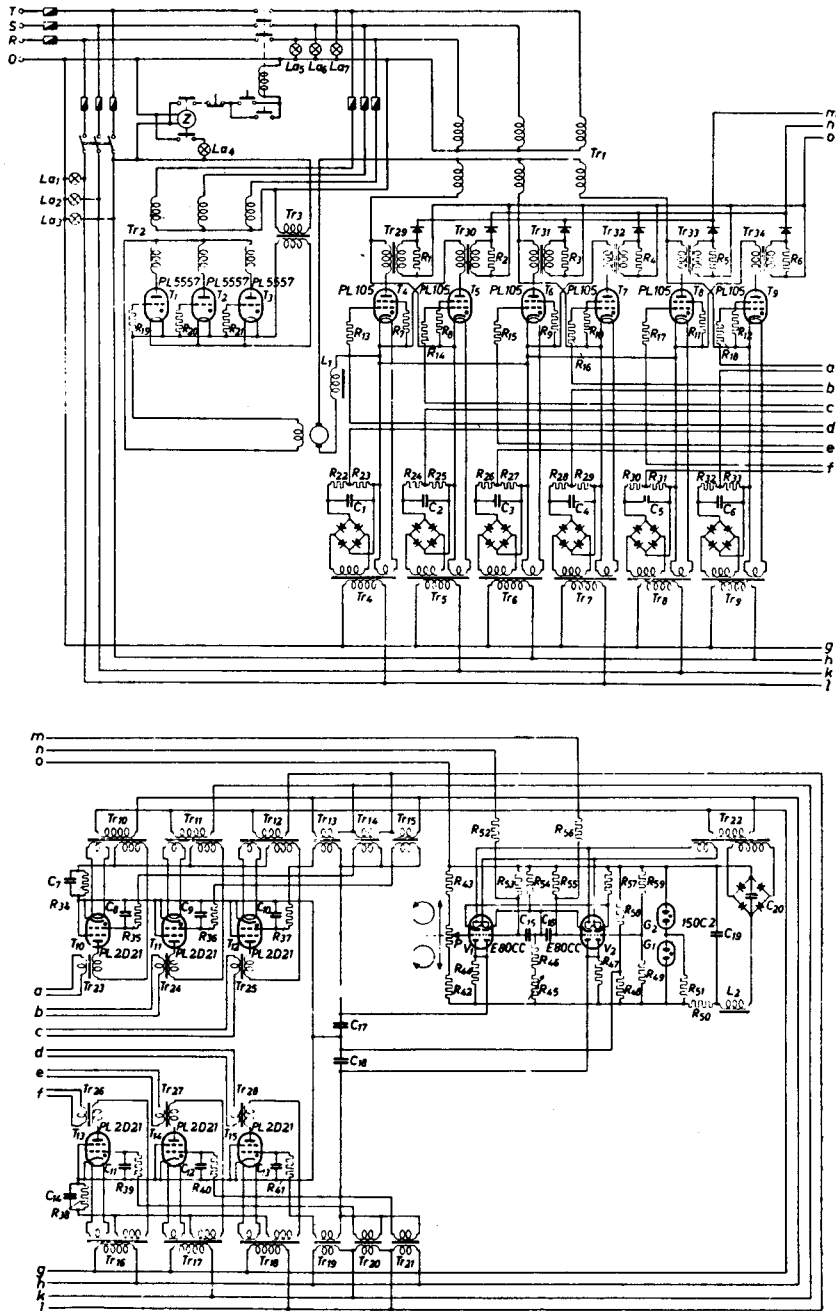


Fig. 17-21. Circuit of a three-phase installation for controlling motors.

transformer Tr_1 . The field current is constant and supplied by the thyratrons T_1 to T_3 , which are fed by the transformer Tr_2 ; the control grids of these tubes are connected to the filaments via leak resistors. The thyratrons for supplying the armature current are connected two by two in anti-parallel, so that they must be controlled horizontally by phase-shifted pulses. The fixed, negative grid bias is taken from six rectifiers with their transformers Tr_4 to Tr_9 , and the pulses are supplied by the peaking transformers Tr_{23} to Tr_{28} . These transformers are included in the anode circuits of the small thyratrons T_{10} to T_{15} . Of these tubes the thyratrons T_{13} to T_{15} serve for controlling the motor when it rotates in a clockwise direction (the large thyratrons T_4 , T_6 and T_8 then come into operation), whereas the thyratrons T_{10} to T_{12} serve for controlling the motor in the opposite direction (the armature is then fed by the large thyratrons T_5 , T_7 and T_9).

The cathodes of the small thyratrons are at the same potential, so that vertical control can be applied. Alternating grid voltages, in quadrature with the anode voltages, are then required for these tubes; these voltages are supplied by the transformers Tr_{13} to Tr_{15} and Tr_{19} to Tr_{21} , which are connected to the corresponding phases of the mains.

The variable direct grid voltages are produced across the capacitors C_{17} and C_{18} . These voltages must meet the requirements that they have the same value when the motor is at a standstill, and that they are just sufficiently negative to prevent the small thyratrons from igniting. To start the motor in a clockwise or anti-clockwise direction, the voltage across C_{18} or C_{17} respectively must become less negative and, finally, even positive, the voltage across the other capacitor being rendered more negative or at least maintained at its original value. This opposed variation of the two voltages is obtained by the control circuit equipped with the high-vacuum tubes V_1 and V_2 (E 80 CC). When the potentiometer P is adjusted to its centre position, the currents flowing through the left triode section of V_1 and through the right triode section of V_2 are equal. The anode potentials of these two tube sections are then also equal, and negative with respect to the potential at the tapping of the voltage divider formed by the resistors R_{48} and R_{58} .

If the sliding contact of the potentiometer is moved upwards (in a clockwise direction), the control grid of the left section of V_1 is rendered less negative. Owing to the larger anode current then flowing through this system, the potential at the anodes of V_1 is lowered, and since the voltage drop across the common cathode resistor R_{57} increases simultaneously, the grid of the right section of V_2 becomes more negative, and the anode potential of this section rises. The voltage across C_{18} therefore becomes more positive, and the thyratrons T_{13} to T_{15} are ignited with decreasing firing angle; the tubes T_{10} to T_{12} remain extinguished. Since the large thyratrons T_4 , T_6 and T_8 are ignited in a similar way, the motor starts in a clockwise direction.

If, on the other hand, the sliding contact of the potentiometer is shifted downwards (in an anti-clockwise direction), the anode current of the left section of V_1 decreases, whereas the current of the right section of V_2 increases. The thyratrons T_{13} to T_{15} are now extinguished, whereas the thyratrons T_{10} to T_{12} are ignited with decreasing firing angle, so that the large thyratrons T_5 , T_7 and T_9 start to pass current. Since the motor is still rotating in a clockwise direction, thus generating a corresponding e.m.f., recuperation takes place. This results

in the motor being braked very rapidly and effectively, after which it is reversed.

To limit high current peaks that may occur in the armature circuit, for example after an overhaul, the current transformers Tr_{29} to Tr_{34} are incorporated in the anode circuits of the large thyratrons; these transformers are linked according to the direction of current flow, and supply — via selenium rectifiers — two voltages which are proportional to the direction and the value of the armature current. These voltages are applied to the right section of V_1 and the left section of V_2 respectively. Normally, these two triode sections are cut off owing to the positive voltage applied to their cathodes which is taken from the divider formed by R_{45} , R_{46} and R_{54} . It will be assumed that the motor rotates in a clockwise direction; the left triode section of V_1 will then pass current, whereas the right triode section of V_2 is almost cut off. As soon as there is a risk of the armature current exceeding a certain maximum value, the voltage originating from the transformers Tr_{29} , Tr_{31} and Tr_{33} becomes so high that the left section of V_2 starts to pass current. This causes the voltage at the anode of V_2 to decrease, as a result of which the firing angle of tubes T_{13} to T_{15} increases, and this amounts to a reduction of the armature current. The right section of V_1 will start to pass current in a similar way when the armature current flowing in the opposite direction exceeds its permissible value; in that case the firing angle of the thyratrons T'_{10} to T'_{12} is increased. The value to which the armature current is limited can be adjusted by means of the variable resistor R_{45} .

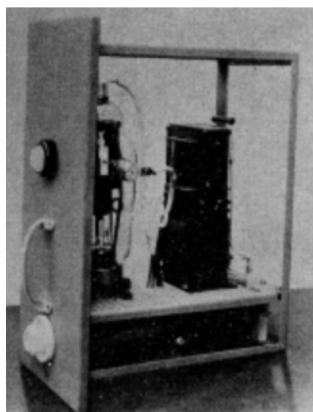


Fig. 17-23. Unit of the rectifier for supplying the armature voltage according to the circuit of fig. 17-21.



Fig. 17-22. Example of a three-phase installation for controlling motors.

Fig. 17-22 shows an installation according to the circuit of fig. 17-21. The individual thyratrons with their components and the control circuit equipped with high-vacuum tubes have been constructed as separate units, which can easily be replaced. Fig. 17-23 shows one of the

six separate units of the rectifier for the armature current. It contains a thyatron with the transformers required for the filament supply and for controlling the tube, as well as the other components for the grid control. In this way disturbances can easily be traced and removed without loss of time by simply replacing the defective unit.

In some installations the main power transformer is omitted in order to save cost, and the mains supply is rectified directly. The three-phase full-wave rectifying circuit is particularly suitable for this arrangement. As, however, the anode supply voltages are then predetermined by the mains voltage it is not possible to utilise to the full the maximum permissible inverse anode voltage rating of the thyatrons. To obtain a given motor power, therefore, it may be necessary to use thyatrons with higher average anode current ratings and these, of course, are more costly. It is thus clear that the design of a motor control system must often be a compromise between technical advantages and commercial considerations.

The following table gives suitable tube arrangements for motor control systems for various power outputs for 440-volt and 220-volt motors.

The powers quoted refer to the motor output power, a motor efficiency of 0.8 being assumed, and a 50 per cent reserve for starting and overload conditions being included.

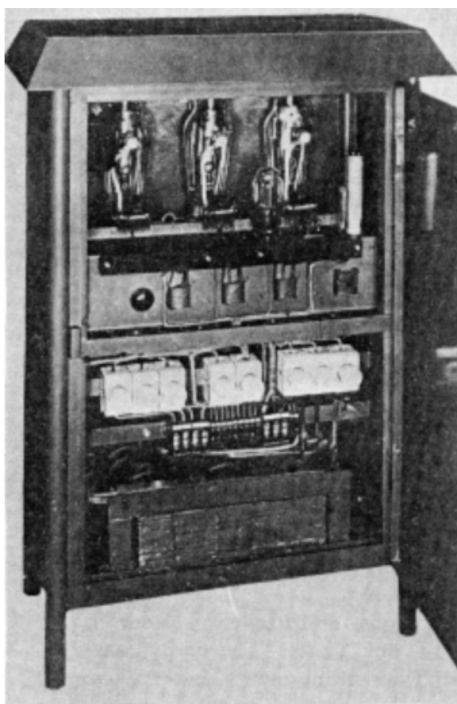


Fig. 17-24. Three-phase motor control equipment used for textile machine drives.

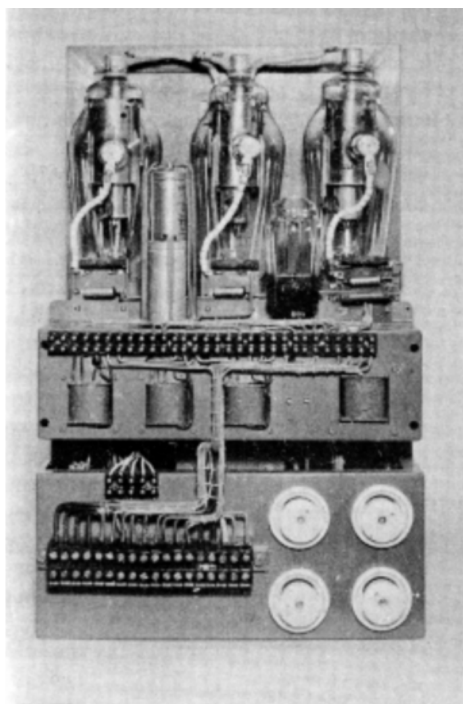


Fig. 17-25. Three-phase motor control device to be mounted into the base of a lathe.

440-VOLT MOTORS

Power (kW)	0.3	0.6	1.2	3	4.5	6	9
Armature rectifier	3× PL 5557	1× PL 5559	2× PL 5559	2× PL 105	3× PL 105	4× PL 105	3× PL 255
Field rectifier	2× PL2D21		2× PL 5557			2× PL 5559	

220-VOLT MOTORS

Power (kW)	0.3	0.75
Armature rectifier	1× PL 5559	3× PL 5559
Field rectifier	2× PL2D21	2× PL 5557

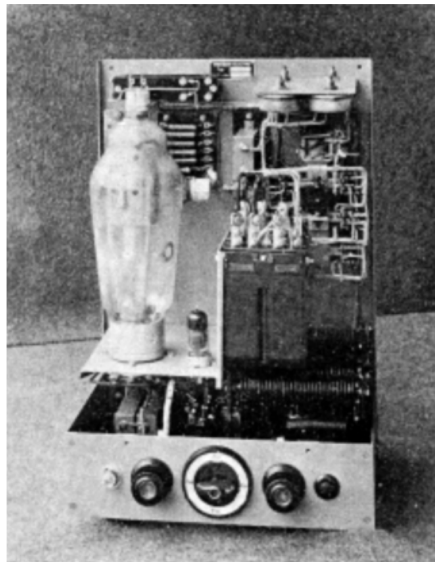


Fig. 17-26. Half-wave motor control for feed drives.

To illustrate the practical form of electronic motor control equipment, a number of photographs are reproduced in figs. 17-24 to 17-29. Fig. 17-24 shows a three-phase unit for 8.5 kW peak output power to control a textile drive and to give a range of speed regulation of 1 : 50. Fig. 17-25 is a similar device designed to be built into the base of a machine tool.

Fig. 17-26 shows a single phase control equipment especially suitable for feed drives with an output power of approximately 1 kW.

A three-phase unit incorporating armature and field control and driving a production lathe is illustrated in fig. 17-27. The nominal output power is 4.5 kW. The housing is partially removed in order to show the tubes. Part of the power transformer and the armature circuit smoothing reactor can be seen mounted

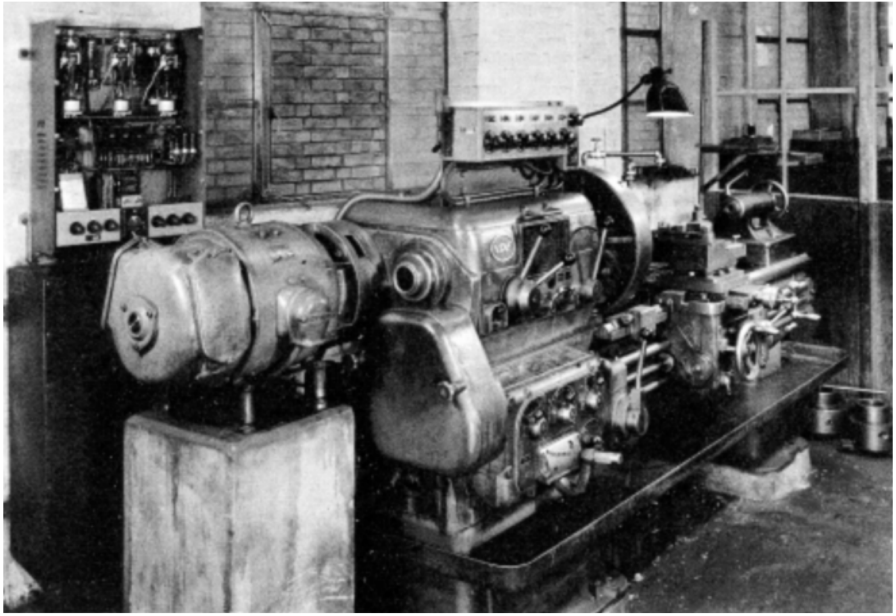


Fig. 17-27. Three-phase motor control system driving a production lathe.

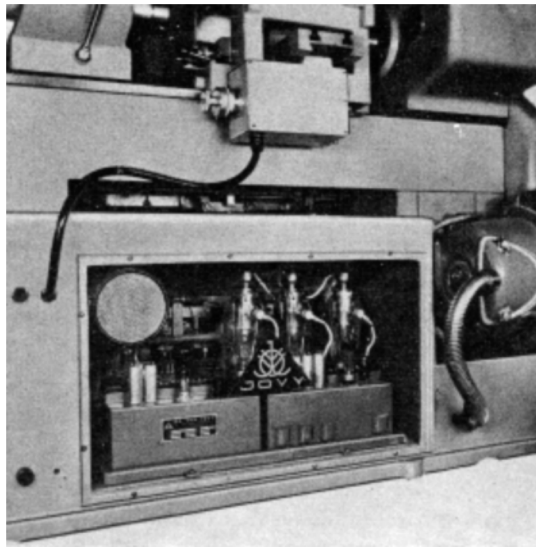


Fig. 17-28. Three-phase built-in' motor control system of a lathe.

on the wall below the apparatus. On the lathe itself is mounted an operating panel with six potentiometers and switches for pre-setting the speed and running direction. A six-way selector switch mounted alongside permits quick change-over to any of the pre-set conditions.

Fig. 17-28 illustrates an electronically-controlled drive built into the base of a production lathe. It is designed for a maximum power of 10 h.p., and incorporates armature and field control, quick reverse, quick braking, and overload protection. Fig. 17-29 shows the carriage of the same lathe, with water-tight built-in potentiometers, push-buttons, master switch and pilot lamps.

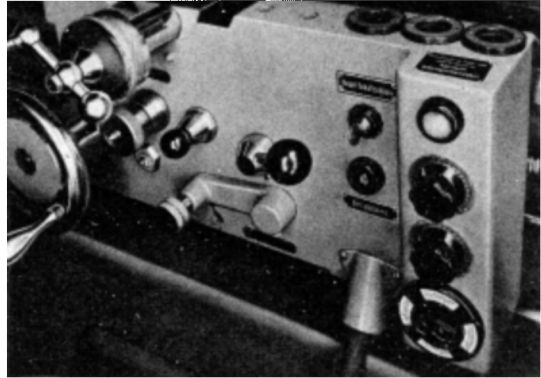


Fig. 17-29. Carriage of the lathe with built-in controls and switches.

Electronic Control of Ward-Leonard Systems

Controlled rectifiers are also often used in Ward-Leonard systems for supplying the field currents for the motor and the generator. When the speed is decreased, the surplus of kinetic energy is directly converted into electric energy in these systems, and this energy is returned to the supply mains. The driving motor then operates as a generator and feeds the generator which then operates as a motor and drives the three-phase machine, so that the latter — working as an over-synchronous generator — returns energy to the mains (recuperation).

The particularly good matching of Ward-Leonard systems to the conditions prevailing in practice can be considerably improved by using instead of a special

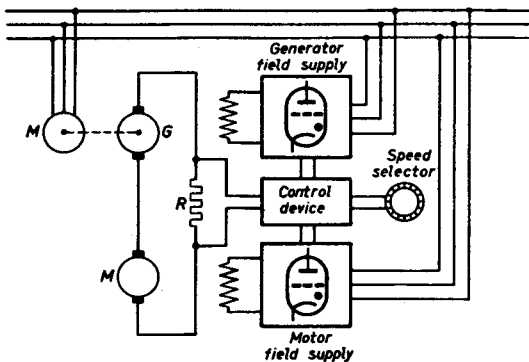


Fig. 17-30. Electronic control of a Ward-Leonard system.

field generator two electronically controlled rectifiers for the field supply of the motor and the generator (fig. 17-30). The two rectifiers are controlled electronically by means of the speed control, so that the speed is first adjusted to the nominal value by means of the armature voltage of the d.c. motor, after which the speed is increased to the final value by attenuating the field. The only difference compared with a normal installation for electronic control of the

motor therefore consists in the fact that the rectifier for the armature supply is replaced by a d.c. generator with a variable field excitation.

The voltage drop across the resistor R is used as a measure of the current flowing through the armature, and is used as a control voltage for the rectifier. When the predetermined maximum value is exceeded this will result in a decrease of the field current and hence of the armature current of the motor, so that the torque is prevented from exceeding the maximum permissible value. For reversing the motor it is possible either to reverse the field of the dynamo by means of an electromagnetic contactor, or to use a rectifier the polarity of which can be reversed.

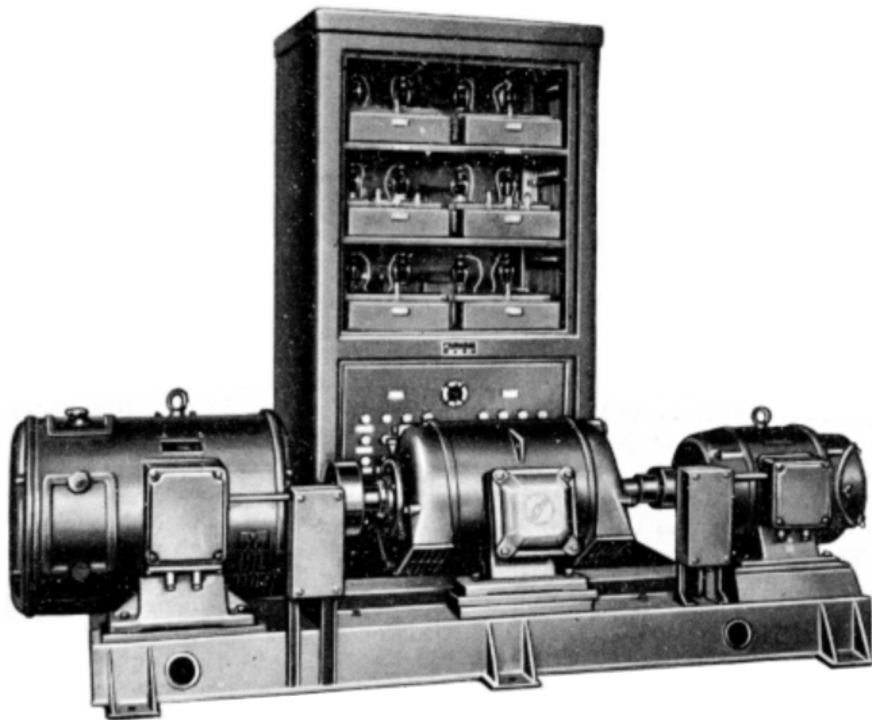


Fig. 17-31. Example of an electronically controlled Ward-Leonard system with its electronic switchgear at the rear.

Fig. 17-31 shows an electronically controlled Ward-Leonard system consisting of a three-phase driving motor (centre), a d.c. generator with a power of 30 kW (left) and one with a power of 4 kW (right). At the rear can be seen the switchgear with the rectifiers for supplying the field current for controlling these two generators and the motors which are fed thereby. By means of this installation it is possible to stabilise the speed, to limit the torque, and to decrease or increase the speed.

Finally fig. 17-32 shows an electronically controlled Ward-Leonard converter, mounted in a control desk, for feeding a spooling machine with a motor

power of 4 kW. The control desk is provided with a potentiometer by means of which the pull can be adjusted; this is indicated by the measuring instrument at the top, which is calibrated in percentages. Pushbutton switches are provided for switching on the electronic equipment and the Ward-Leonard system, and rotary switches for reversing the motor. The two rectifiers for the field supply can be seen on the rear view.

There are many different applications of electronic motor control devices. Such a control system offers particular advantages for feeding machine tools,

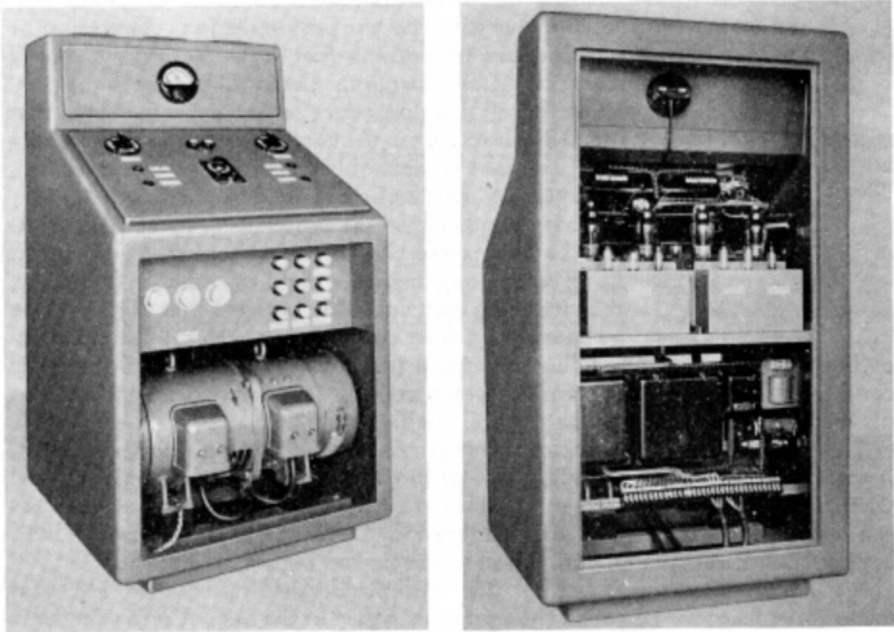


Fig. 17-32. Electronically controlled directly coupled Ward-Leonard convertor.

in installations for controlling the feed of drilling, milling, grinding and planing machines, in transport installations and lifts, in driving mechanisms for spooling various materials in the textile and paper working industries, and so forth, in brief for all driving mechanisms that have to meet special requirements as to the control characteristic, i.e. the curve representing the velocity as a function of the torque.

The cost of an electronic motor control system, including a d.c. shunt-wound motor, is now so low that these devices can compete on economic grounds with Ward-Leonard units. Compared with the latter, electronic control systems have the following advantages: (1) smaller space requirements; (2) no foundations are necessary; (3) great adaptability; (4) silent operation; (5) very low maintenance cost. To these must be added the further advantages of very low power consumption and instantaneous operation, and the ease with which the control can be adapted to meet individual requirements. The overall efficiency of electronic motor control is also higher than that of a Ward-Leonard unit, as the

latter involves three different conversions of the energy delivered by the three-phase power line before it is finally made use of.

18. HIGH-FREQUENCY INDUCTIVE HEATING OF METALS

In all branches of the metal industry the necessity arises of heating a workpiece up to a certain temperature, for instance, for hardening, casting, soldering, melting, tempering, etc. Usually this is done by generating the necessary heat outside the workpiece and applying it to the work by radiation, convection or conduction, the heat being produced by burning fuel, by some other chemical reaction, or by changing electrical energy into heat. However, all these processes have the disadvantage that it is difficult to ensure the correct heat dosage, especially if it is desired to heat only a part of the workpiece as, for example, to avoid deformation.

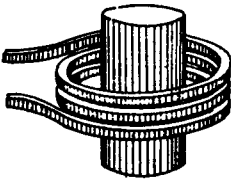


Fig. 18-1. Load coil surrounding the workpiece.

Today another method is preferred, namely, generating the heat inside part of the workpiece by placing it in an alternating magnetic field of suitable frequency, thus inducing electrical eddy currents in it. This is done by means of a load coil through which flows an alternating current of corresponding frequency (fig. 18-1). The load coil generally consists of copper tubing through which cooling water is passed in order to dissipate the heat generated in the coil itself.

The required alternating electrical energy can be supplied by a rotary generator, if the frequency is not too high, but in most cases H.F. generators equipped with transmitting tubes are used, since they are easy to install, require less room, and need practically no maintenance. Moreover, tube generators also operate satisfactorily and with good efficiency at higher frequencies.

The eddy currents generated in the workpiece cause the desired I^2R heat losses to which must be added the hysteresis losses in the case of ferro-magnetic materials. According to the laws of induction the currents in the workpiece flow in such a direction that they tend to oppose their originating source, i.e. the magnetic field. This explains the characteristic feature of inductive heating, namely that the heat generated is concentrated mainly in that part of the workpiece which is in the immediate neighbourhood of the load coil. It is therefore possible, by suitably shaping the load coil, to heat any desired part of the workpiece.

Another important feature of inductive heating is that the density of the currents induced in the workpiece is greatest at the surface, due to the so-called skin effect, and decreases very rapidly towards the interior. This is expressed with good approximation by the formula

$$I_x = I_0 e^{-x/\delta}, \quad (18.1)$$

where I_x is the current density at depth x ; I_0 the current density at the surface; and δ the penetration, that is to say the depth at which I_x is equal to I_0/e . In fig. 18-2 this function is depicted graphically. Approximately 83% of the total generated heat is concentrated on an external layer of thickness δ . The penetration δ is greatly dependent on the kind of material being heated; it is a function of

the volume resistivity ρ , the relative permeability μ , and the frequency f . The following formula is valid:

$$\delta = \frac{1}{2\pi} \sqrt{\frac{\rho \cdot 10^7}{f \cdot \mu}} \approx 503 \sqrt{\frac{\rho}{f \cdot \mu}} \text{ (mm)}, \quad (18.2)$$

where ρ is measured in ohm \cdot mm²/m and f in c/s. Since the magnetic field strength within the load coil is very high, magnetic saturation will be reached in the case of workpieces made of ferromagnetic materials such as steel. At

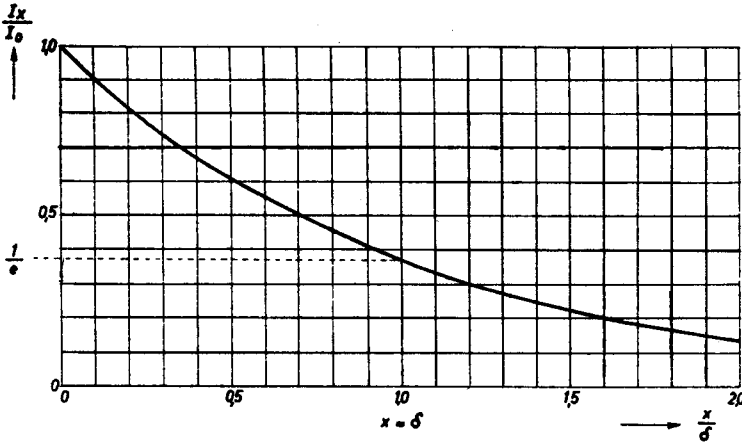


Fig. 18-2. Experimental decrease of current density towards the interior of the workpiece.

saturation the permeability μ becomes very low. If the temperature rises above the Curie point (approximately 760 °C for steel) the magnetic properties disappear, and μ becomes unity, and consequently the penetration for a given frequency increases.

In fig. 18-3 the penetration δ is shown as a function of the frequency for different materials according to equation (18.2) (*graph* = graphite, *Fe* = steel, *M* = brass, *Cu* = copper). The dotted lines are valid for room temperature, the full lines for temperatures of 800 to 1000 °C. The line for graphite is practically independent of temperature. For steel at room temperature, $\mu = 100$ has been assumed. It is seen, for instance, that the penetration is about 0.85 mm for steel heated to a temperature of 1000 °C at a frequency of 500 kc/s.

Of course, the interior of the workpiece is also heated if the high-frequency field is applied for a longer time. Thus the penetration is also a function of the heating time and the heat conductivity of the material. A part of the power applied to the load coil is lost because of the resistance of the coil, and is dissipated mainly by the cooling water. If the power loss is P_c , the total power is

$$P = P_c + P_o, \quad (18.3)$$

P_o being the useful power applied to the workpiece. The efficiency of the load coil is then

$$\eta = \frac{P_o}{P_c + P_o}. \quad (18.4)$$

Obviously the efficiency depends within certain limits, on the ratio of the volume resistivities of the workpiece to the load coil. Moreover, the ratio of the diameter of the workpiece to the penetration δ affects the efficiency, as is shown below. With certain simplifying assumptions the efficiency can be expressed:

$$\eta = \frac{1}{1 + \frac{D^2}{d^2} \left(1 + 6.25 \cdot \frac{\delta^2}{d^2}\right) \sqrt{\frac{\rho_1}{\mu \cdot \rho_2}}}, \tag{18.5}$$

D being the diameter of the load coil, d the diameter of the workpiece which

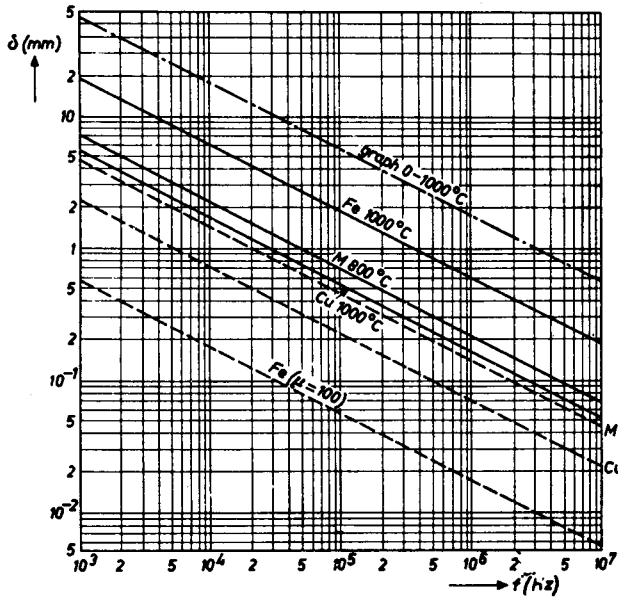


Fig. 18-3. Penetration δ against frequency for different materials (see text).

has been assumed to be cylindrical, ρ_1 the volume resistivity of the load coil, and ρ_2 that of the workpiece. Obviously the efficiency increases as the factor D^2/d^2 is made small, i.e. if the load coil tightly surrounds the workpiece. The factor $(1 + 6.25 \cdot \delta^2/d^2)$ is greater than unity. It should, however, be as small as possible if high efficiency is to be obtained. If a maximum value of 1.1 is achieved (which is generally easily attainable), then

$$\frac{\delta}{d} \leq \frac{1}{\sqrt{62.5}} \approx \frac{1}{8}, \tag{18.6}$$

i.e. the frequency must be chosen so high that the penetration is not greater than approximately $\frac{1}{8}$ of the diameter of the workpiece. In view of equation (18.2) the minimum frequency becomes

$$f_{\min} = 16 \cdot 10^6 \cdot \frac{Q_2}{\mu d^2} \quad (\text{c/s}). \quad (18.7)$$

If the frequency is chosen higher than f_{\min} , the efficiency increases only slightly, and since at the higher frequency the efficiency of the H.F. generator may become lower, f_{\min} is in practice the most favourable working frequency. In fig. 18-4 f_{\min} is shown for different materials (*graph* = graphite, *Cu* = copper, *M* = brass, *Fe* = steel) at room temperature (dotted lines) and at 800 to 1000 °C (full lines) as a function of the diameter of the workpiece.

The efficiency η as a function of the ratio D/d is indicated in fig. 18-5 for

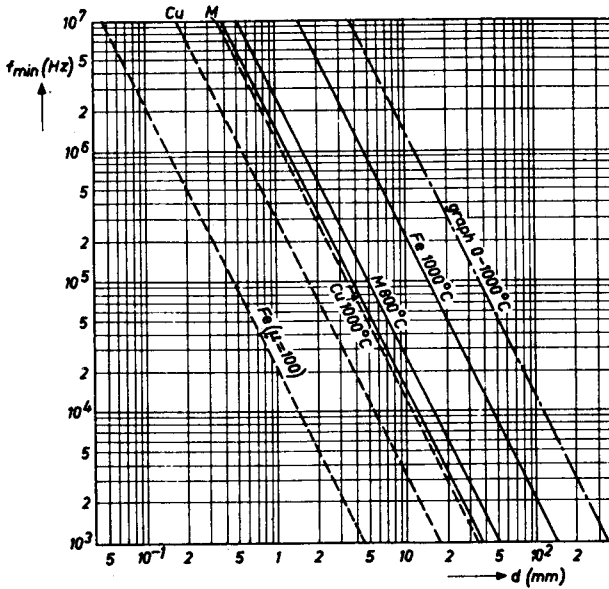


Fig. 18-4. Minimum working frequency plotted against diameter of the workpiece for various materials (see text).

the same materials, assuming that the frequency is equal to f_{\min} . In the case of non-magnetic materials, the volume resistivity of which increases with increasing temperature, the efficiency also increases if the temperature rises. In the case of steel, however, the efficiency decreases as soon as the Curie point is exceeded as the relative permeability is then unity, and the volume resistivity does not increase in the same proportion.

As can be seen from fig. 18-4, for larger workpieces with comparatively high volume resistivity a frequency of some thousands of cycles per second will suffice; if, however, steel workpieces are to be treated, for instance steel hardening processes where a penetration of less than one millimeter is required, frequencies of 500 kc/s and over must be used, and are usually produced by a tube generator. Such a device consists essentially of an oscillator of the type shown in fig. 1-11, the load coil in the simplest case being connected in series

with the inductance of the oscillatory circuit inserted in the anode lead of the oscillator tube.

The basic circuit of such a generator is illustrated in fig. 18-6; for the sake of simplicity the voltage supply sources are omitted. The anti-phase alternating grid voltage required for the excitation of oscillations is in this circuit derived from capacitor C_2 which forms together with C_1 the capacitance C of the oscillatory circuit (Colpitts oscillator). By using this capacitive voltage divider the

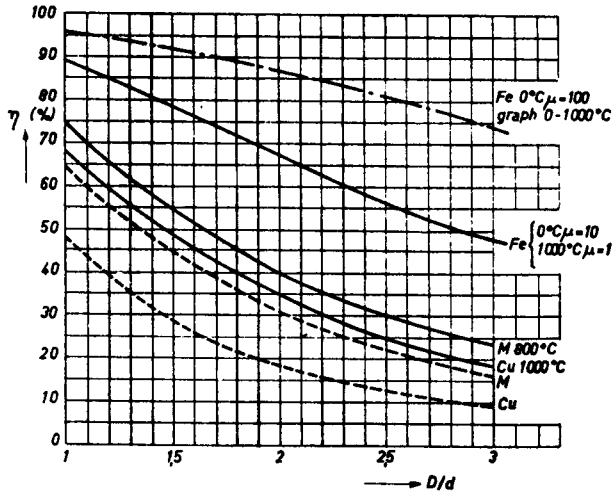


Fig. 18-5. Efficiency as a function of the ratio D/d (see text).

inclusion of the reaction coil (fig. 1-11) is rendered unnecessary, and the constructional difficulties connected with the use of such a coil in generators for inductive heating are avoided. The capacitor C_3 serves only for isolation of the grid circuit from the anode circuit for direct currents. The resistance at resonance of an oscillatory circuit containing a resistive component R is

$$R_{res} = \frac{L}{CR} \tag{18.8}$$

If the quality Q of the circuit is introduced, which may in this case be put equal to that of the coil, i.e.:

$$Q = \frac{1}{R} \sqrt{\frac{L}{C}}, \tag{18.9}$$

R_{res} becomes

$$R_{res} = Q^2 R. \tag{18.10}$$

In the case considered, the resistance R is chiefly formed by the resistance of the load coil including the workpiece. The resistance R_{res} which is present in the anode circuit of the generating tube at resonance results therefore from R

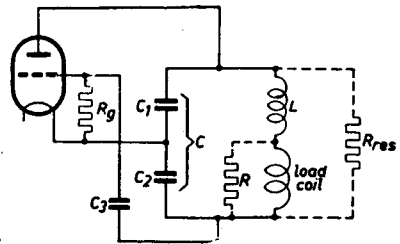


Fig. 18-6. Basic circuit of an H.F. generator for inductive heating.

multiplied by the factor Q^2 . In order to operate the tube at high efficiency, R_{res} should be as near as possible equal to the optimum anode resistance R_{opt} , which, for tubes with a tungsten cathode, can be approximately calculated from the formula

$$R_{opt} = \frac{2V_a}{I_s}, \quad (18.11)$$

V_a being the anode supply voltage and I_s the saturation current of the generator tube.

If R_{res} is smaller than R_{opt} the tube is under-excited, the output power is reduced, the anode dissipation power increases, and the efficiency is low. It is therefore necessary when constructing a generator to prevent R_{res} becoming too low, not merely because it results in low efficiency, but also to safeguard the oscillator tube since the maximum permissible anode dissipation power might be exceeded.

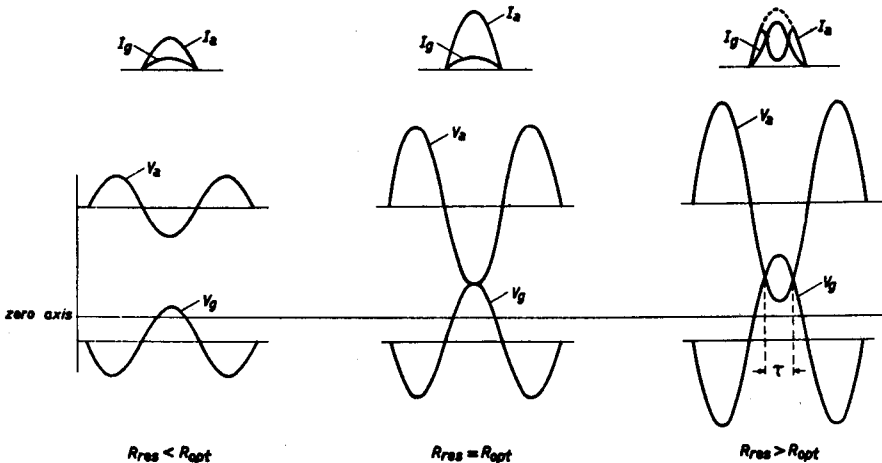


Fig. 18-7. Anode and grid voltage and current wave forms of a generator tube for different values of its anode resistance.

Somewhat similar effects occur if the circuit resistance at resonance becomes larger than the optimum resistance value. In this case the tube is over-excited; the output power decreases, but the anode dissipation also decreases so greatly that the efficiency of the tube may become even higher. This in itself would be no disadvantage but, since under these conditions the grid current increases very greatly, the danger of overloading arises because of the corresponding increase of the grid dissipation. This effect can be studied by means of fig. 18-7. If $R_{res} = R_{opt}$, i.e. under optimum operating conditions, the grid and anode voltage are of such values that the lowest instantaneous value of the anode voltage is nearly equal to the highest instantaneous value of the grid voltage. Anode and grid are therefore for a short time within each cycle at approximately the same potential. Under over-excited operating conditions, however, the amplitude of the alternating component of the anode voltage is increased so that during a certain time interval the grid is at a higher potential than the anode. Moreover, the alternating grid voltage can also increase due to the constant

feedback resulting from the increasing alternating anode voltage. Thus the grid current is considerably increased while the anode current is correspondingly reduced, losing its sinusoidal form and developing the form shown on the right in fig. 18-7. These graphs explain clearly the behaviour of the tube as regards output power, efficiency, etc. and emphasize the necessity of avoiding overloading the anode or the grid.

The optimum anode resistance of normally used oscillator tubes is generally about some thousands of ohms. If, for instance, a value of 2500 ohms for R_{opt} and a circuit Q of 50 is assumed, the resistance of the load coil should be 1 ohm, according to equation (18.10). However, the resistance of the workpiece is considerably lower (some milliohms). A matching transformer is therefore

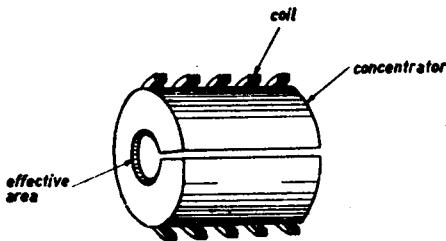


Fig. 18-8. Concentrator forming a matching transformer with a ratio of 5 : 1.

usually necessary and must be connected between the workpiece and load coil. The coil of this device generally consists of a single turn which at the same time allows the high frequency alternating field to be concentrated on a limited area of the workpiece. It is therefore called a concentrator.

If, for instance, the resistance of the workpiece is 10 milliohms, the resistance ratio in the case considered is 100 : 1. Since resistances are transformed in a ratio equal to the square of the transformer ratio, a matching transformer having a turns ratio of 10 : 1 will then be required.

A concentrator is shown in fig. 18-8, consisting of a copper cylinder which, together with the surrounding load coil, forms a matching transformer with a ratio of 5 : 1. The current induced in the cylinder tends to flow around the

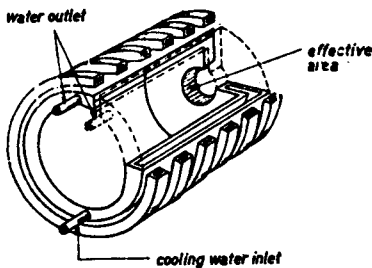


Fig. 18-9. Water-cooled concentrator.

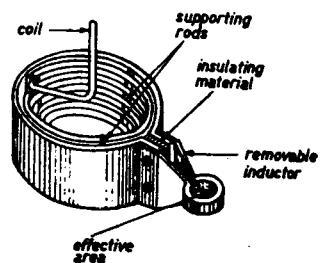


Fig. 18-10. Concentrator equipped with removable inductor coil.

outside, but because of the slot it is forced to flow along the slot and the small inner surface. Thus a very high flux density is achieved. Fig. 18-9 shows a concentrator of similar form having double walls which are water-cooled in order to dissipate the heat losses. In fig. 18-10 another type of concentrator is illustrated, possessing a removable inductor coil in order to adapt it to workpieces of different dimensions.

One of the most important applications of H.F. inductive heating is the

hardening of steel. This process can be done very easily if sufficient energy concentration is present since heating occurs only at the surface of the material. The heating time is generally only a fraction of a second, the exact time depending upon the energy concentration which will be usually from 1 to 5 kilowatts per square centimeter. All kinds of steel, the carbon content of which is greater than 0.3%, can be satisfactorily

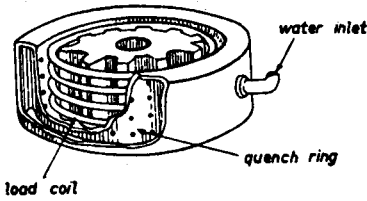


Fig. 18-11. Arrangement for hardening cog-wheels.

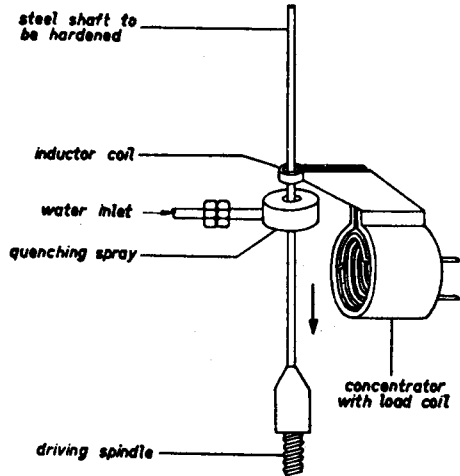


Fig. 18-12. Experimental arrangement for hardening a steel shaft.

treated. Steels with a carbon content up to 0.8% give better hardening results by H.F. heating than by any other method hitherto employed. A great advantage is that the quenching spray can be mounted in the immediate neighbourhood of the heated part of the workpiece. As an example, fig. 18-11 shows how, in

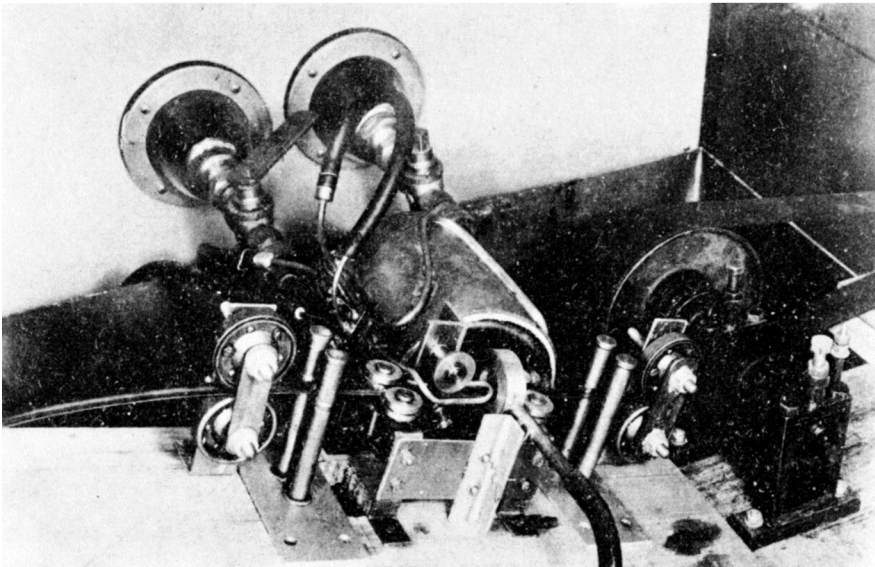


Fig. 18-13. Arrangement for hardening endless saw blades.



Fig. 18-14. High frequency soldering of capacitor housings.

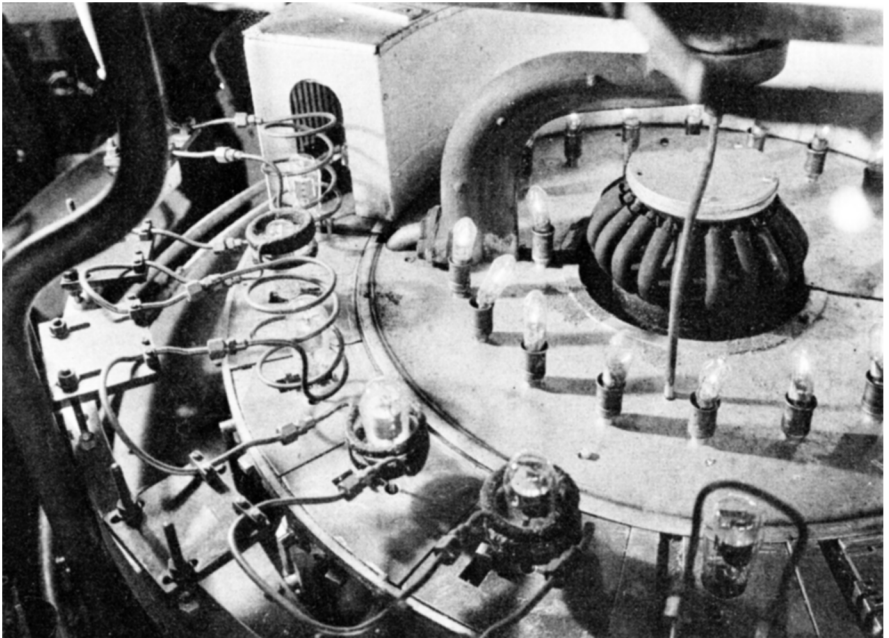


Fig. 18-15. Rotating pump in radio tube manufacture equipped with HF heating arrangement in order to release gas from electrode assemblies.

hardening cogwheels, a quench ring (which must not consist of metal in this case) surrounds the load coil so that the cooling liquid is sprayed on the work-piece through small holes between the turns of the coil.

Fig. 18-12 shows an experimental arrangement for hardening a steel shaft (6 mm diameter) which is fed by the spindle shown below, at a speed of two meters per minute through the inductor coil of a concentrator. Immediately below the inductor coil is mounted the quenching spray. The load coil is placed in the interior of the cylindrical concentrator.

The arrangement shown in fig. 18-13 serves for hardening endless saw blades which run below the inductor coil at a speed of 2 m/min. The shape of the inductor coil ensures that both the toothed and the plain edge of the blade is heated and hardened, thus eliminating deformation. The central part of the blade remains soft, maintaining the required flexibility.

Inductive heating is also very suitable for soft or hard soldering. Fig. 18-14 shows the soldering of capacitor housings in mass production by means of a 6 kW H.F. generator. The load coil consists of metal bands below which the housings are fed. On the left-hand side the longitudinal seams are soldered, on the right hand side the bottoms.

Finally, a very important application of inductive heating is found in the manufacture of electronic valves and tubes. The electrode systems are heated by high frequency during the evacuating process in order to release all gas from the electrode materials. Fig. 18-15 shows a view of part of a rotating pump used for evacuating radio tubes. The load coils are movable vertically; they are of various shapes depending on the heating effect desired in different parts of the assembly.

The operation of a generator for inductive heating does not differ in principle from that of a generator designed for capacitive heating of dielectric materials, and a full description of complete circuits is therefore reserved for Chapter 19, which deals with capacitive heating.

19. HIGH FREQUENCY CAPACITIVE HEATING OF DIELECTRIC MATERIALS

Materials which are not good conductors of electricity, such as wood, textiles, cork, and various plastics can be heat-treated by means of a high frequency electric field. This method of heating is based on the physical conception that the molecules of the material to be treated are polarised under the influence of an electric field and set themselves in the direction of the field. If the direction of the field is changed the molecules re-orient themselves accordingly so that, if subjected to a high frequency alternating field, the molecules change their direction with the same frequency. Due to inter-molecular friction, heat is generated in the material.

It is obvious that the quantity of heat developed in unit time, i.e. the electrical power transformed into heat, increases as the frequency of the field increases. For this reason the frequencies employed for capacitive heating are considerably higher than for inductive heating.

The material to be heated acts as the dielectric between the plates of a capacitor which is coupled in a suitable manner to the anode tank circuit of a high frequency generator. In an "ideal" capacitor, i.e. one which has no losses, the

current leads the voltage by 90 degrees. However, if losses occur in the dielectric, due, for example, to internal friction between the molecules as already mentioned, and to the fact that the dielectric is not a perfect non-conductor, the current no longer leads by 90 degrees but by some smaller angle. The difference between this angle and 90 degrees is called the loss angle. Fig. 19-1 shows the corresponding vector diagram. The equivalent network of a capacitor having losses is the parallel connection of an ideal capacitor C and a "loss" resistance R . The resulting impedance is:

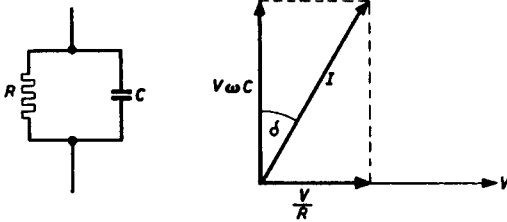


Fig. 19-1. Vector diagram of a capacitor having losses.

The equivalent network of a capacitor having losses is the parallel connection of an ideal capacitor C and a "loss" resistance R . The resulting impedance is:

$$Z = \frac{1}{\frac{1}{R} + j\omega C}, \tag{19.1}$$

and the loss angle,

$$\delta = \tan^{-1}\left(\frac{1}{\omega RC}\right). \tag{19.2}$$

Hence the loss resistance, which determines the amount of heat generated in the workpiece, is:

$$R = \frac{1}{\omega C \tan \delta}, \tag{19.3}$$

and the power transformed into heat is:

$$P = \frac{I^2}{\omega C} \cdot \frac{1}{\tan \delta} = \omega C V^2 \tan \delta, \tag{19.4}$$

where V is the r.m.s. value of the alternating voltage across the capacitor. If the electric field between the plates of the capacitor is assumed to be uniform, and edge effects are neglected, equation (19.4) can be rewritten:

$$P = 0.556 \times 10^{-10} \cdot \epsilon \tan \delta E^2 f \frac{m}{\rho} \text{ watt}, \tag{19.5}$$

where ϵ is the (relative) dielectric constant;

m the mass in kg;

ρ the density in kg/m^3 of the material to be treated;

E the electric field strength in V_{rms}/m , and

f the frequency.

The temperature rise dT in a material is proportional to the amount of heat applied and inversely proportional to the mass m and the specific heat c , so that, radiation and conduction losses being neglected,

$$\frac{dT}{dt} = \frac{P}{mc}. \tag{19.6}$$

Combining equations (19.5) and (19.6):

$$\frac{dT}{dt} = 0.556 \times 10^{-10} \cdot \frac{\epsilon \tan \delta}{\rho c} \cdot E^2 f. \quad (19.7)$$

Since ϵ and $\tan \delta$ are only slightly dependent upon frequency in the case of most normal materials of the types under consideration, the factor $\epsilon \tan \delta / \rho c$ can be regarded as a material constant.

The temperature rise in unit time is thus proportional both to the frequency and to the square of the field strength. For the latter, the upper limit is ap-

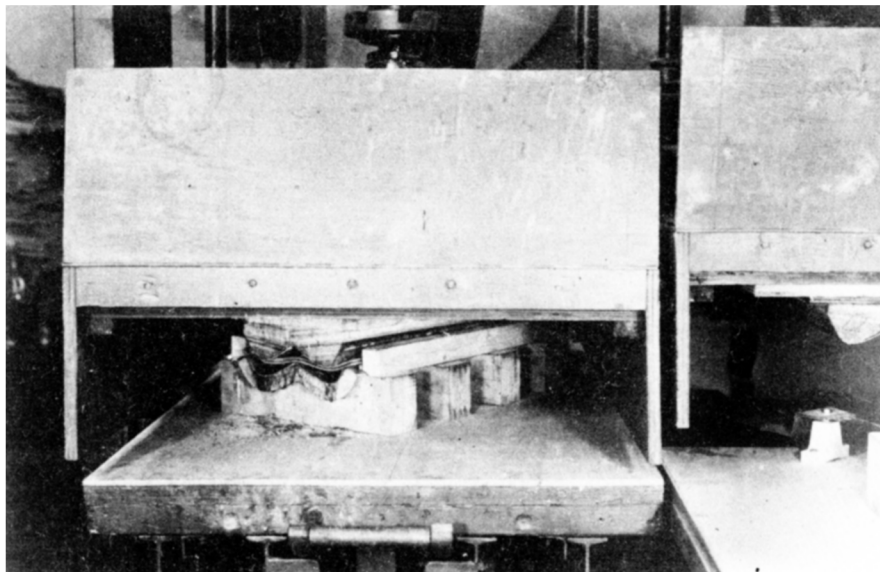


Fig. 19-2. Inserting veneers dressed with glue into a press.

proximately 10^6 V/m or 100 V/mm. This value may be used for rapid heating without involving risk of spark discharge through the workpiece. If, however, steam is released from the material during the heating process, it may condense at the electrodes, thus increasing the risk of flash-over. It is then recommended that the field strength be reduced and, if possible, a higher frequency be employed.

The most important characteristic of high frequency capacitive heating is that the generation of heat is uniform throughout the material. This is especially valuable in view of the fact that materials which are poor electrical conductors are also bad conductors of heat. Heat applied externally by radiation or conduction therefore penetrates only very slowly to the interior of the material while the external parts are exposed to high temperatures, which may even be harmful to the workpiece. This disadvantage is avoided in high frequency heating and the method is therefore ideal for many applications such as the manufacture of plywood. If in the press the steam heating hitherto used is replaced by high frequency heating, the capacity of the press is very greatly

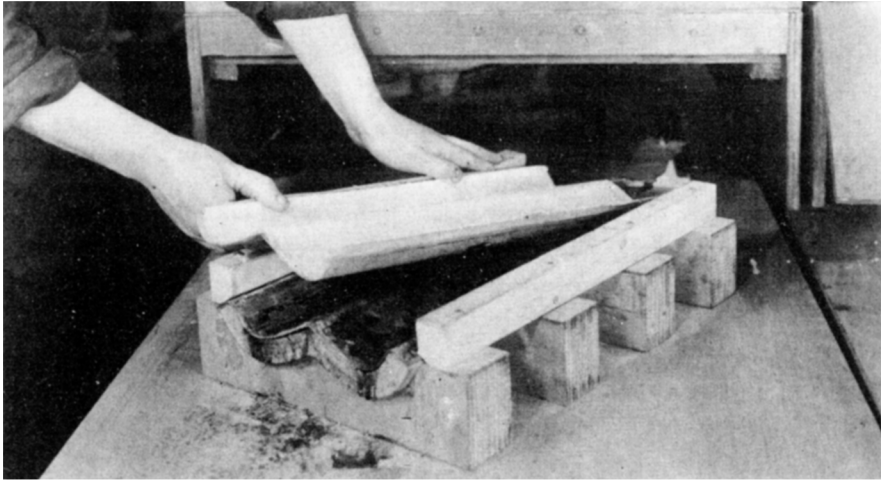


Fig. 19-3. Removing the completed workpiece.

increased; the useful space inside the press is in most cases nearly doubled by removal of the steam-heated plates.

Another important advantage of high frequency heating is the very much shorter time required for glueing processes. The quality of plywood manufactured by means of high frequency heating is excellent since the moisture content of the material is practically unchanged. Usually a frequency of approximately

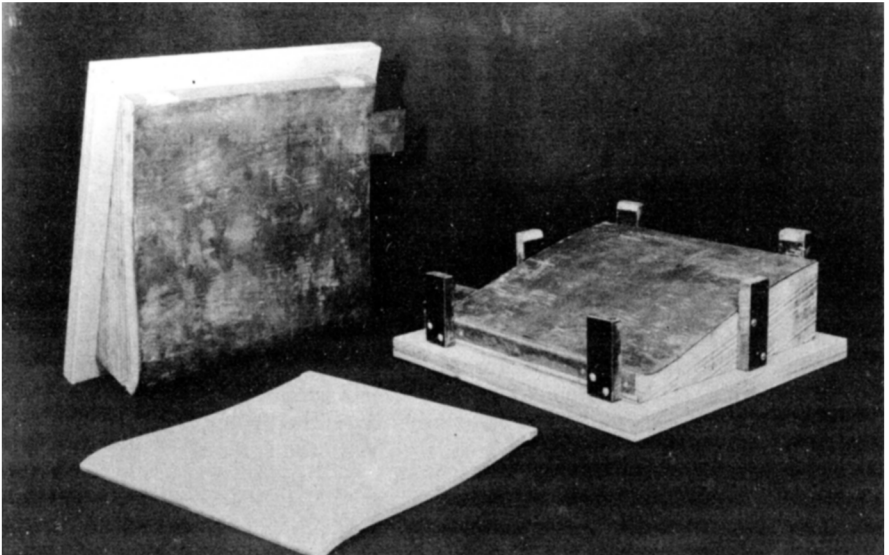


Fig. 19-4. Jig used for the manufacture of curved seats.

2.5 Mc/s is used for plywood glueing. The heat generated remains in the material for a considerable time so that no further energy is needed during polymerisation of the glue.

Very hard and resistive sheets can easily be made from chips or shavings pressed together with a binding medium under heat generated by high frequency. When glueing plywood or veneers by means of high frequency heating the workpiece can be bent or formed to any desired contour by a suitable jig. This method is used, for instance, in the manufacture of cabinets for radio sets. Fig. 19-2 shows veneers dressed with glue being loaded into a press, the jig consisting of wood covered with thin aluminium sheets, which form the plates of the load capacitor and are connected to the high frequency generator. Fig. 19-3 shows the finished work being removed from the jig. No after-treatment is necessary and the workpiece can be passed on immediately to the next stage in manufacture. The curvature of the workpiece does not change in any way after treatment in contrast to the deformation experienced with older methods of heating.

In the manufacture of curved components for furniture high frequency heating also offers many advantages. Fig. 19-4 shows a jig consisting of hard wood covered with aluminium sheet used for the manufacture of curved seats, together with a specimen of the finished article.

In the plastics industry it is necessary to pre-heat the charges of plastic material to make them flexible before inserting them in the press. For this purpose the charges, in pellet form, are heated up to 105 to 115 °C for a short period in a specially designed high frequency generator. This method confers many advantages. In the first place the product can be manufactured in nearly half the time previously required and has a much better surface finish. The greater the thickness of the product, the greater is the gain resulting from the reduction in operating time. Again, the working pressure required in the moulding machine is only 50 to 75% of that required in the



Fig. 19-5. 2 kW H.F. generator for preheating plastic material.

older method so that wear on the matrix is reduced by about half.

Plastic powder can also be preheated by high frequency in special vessels consisting of a material having low dielectric losses. A commercial 2 kW high frequency generator (Philips Type SFG 136/21) designed for pre-heating plastics is illustrated in fig. 19-5.

In view of the high frequencies (up to 40 Mc/s and more) used for the capacitive heating of dielectric materials, the only suitable form of high frequency generator is the tube oscillator type. The capacitor formed by the electrodes and the workpiece must be coupled in a suitable manner to the anode tank circuit of the generator tube. Since, however, the "loss resistance" R (see fig. 19-1) is usually considerably smaller than the optimum load of the tube,

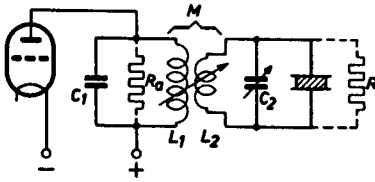


Fig. 19-6. Load circuit coupled inductively to the anode tank circuit.

problems arise which are similar to those encountered in high frequency inductive heating.

In the simplest case matching can be achieved by a transformer coupling between the load circuit and the anode circuit of the generator tube, as shown in fig. 19-6. If the resistance of the anode tank circuit at resonance is R_a under full-load conditions and R_1 under no-load conditions, then:

$$R_a = \frac{R_1}{1 + k^2 Q_1 Q_2} \tag{19.8}$$

where the quality factor of the anode tank circuit is:

$$Q_1 = \omega C_1 R_1 \tag{19.9}$$

and the quality factor of the load circuit:

$$Q_2 = \omega C_2 R. \tag{19.10}$$

C_2 is the sum of the capacitance of the load capacitor and the variable capacitor used for tuning the load circuit, and R is the resistance of the workpiece. It is assumed that R is small compared with the resistance R_2 of the load circuit at resonance under no-load conditions, so that the latter can be neglected when connected in parallel.

$$k = \frac{M}{\sqrt{L_1 L_2}} \tag{19.11}$$

is the coupling coefficient. It is thus clear from equation (19.8) that the resistance value transferred to the anode tank circuit can be varied over a wide range by adjusting the coupling M of the circuits.

Another method of matching is indicated in fig. 19-7. The primary resistance of the four-terminal network consisting of L , C_1 , L_1 and C is

$$R_a = nR^2, \tag{19.12}$$

where

$$n = \frac{L}{L + L_1} \cdot \frac{C + C_1}{C}. \tag{19.13}$$

By suitable choice of L_1 and/or C_1 , n can be given any desired value, thus matching the load.

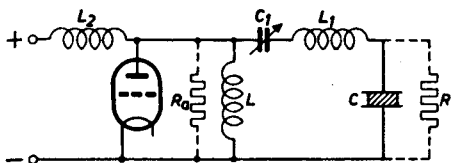


Fig. 19-7. Matching the load by an LC network.

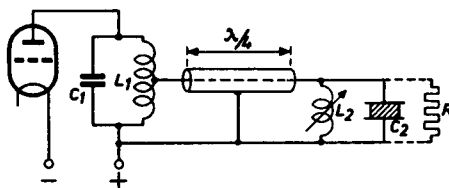


Fig. 19-8. Matching the load by a concentric transmission line.

A third method, often used, is shown in fig. 19-8. Here a concentric transmission line is used, the length of which may be, for instance, a quarter of the wavelength. If the losses in the transmission line can be neglected, the following general equations are valid:

$$V_i = V_o (\cos \alpha + j \cdot \frac{Z}{R_o} \cdot \sin \alpha), \quad (19.14)$$

$$I_i = \frac{V_o}{R_o} (\cos \alpha + j \cdot \frac{R_o}{Z} \cdot \sin \alpha). \quad (19.15)$$

where

V_i = r.m.s. input voltage to the transmission line;

I_i = r.m.s. input current in the line;

$Z = \sqrt{L/C}$ = surge impedance of the line;

V_o = r.m.s. output voltage of the line;

R_i = input impedance;

R_o = output impedance;

α = electrical length of the line in degrees

$$= \frac{\text{length of line}}{\text{wavelength}} \cdot 360^\circ.$$

From (19.14) and (19.15):

$$R_i = \frac{V_i}{I_i} = \frac{Z (R_o + jZ \tan \alpha)}{Z + jR_o \tan \alpha}. \quad (19.16)$$

The impedance-matching characteristics of such a line become evident from consideration of a quarter-wave line such as is normally used. In this case:

$$V_i = jV_o \cdot \frac{Z}{R_o}, \quad (19.17)$$

and

$$I_i = j \cdot \frac{V_o}{Z}, \quad (19.18)$$

so that

$$R_i = \frac{Z^2}{R_o}. \quad (19.19)$$

It is thus seen that the output impedance is reflected with the square of the surge impedance to the input side, so that a low impedance at the output side appears as a high impedance at the input side. If the output side is short-circuited ($R_o = 0$), the theoretical value of the input impedance is infinite, i.e. the resistance present in the anode circuit of the generator tube is merely that provided by the resistance of the tank circuit comprising L_1 and C_1 at resonance and its loss resistance.

Equation (19.14) shows that the output voltage V_o appearing across the workpiece is a function of R_o if V_i is assumed to be constant. From equation (19.7) it is clear that for rapid heating a high field strength, i.e. a high voltage, should be applied across the load capacitor, so that R_o should be as high as possible. This can be achieved by connecting a variable inductance L_2 in parallel with the load capacitor, and tuning the oscillating circuit formed by L_2 and C_2 to resonance, to ensure that the resistance of the parallel connection of the capacitor and inductor is a maximum.

5 kW High Frequency Generator

In the following paragraphs is described the design of a high frequency generator giving an output of approximately 5 kW at a frequency of 27.2 Mc/s. Such a generator would, for example, raise the temperature of 4800 cm³ of wood by about 37 °C in one minute, or would treat a layer of glue with a surface area of 1500 cm² in only fifteen seconds.

The complete circuit of the generator is shown in fig. 19-9. It should be pointed out that great care must be taken in the layout and construction of high frequency generators in order to avoid parasitic oscillations. For this purpose filters and damping resistors are included at suitable points, their location depending upon the arrangement of the circuit components and the lengths of the various leads. The construction of high frequency generators thus demands specialised experience which can be acquired only by long practical work in the field of high frequency technique.

The input power of this generator is 4.2 kW at no load and will increase up to 16.2 kW at full load with a power factor of 0.9. The equipment consists of two parts: the power pack including the necessary transformers, rectifying tubes, relays, switches and measuring instruments; and the oscillator which generates the high frequency energy. So that the equipment may be adopted for different applications, two models (*a*) and (*b*) of the oscillator are provided. Model (*a*) is especially suitable for use in mass production, and model (*b*) is more suitable for experimental work.

In model (*a*) the positive terminal of the voltage supply is earthed and the anode tank circuit is connected to the load by means of a quarter-wave transmission line. The coil L_4 is provided with taps, not shown in the diagram, so that it can be tuned to the most suitable value for the particular work being done. Such load fluctuations as may then occur during operation do not greatly influence either the frequency or the output power of the generator.

In model (*b*) the anode of the oscillating triode is connected to the positive terminal of the power pack via the high frequency reactor coil L_8 , the negative terminal being earthed. In order to simplify operation with different values of load capacitance and output power, the generator is adjusted initially for maximum output and no further adjustment is needed. The circuit is so designed

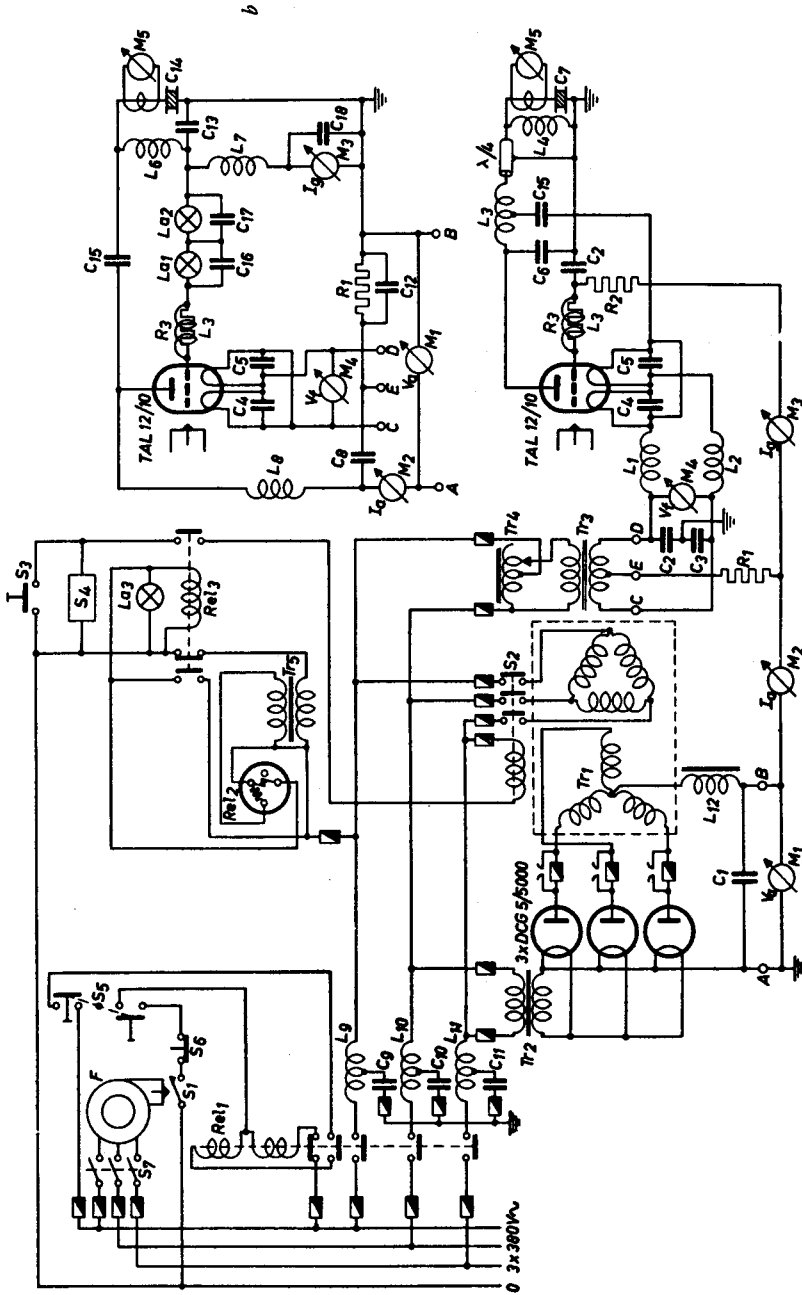


Fig. 19-9. Complete circuit of a 5 kW H.F. generator.

that no overloading of the generator tube occurs in the event of incorrect matching.

The high frequency generator tube is an air-cooled triode with external anode, Philips Type TAL 12/10. For rectifying the current supplied by the mains transformer three high voltage rectifying tubes Type DCG 5/5000 are employed. The fan F has a capacity of 7.5 m³ per minute and is driven by a motor of approximately 1 h.p.

The main contactor Rel_1 is operated by a push-button switch S_5 ; a door contact S_6 and air-vane relay S_1 prevent the generator from starting if the housing doors are open or if the fan is not running. A mains filter $L_9, L_{10}, L_{11}, C_9, C_{10}, C_{11}$ prevents the high frequency energy from reaching the power line. A separate contactor S_2 , operated by push-button switch S_3 connects the power transformer Tr_1 with the mains. The bi-metal relay Rel_2 provides the pre-heating time for the gas-filled rectifying tubes. Instead of S_3 a mechanical or electronic timer, indicated by S_4 , can be used and can be adjusted for any desired treatment time of the workpiece.

Since in model (a) the positive terminal of the power pack is earthed, the secondary winding of filament transformer Tr_3 together with the connecting leads of the cathode of the generator tube, and also the secondary windings of the power transformer Tr_1 must be insulated against high voltage. This also applies to the whole of the grid circuit of the generator tube. The secondary of filament transformer Tr_2 , however, need not be insulated against high voltage in this case. In model (b) the negative terminal of the power pack is earthed so that only the secondary winding of Tr_2 needs to be insulated against high voltage.

The fan is switched on by means of S_7 . About two minutes after switch S_5 has been closed the pilot lamp La_3 will light, indicating that the generator is ready for operation. A workpiece can now be placed between the electrodes, and heat treatment can be started by operating push-button switch S_3 or timing switch S_4 .

As previously mentioned, for model (a) the load circuit must be tuned to the generator frequency by adjusting coil L_4 , by choosing a suitable tapping. It is recommended that this adjustment be made under reduced anode voltage. This can be achieved by removing one of the fuses inserted in the primary circuit of power transformer Tr_1 .

H.F. Generator with High Frequency Stability

At the international conference held in Atlantic City it was agreed that in order to avoid radio interference one of the most suitable frequency bands allotted for industrial and electro-medical purposes, viz. the 40.680 Mc/s band, should be adhered to with an accuracy of $\pm 0.05\%$. Since it is practically impossible to achieve such a high degree of frequency stability with a self-oscillating circuit, a generator with a crystal-controlled driving stage must be used.

In fig. 19-10 is shown the circuit of an equipment in which the small transmitting tetrodes V_1 to V_4 , Type QE 04/10, are used in the oscillator, frequency-multiplier and driver stages, a larger transmitting tetrode V_5 , Type QB 3.5/750 being used in the final stage and giving an output of about 500 watts. The use of tetrodes offers the great advantage of requiring relatively small drive power.

The master oscillator consists of a crystal-controlled tetrode V_1 which

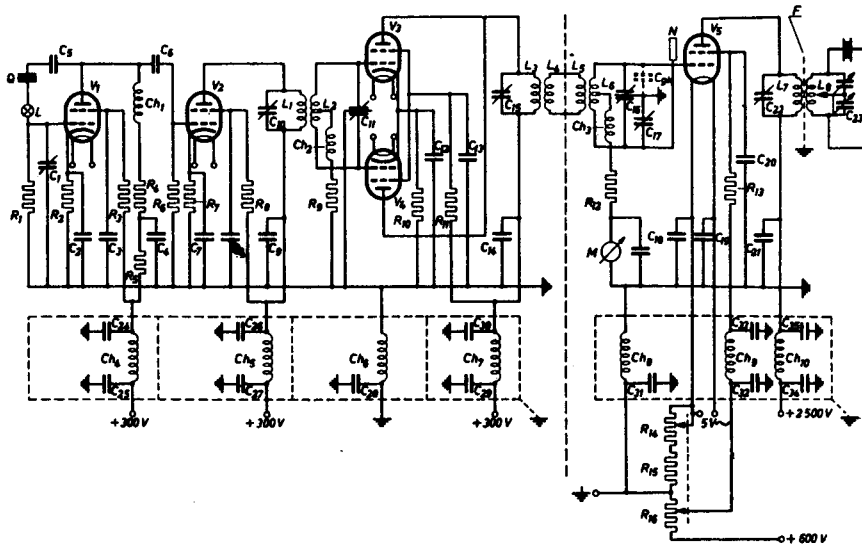


Fig. 19-10. H.F. generator with high frequency stability.

- | | | |
|--|--|---|
| $R_1 = 33 \text{ k}\Omega, 0.25 \text{ W}$ | $C_5 = 5000 \text{ pF}, 350 \text{ V}$ | $C_{28} = 1000 \text{ pF}, 350 \text{ V}$ |
| $R_2 = 1000 \Omega, 1 \text{ W}$ | $C_6 = 100 \text{ pF}, 450 \text{ V}$ | $C_{27} = 1000 \text{ pF}, 350 \text{ V}$ |
| $R_3 = 25 \text{ k}\Omega, 0.5 \text{ W}$ | $C_7 = 5000 \text{ pF}, 350 \text{ V}$ | $C_{28} = 1000 \text{ pF}, 350 \text{ V}$ |
| $R_4 = 1 \text{ k}\Omega, 4 \text{ W}$ | $C_8 = 5000 \text{ pF}, 350 \text{ V}$ | $C_{29} = 1000 \text{ pF}, 350 \text{ V}$ |
| $R_5 = 5 \text{ k}\Omega, 4 \text{ W}$ | $C_9 = 5000 \text{ pF}, 350 \text{ V}$ | $C_{30} = 1000 \text{ pF}, 350 \text{ V}$ |
| $R_6 = 0.15 \text{ M}\Omega, 0.25 \text{ W}$ | $C_{10} = 25 \text{ pF}$ | $C_{31} = 1000 \text{ pF}, 350 \text{ V}$ |
| $R_7 = 750 \Omega, 1 \text{ W}$ | $C_{11} = 2 \times 16 \text{ pF}$ | $C_{32} = 1000 \text{ pF}, 450 \text{ V}$ |
| $R_8 = 40 \text{ k}\Omega, 1 \text{ W}$ | $C_{12} = 0.01 \mu\text{F}, 350 \text{ V}$ | $C_{33} = 1000 \text{ pF}, 450 \text{ V}$ |
| $R_9 = 30 \text{ k}\Omega, 0.5 \text{ W}$ | $C_{13} = 0.01 \mu\text{F}, 350 \text{ V}$ | $C_{34} = 750 \text{ pF}, 6 \text{ kV}$ |
| $R_{10} = 200 \Omega, 4 \text{ W}$ | $C_{14} = 0.01 \mu\text{F}, 350 \text{ V}$ | $C_{35} = 750 \text{ pF}, 6 \text{ kV}$ |
| $R_{11} = 10 \text{ k}\Omega, 4 \text{ W}$ | $C_{15} = 25 \text{ pF}$ | $Cb_1 = 1.6-2.0 \text{ mH}$ |
| $R_{12} = 5 \text{ k}\Omega, 4 \text{ W}$ | $C_{16} = 2 \times 34 \text{ pF}$ | $Cb_2 = 1.6-2.0 \text{ mH}$ |
| $R_{13} = 3 \text{ k}\Omega, 12 \text{ W}$ | $C_{17} = 20 \text{ pF}$ | $Cb_3 = 70 \mu\text{H}$ |
| $R_{14} = 1 \text{ k}\Omega, 150 \text{ W}$ | $C_{18} = 4000 \text{ pF}, 300 \text{ V}$ | $Cb_4 = 1.6-2.0 \text{ mH}$ |
| $R_{15} = 500 \Omega, 75 \text{ W}$ | $C_{19} = 4000 \text{ pF}, 300 \text{ V}$ | $Cb_5 = 1.6-2.0 \text{ mH}$ |
| $R_{16} = 6 \text{ k}\Omega, 150 \text{ W}$ | $C_{20} = 4000 \text{ pF} +$ | $Cb_6 = 1.6-2.0 \text{ mH}$ |
| | $4000 \text{ pF}, 600 \text{ V}$ | $Cb_7 = 1.6-2.0 \text{ mH}$ |
| $C_1 = 25 \text{ pF var.} +$ | $C_{21} = 1250 \text{ pF}, 6 \text{ kV}$ | $Cb_8 = 1.6-2.0 \text{ mH}$ |
| 75 pF | $C_{22} = 25 \text{ pF}$ | $Cb_9 = 1.6-2.0 \text{ mH}$ |
| $C_2 = 5000 \text{ pF}, 350 \text{ V}$ | $C_{23} = 2 \times 80 \text{ pF}$ | $Cb_{10} = 70 \mu\text{H}$ |
| $C_3 = 5000 \text{ pF}, 350 \text{ V}$ | $C_{24} = 1000 \text{ pF}, 350 \text{ V}$ | $Q = \text{quartz crystal}$ |
| $C_4 = 5000 \text{ pF}, 350 \text{ V}$ | $C_{25} = 1000 \text{ pF}, 350 \text{ V}$ | 10.17 Mc/s |
-
- | | |
|--|---|
| $L_1 = 12 \text{ turns } 2.5 \text{ mm}^2, 35 \text{ mm } \varnothing$ | $L_6 = 3 \text{ turns } 2.5 \text{ mm}^2, 20 \text{ mm}$ |
| $L_2 = 27 \text{ turns } 2.5 \text{ mm}^2, 20 \text{ mm } \varnothing$ | $L_7 = 5 \text{ turns of copper tubing}$ |
| $L_3 = 8 \text{ turns } 2.5 \text{ mm}^2, 20 \text{ mm } \varnothing$ | $6 \text{ mm} \times 8 \text{ mm } \varnothing, 65 \text{ mm } \varnothing$ |
| $L_4 = 5 \text{ turns } 2.5 \text{ mm}^2, 35 \text{ mm } \varnothing$ | $L_8 = 2 \text{ turns of copper tubing}$ |
| $L_5 = 20 \text{ turns } 2.5 \text{ mm}^2, 12 \text{ mm } \varnothing$ | $6 \text{ mm} \times 8 \text{ mm } \varnothing, 65 \text{ mm } \varnothing$ |

generates a frequency of $10.17 \text{ Mc/s} \pm 0.05\%$, i.e. from 10.165 Mc/s to 10.175 Mc/s . As shown in the circuit diagram of fig. 19-10, this stage contains no tuned circuit; nevertheless the prescribed frequency limits are maintained, provided the construction is such that the temperature of the crystal is not affected by external influences. The oscillator frequency is applied to the control grid of V_2 via the coupling capacitor C_6 . This tube is operated as a frequency doubler with its anode tank circuit tuned to the second harmonic, i.e. 20.34 Mc/s . This circuit is inductively coupled to the grid circuit of the second frequency doubler stage, which is tuned to the same frequency. This special driver stage contains two tetrodes V_3 and V_4 with a push-pull input. The common anode

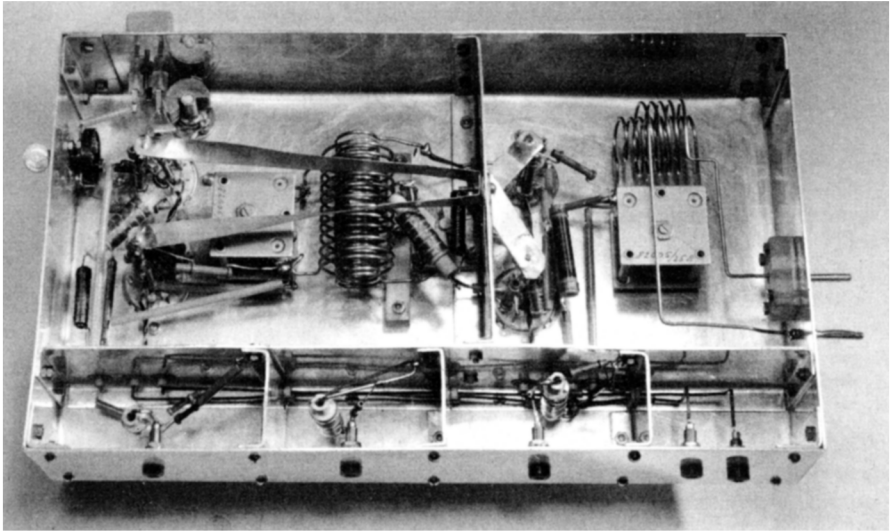


Fig. 19-11. Underside of a prototype chassis including master oscillator, frequency-doubler and driver stage.

circuit is tuned to the operating frequency of 40.68 Mc/s . It is coupled inductively to the grid circuit of tetrode V_5 which forms the output stage. The output power of this stage is controlled by adjusting both potentiometer R_{16} connected across the screen grid voltage source, and the cathode resistor R_{14} , the value of which is increased as the screen-grid voltage is decreased. The load circuit is inductively coupled to the anode tank circuit, with an earthed Faraday screen placed between the coils of the anode and load circuit in order to suppress radiation of harmonics, to eliminate capacitive reaction from the load circuit to the tuned tank circuit, and to safeguard the load circuit against coming into contact with the live coil.

At no-load tetrode V_5 may be liable to self-oscillation and it is therefore advisable to neutralize the output stage. The most practicable method is to apply grid neutralisation, which may be done by making the grid circuit symmetrical with respect to earth. The top of the grid circuit is connected to the control grid of the output tube and the bottom is connected to a copper

strip which faces the anode of the tube. The copper strip and the anode form the neutralizing capacitance, and by bending the strip correct neutralisation can easily be achieved. This method is used in the circuit of fig. 19-10. The small trimmer C_{17} serves for compensating the grid-to-cathode capacitance of V_5 indicated by C_{gk} .

To avoid undesirable feed-back between the stages these must be carefully screened from each other. Furthermore, screened filters are provided in the supply leads.

Fig. 19-11 shows an experimental chassis, seen from beneath, on which the master oscillator, the frequency-doubler and the driver stage are mounted

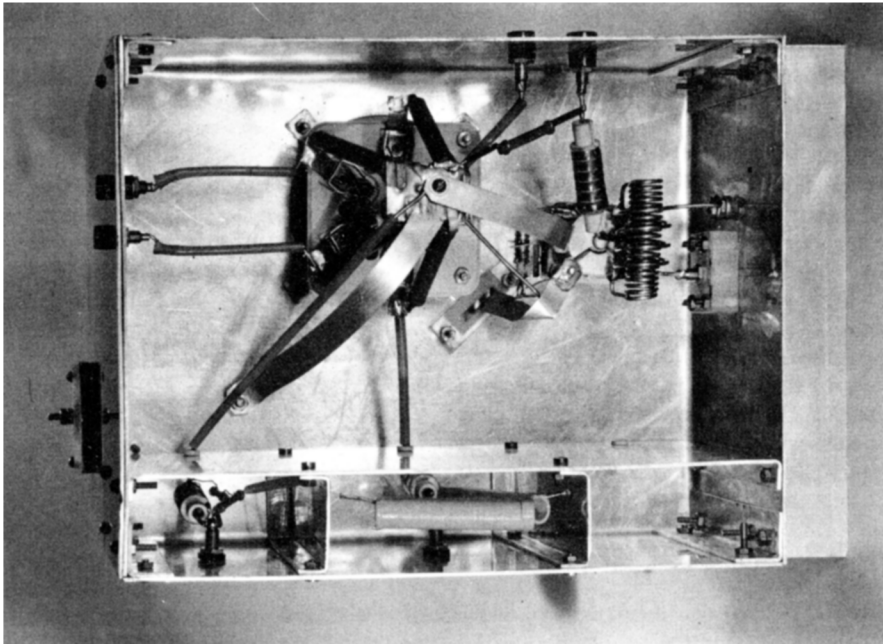


Fig. 19-12. Underside of the power stage of fig. 19-10.

(from left to right). The screened filters in the supply leads can be seen at the front.

Fig. 19-12 illustrates the output stage seen from below, clearly showing how the various components are mounted underneath the tube holder, and the earthing strips are led to a "star" point.

The master oscillator is provided with a variable feed-back capacitor C_1 which should be adjusted so that the current flowing through the crystal Q is a minimum. This can be checked easily by means of the small incandescent lamp L connected in series with the crystal.

Only two controls need be adjusted, namely one for tuning the load circuit (capacitor C_{23}) and one for adjusting the output power (ganged potentiometers R_{14} and R_{16}). The latter control may be calibrated in watts to give a rough indication of the output power at correct tuning of the load circuit. The other

trimmers and tuning capacitors are pre-set, and need be adjusted only when the tubes are replaced. There is, however, no risk of the adjustment of the set being considerably upset when tubes have to be renewed.

Automatic Load Circuit Tuning Device

During operation of a high frequency generator for capacitive heating, considerable changes of ϵ and $\tan \delta$ of the workpiece material may occur, and may cause the load circuit to fall out of resonance, thus reducing the power applied to the workpiece. This not only lowers the efficiency but, because of incorrect matching, introduces a risk of damage to the generator tube. Manual correction of the tuning will always be unsatisfactory, and does not eliminate the danger of damaging the equipment as a consequence of inattention on the part of the operator.

The stringent requirements as to frequency stability of high frequency generators imposed by the Atlantic City agreement, which necessitate the use of crystal drive on certain frequency bands, still further increase the risk that power tubes will be overloaded if the load circuit is detuned. The best solution, therefore, is to provide an automatic tuning device which acts upon the tuning of the load circuit. The cost of providing such a device is more than counterbalanced, especially in high

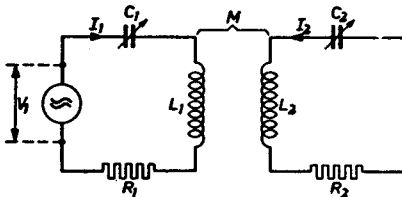


Fig. 19-13. Inductively coupled oscillating circuits.

power generators, by the increased average output, saving in time, reduced running expenses, increased life of tubes and, finally, much simpler operation, since no tuning of the load circuit is necessary.

The arrangement described below is suitable for a high frequency generator in which the load circuit is coupled inductively to the tank circuit, as in the circuit shown in fig. 19-10. The operation of the device can be followed by reference to fig. 19-13 in which are shown two oscillating circuits which are coupled inductively, the left-hand circuit being fed with a high frequency voltage V_1 . Obviously this voltage must be equal to the sum of the voltage drops appearing across the circuit elements and caused by the current I_1 , so that:

$$V_1 = I_1 \cdot R_1 + I_1 \cdot j\omega L_1 + I_1 \cdot \frac{1}{j\omega C_1} + I_2 \cdot j\omega M. \tag{19.20}$$

The last term in this equation represents the voltage which is induced in the primary circuit due to the mutual inductive coupling M between the circuits, as a consequence of the current I_2 flowing in the secondary circuit. At resonance this voltage is:

$$V_1 = I_1 \cdot R_1 + I_2 \cdot j\omega M, \tag{19.21}$$

and the voltage in the secondary circuit is:

$$V_2 = -I_1 \cdot j\omega M = I_2 \cdot R_2 + I_2 \cdot j\omega L_2 + I_2 \cdot \frac{1}{j\omega C_2}. \tag{19.22}$$

If the last term is greater than the term preceding it, the circuit is detuned capacitively; if the last term is smaller the circuit becomes inductive. If both

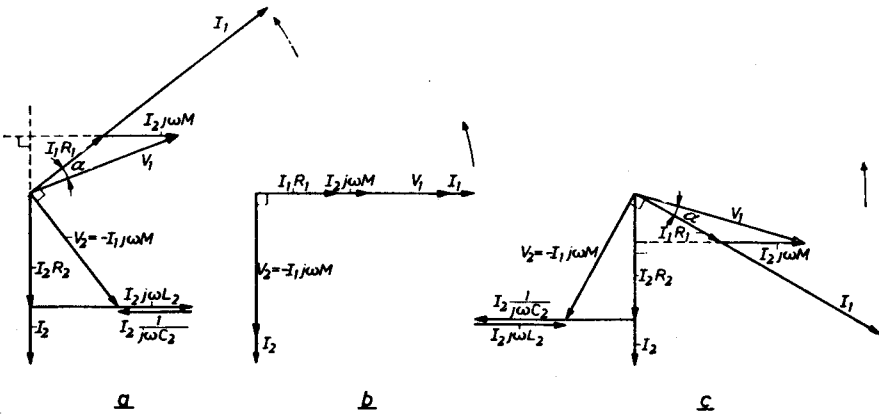


Fig. 19-14. Vector diagram of fig. 19-14, the secondary circuit being (a) inductively detuned, (b) in resonance, (c) capacitively detuned.

terms are equal the circuit is in resonance. The vector diagrams illustrating these three possibilities are given in fig. 19-14. The phase relations between the different quantities are summarised in the table below:

Phase relation between	Load		
	inductive	resistive	capacitive
I_1 and V_1	$+\alpha$	0°	$-\alpha$
I_1 and I_2	$< +180^\circ$	$+90^\circ$	$> 0^\circ$
I_1 and V_2	$+90^\circ$	$+90^\circ$	$+90^\circ$
V_1 and I_2	$< +180^\circ$	$+90^\circ$	$> 0^\circ$
V_1 and V_2	$< +90^\circ$	$+90^\circ$	$> +90^\circ$
V_2 and I_2	$< +90^\circ$	0°	$< -90^\circ$

It is seen that the phase difference between the primary current I_1 and the primary voltage V_1 varies between comparatively narrow limits which are determined by the magnitude of the counter voltage $I_2 \cdot j\omega M$ and depend, therefore, on the coupling M . The same is true for the phase difference between the voltages V_1 and V_2 . The primary current I_1 and the secondary voltage V_2 are, according to equation (19.22) always exactly in quadrature. A signal depending on any of these three phase relationships is therefore unsuitable for the purpose in view, so that use must be made of one of the three remaining phase relationships. For the control to be effective in two directions it is also

necessary that the controlling signal fluctuates around a mean value so that its polarity can be changed by means of a bridge circuit, and this stipulation precludes the use of the phase relation between V_2 and I_2 . Hence only the phase relations between I_1 and I_2 or V_1 and I_2 are suitable. In both cases the phase shift may vary almost from 180° to zero.

In principle any circuit suitable for frequency or phase detection can be used as a detecting element. The enneode EQ 80, for example, gives excellent results because of its large grid base and effective limiting action, which renders the control

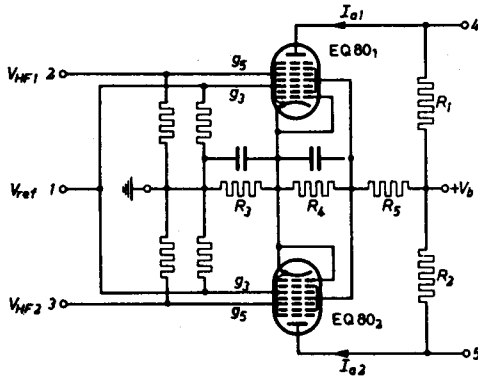


Fig. 19-15. Bridge circuit with two EQ 80 enneodes.

system practically independent of the amplitude of the signals derived from the generator. In order to obtain a control voltage of varying polarity, as is required for reversal of the control unit, a bridge circuit with two EQ 80 tubes is used, as shown in fig. 19-15.

In this circuit the tubes EQ 80₁ and EQ 80₂ and the resistors R_1 and R_2 form the four arms of a bridge which is fed by the common voltage source $+V_b$. The potentials for the screen grids g_2 , g_4 and g_6 and the negative bias of g_3 and g_5 are taken from the voltage divider consisting of

the resistors R_3 , R_4 and R_5 . The grids g_3 of both tubes are interconnected. An H.F. voltage V_{ref} of suitable magnitude and phase and taken directly from the driver or final stage of the high frequency generator, is applied to grids g_1 and serves as a reference voltage.

The control voltages V_{HF1} and V_{HF2} are applied to the grids g_5 of both tubes. These voltages have a constant phase difference of 180° . At correct tuning the voltage V_{ref} has phase differences of exactly -90° and $+90^\circ$ with respect to V_{HF1} and V_{HF2} so that the voltages at g_3 and g_5 of each tube are then in quadrature. This condition is represented in fig. 19-16a. The mean value of the resulting anode currents $I_{a1\text{ mean}}$ and $I_{a2\text{ mean}}$ are then equal, so that the bridge is in balance and the voltage between terminals 4 and 5 is zero.

When, as a result of the load circuit being detuned, V_{HF1} (i.e. the voltage applied to grid g_5 of EQ 80₁) lags less than 90° and V_{HF2} (i.e. the voltage applied to grid g_5 of EQ 80₂) leads by more than 90° with respect to the reference voltage V_{ref} , tube EQ 80₁ remains conductive for a longer interval than tube EQ 80₂. As a result the mean anode current of the first tube is larger than that of the second tube (see fig. 19-16b), and a voltage difference of a certain polarity appears across terminals 4 and 5.

If, however, the load is detuned in the other direction, it is obvious that the anode currents of the tubes will be as shown in fig. 19-16c, and the voltage across terminals 4 and 5 will be of opposite polarity.

It is this voltage between terminals 4 and 5 which is used to operate the controlling unit (see fig. 19-17), which contains two small thyatron tubes T_1 and T_2

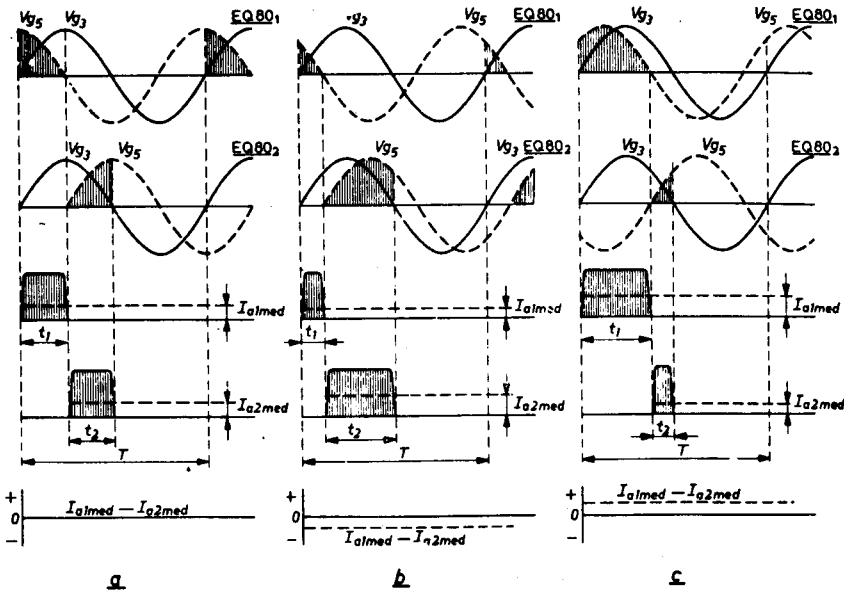


Fig. 19-16. Anode currents of the EQ 80 tubes in fig. 19-15, the load circuit being in resonance (a) and detuned (b and c).

connected in push-pull. Instead of the customary d.c. shunt-wound motor with the armature fed by two inverse-parallel connected thyratrons, a series motor having two field windings is used in this circuit.

According to the polarity of the terminals 4 and 5, this motor is fed from the a.c. mains either via tube T_1 and field winding F_1 , or via tube T_2 and field winding F_2 , thus being made to run in one direction or the other.

Since the grids of the thyratrons are directly connected to the anodes of the preceding EQ 80 tubes, it is necessary to apply a fixed negative grid bias to the thyratrons. This should have such a value that the grids remain sufficiently negative with respect to the cathodes of the thyratrons so that the tubes will not ignite at correct tuning, i.e. when the bridge circuit is balanced. The bias can be adjusted by means of potentiometer R_5 . To avoid simultaneous ignition of both thyratrons a resistor R_6 is also included in the common cathode lead. As soon as one of the thyratrons ignites an additional voltage drop is produced across this resistor, increasing the negative bias of the other tube.

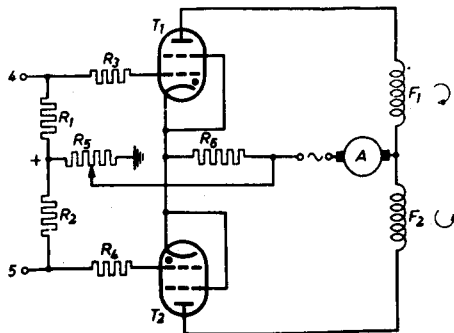


Fig. 19-17. Thyatron control circuit.

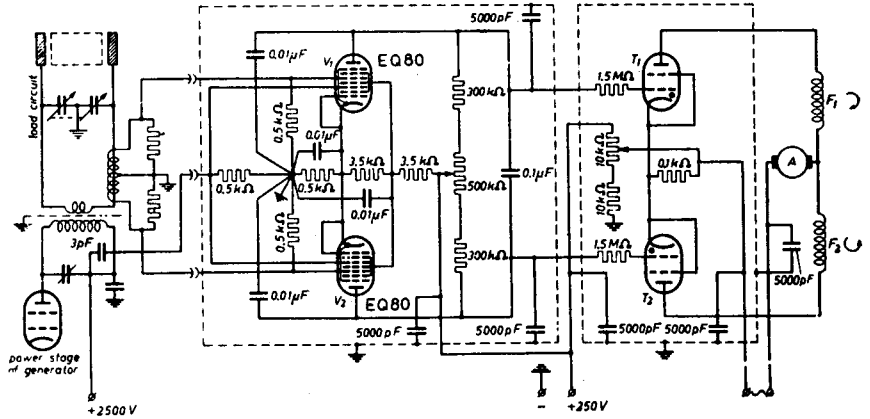


Fig. 19-18. Complete circuit of an automatic load circuit tuning device.

Fig. 19-18 shows the complete circuit of an automatic tuning device based upon the simplified circuits already described. The residual voltage across the smoothing capacitor in the d.c. supply lead of the generator tube is used as the reference voltage V_{ref} ; its phase corresponds to that of the voltage V_1 in the

anode tank circuit. The reference voltage is fed to the grids g_3 of the two EQ 80 tubes by means of a concentric line, via a blocking capacitor of 3 pF. The phase of the control voltage for the grids g_3 of the EQ 80 tubes should correspond to that of the current I_2 flowing in the load circuit. This voltage is obtained by means of a bar-type current transformer surrounding the lead to one of the capacitor plates in the load circuit. The ends of the winding of this current

transformer are connected to the grids g_3 of the EQ 80 tubes, its centre tap being connected to earth.

Zero adjustment at correct tuning is achieved by means of the 500 kΩ potentiometer.

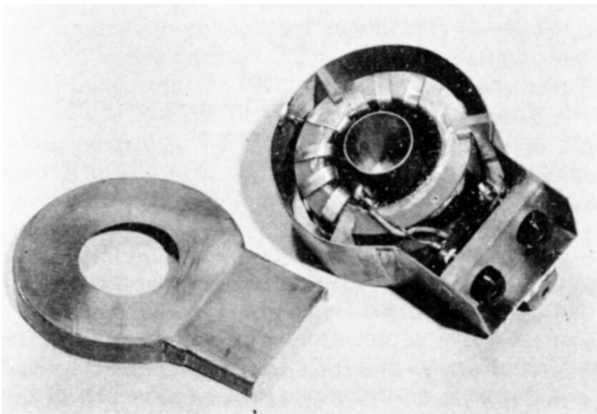
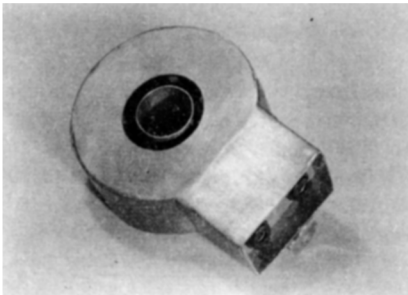


Fig. 19-19. Current transformer used to obtain control voltages from the load circuit.

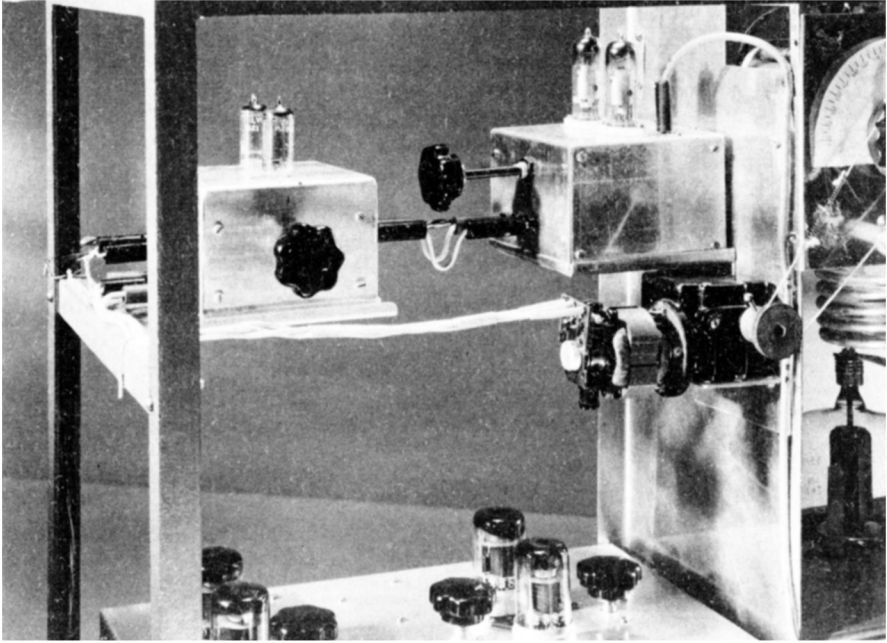


Fig. 19-20. Complete prototype automatic tuning device.

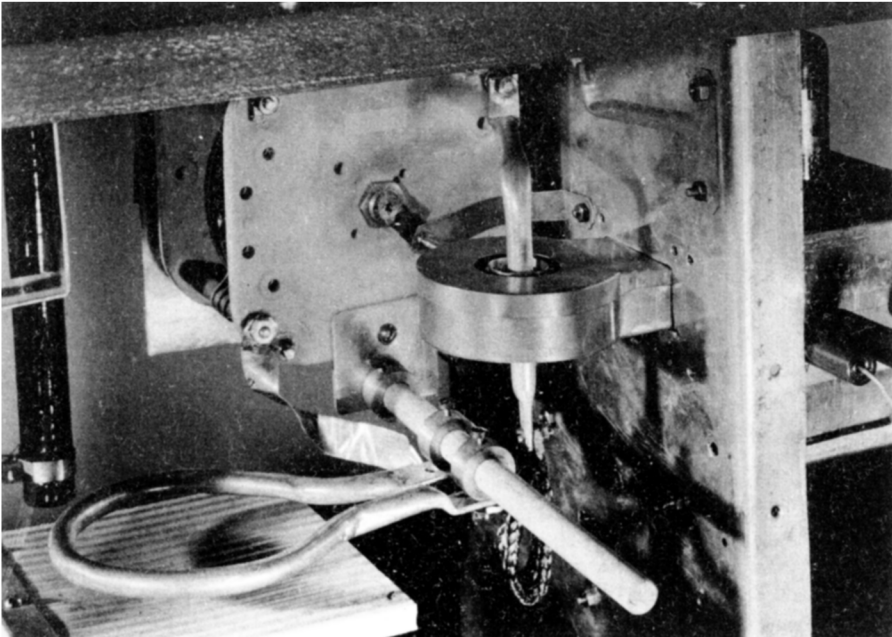


Fig. 19-21. Photograph of the load circuit with coupling coil and current transformer.

meter in the anode circuit of the EQ 80 tubes, which is adjusted so that the voltages at the control grids of the two PL2D21 thyratrons are exactly equal at correct tuning. The sensitivity of the tuning controller can be varied over a wide range by means of the $10\text{ k}\Omega$ potentiometer in the common grid circuit of the thyratrons.

A d.c. series motor with a double worm gear is used for tuning correction of the load circuit. The motor is fed from the a.c. mains via one or other of the thyratrons, and will run in one direction or the other according to which thyatron is conducting, thus adjusting the variable capacitor in the load circuit until correct tuning has been re-established.

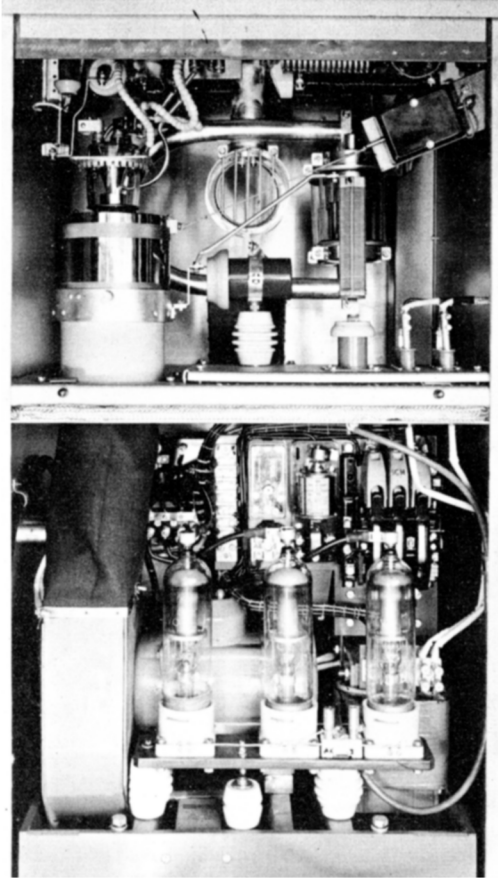


Fig. 19-22. Internal construction of a 6 kW generator.

Fig. 19-19 illustrates the current transformer in detail. Apart from a narrow annular gap, which is necessary in order to avoid excessive eddy currents, the secondary coil is completely surrounded by a copper screen.

Fig. 19-20 shows a complete prototype equipment. On the right can be seen the small motor with gear and transmission to the tuning capacitor, and above it the chassis equipped with two EQ 80 tubes. On the left is the controlling unit containing the thyratrons.

Fig. 19-21 shows the other side of the apparatus, the coupling coil of the load circuit and the current transformer being clearly visible.

Whatever method of load matching is adopted, it is always advisable to take due precautions against overloading

the generator tube, more particularly since the apparatus will often be operated by unskilled personnel who have little or no technical knowledge. As already mentioned in the preceding chapter, incorrect matching introduces a risk of overloading the anode or the grid of the generator tube. Limitation of the power dissipated at the anode is usually a simple matter, especially in the case of a tube having an external anode. The temperature can be automatically and continuously measured, and if a maximum value is exceeded an alarm signal can be given or the anode circuit of the tube can be interrupted. In many cases, however, even

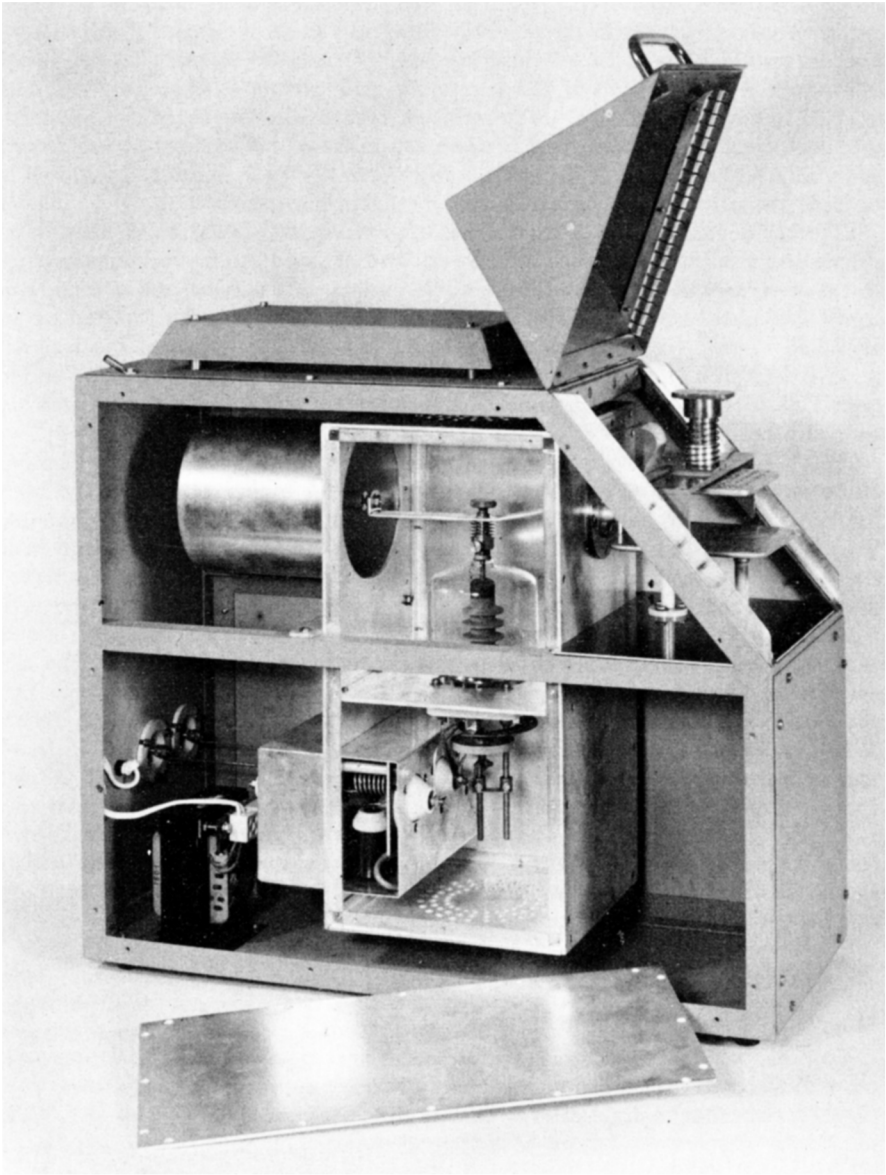


Fig. 19-23. Internal construction of a 500 W generator operating at a frequency of 75 Mc/s.

these measures are superfluous, since the maximum anode dissipation of the tubes employed is usually so great that it is impossible to exceed it in practice.

Limitation of the power dissipated at the grid can be achieved in many ways. For instance, a relay with overload trip can be included in the grid circuit to interrupt the supply leads if the maximum grid current is exceeded. Another method frequently used consists in inserting resistors having a positive temperature coefficient of resistance (for example incandescent lamps) in the grid circuit. Such an arrangement is used in model (*b*) — see fig. 19-9. Finally, it is possible to limit the grid current automatically by electronic means.

Like all devices for industrial duty, high frequency generators should be simple and straight-forward in their layout and of good mechanical construction so that any necessary repairs can be easily and quickly carried out at minimum cost. The tubes should be so mounted that replacements can be inserted by an unskilled operator without risk of damage. Fig. 19-22 illustrates the internal arrangement of a 6 kW generator equipped with a fan-cooled transmitting triode type TBL 6/6000 (see also fig. 1-17), which is very accessible and can therefore be easily replaced.

Fig. 19-23 shows the internal view of a high frequency generator having an output power of 500 watts at a frequency of 75 Mc/s. In view of the high frequency involved, part of the anode circuit is designed as a co-axial resonator. In this way a circuit with a remarkably high Q -factor is obtained, and little energy is lost by radiation. The capacitor of the load circuit is connected directly to the anode circuit, the anode of the generator tube being earthed.

Because in every generator a certain amount of energy is lost as heat, ventilation slots or louvres must be provided in the generator housing to permit free circulation of cooling air, particularly when forced air cooling is employed. The generator should therefore be located in a position where movement of the air is not impeded by other plant or equipment. It should also be borne in mind that deposits of dust inside the generator may lead to leakage or short circuits and the apparatus should therefore be cleaned or blown out at regular intervals.

Finally, in order to ensure consistent operation and freedom from breakdown the instructions issued by the equipment manufacturer and the tube maker should be carefully studied so that the small amount of maintenance required can be correctly and expeditiously carried out.

20. ELECTRONIC APPARATUS FOR SPECIAL PURPOSES

To conclude this description of electronic apparatus for industrial purposes, we will discuss in this chapter various electronic apparatus which, on account of their nature and application, do not come within the scope of any of the preceding chapters.

Electronic Stabilisation of Direct and Alternating Voltages

For feeding apparatus it is often necessary to have available direct or alternating voltages the values of which are substantially independent of the load or of mains voltage fluctuations. In such cases electronic stabilisation circuits can be used to advantage. This will be illustrated by means of the following two examples which have proved to give most satisfactory results in practice.

Fig. 20-1 shows the circuit of a stabilised supply unit which can be used for feeding small transmitters, measuring equipment in laboratories and for similar purposes. A voltage of 2×450 V is taken from the supply transformer and applied to two mercury-vapour rectifying tubes DCG 4/1000. The rectified current of 325 mA is smoothed by the filter L, C_1 and flows through the load via four tubes EL 34, connected in parallel, and a milliammeter. By means of the adjustable voltage divider R_{12}, R_{13}, R_{14} , part of the output voltage is applied to the control grid of the special quality, high-slope pentode E 83 F; a fixed voltage, which is determined by the voltage reference tube 85 A1, is applied to the cathode of this pentode. The resistors R_1 and R_2 prevent the current

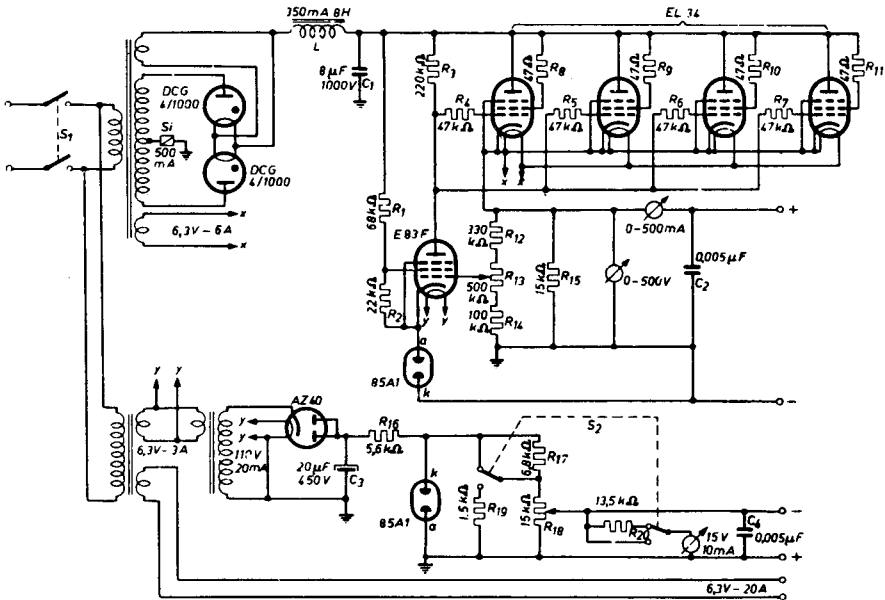


Fig. 20-1. Circuit of a stabilised supply unit.

flowing through the voltage reference tube from dropping below its minimum value. The anode of the E 83 F is connected directly to the control grid of the regulating tubes EL 34 by means of the resistors R_4, R_5, R_6 and R_7 . The output is permanently shunted by a resistor R_{15} of 15 k Ω ; the capacitor C_2 forms a short circuit for any H.F. voltages which may occur in the load circuit.

At an output voltage up to 250 V the output current may reach 325 mA. At an output current varying from 0 to 325 mA the variations of the output voltage are so small that they cannot be discerned on the measuring instrument. At higher output voltages the maximum permissible current becomes smaller (250 mA at 300 V, 200 mA at 350 V and 125 mA at 400 V); the variations of the output voltage may then reach up to 2 V, that is 0.5%, when the current is increased from zero to its maximum value.

The output voltage variations caused by mains voltage fluctuations of $\pm 20\%$ proved to be less than $\pm 1\%$. The maximum effective internal resistance of the supply unit is 10 Ω . The ripple on the output voltage is approximately 50 mV

at a current of 100 mA, 80 mV at a current of 200 mA, and 120 mV at a current of 300 mA. The ripple voltage may obviously be further reduced by improving the smoothing filter. The output voltage can be adjusted by means of the potentiometer R_{13} to any value ranging from 150 V to 400 V.

A grid bias can be taken from a separate rectifier that is also stabilised by means of a voltage reference tube 85 A1. This voltage can be adjusted by means of the potentiometer R_{18} to any value between 0—15 V or 0—85 V according to the position of switch S_2 , which also selects the correct range of the voltmeter.

Fig. 20-2 shows the circuit of a supply unit for a stabilized alternating voltage. The alternating voltage from the transformer Tr_1 flows through the two saturable-

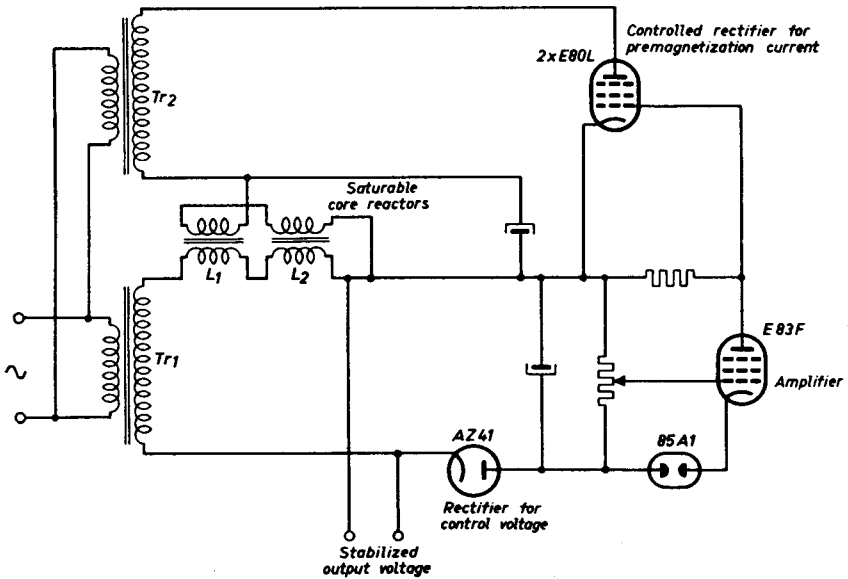


Fig. 20-2. Basic circuit for stabilising an alternating current.

core reactors L_1 and L_2 and the load. The rectifying tube AZ 41 converts the alternating output voltage into a direct voltage which is almost proportional to the amplitude of this output voltage. Variations of the rectified voltage are amplified by the tube E 83 F, the cathode of which is given a fixed potential of 85 V by the voltage reference tube 85 A1. The E 83 F supplies a variable control grid voltage for the two tubes E 80 L which operate as controlled rectifiers and supply the premagnetization current for the reactors L_1 and L_2 .

It will be assumed that at a certain instant the output voltage increases: the direct voltage supplied by the AZ 41 then also increases, and so does the anode current of the E 83 F. As a result, the control grids of the two tubes E 80 L are made more negative so that the premagnetization current flowing through the reactors decreases, and their inductance increases. In this way an increase of the alternating output voltage is counteracted.

Fig. 20-3 shows the complete circuit of this supply unit, which was designed for an output voltage of 110 V. The transformer Tr_1 is provided with tappings

for 125 V, 145 V and 160 V. When the tapping for 125 V is selected an output power of up to 50 W is available. The output voltage is then kept constant at $110\text{ V} \pm 1\%$, the mains voltage then being allowed to fluctuate between 192 V and 228 V. For an output power ranging from 50 W to 100 W the 145 V tapping must be used; for mains voltage fluctuations from 182 V to 236 V the output voltage is then also $110\text{ V} \pm 1\%$. The 160 V tapping is intended for an output power ranging from 100 W to 200 W; the output voltage then remains constant at $110\text{ V} \pm 1\%$ notwithstanding mains voltage fluctuations between 172 V and 240 V. The stabilisation can even be improved by feeding the transformer Tr_2 from a stabilised voltage source instead of from the mains.

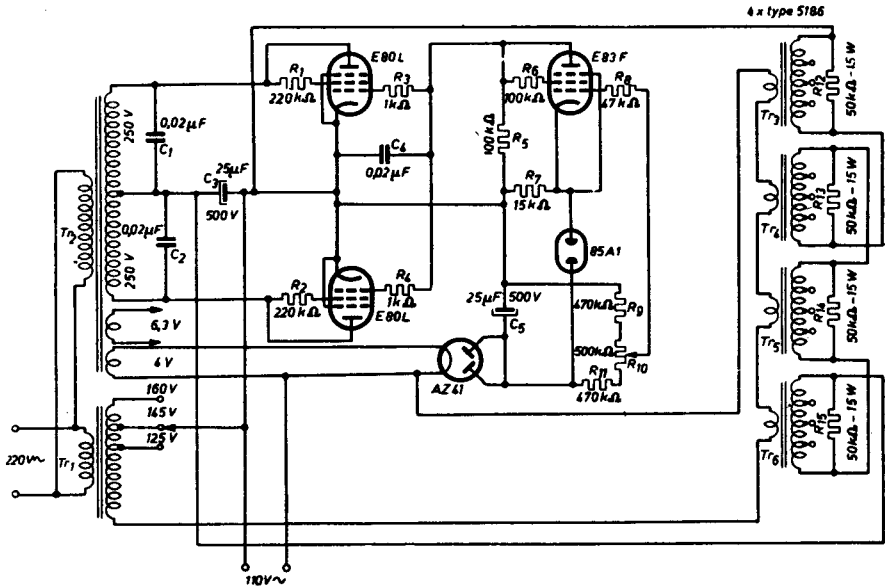


Fig. 20-3. Complete circuit of a unit for supplying a stabilised alternating voltage.

The secondary 7Ω windings of two push-pull radio output transformers (Philips Type No. 5186) connected in series are used for each of the saturable-core reactors; the premagnetization current flows through the transformer primaries. These are connected in series in such a way that the current flows alternately in opposite directions and the alternating voltages induced in these windings cancel each other. The resistors R_{12} to R_{15} shunted across the primaries have been provided to improve the stability of the circuit and to ensure that the waveform of the output voltage is as sinusoidal as possible.

Damping resistors are included in the control and screen-grid circuits of the tubes to prevent H.F. oscillations from being produced. The capacitor C_4 in the control-grid circuit of the tubes E 80 L serves to suppress low-frequency output voltage fluctuations.

Ultrasonic Soldering Iron

Another example of the use of electronics in the treatment of metal is the soldering of aluminium under conditions of ultrasonic vibration. As is well

known, when light metals such as aluminium are exposed to air, a thin layer of inert oxides is almost instantaneously formed on their surface and renders tinning by means of a conventional soldering iron practically impossible. If, however, the molten solder applied to the surface of the metal is subjected to mechanical vibrations having a frequency in the order of 20 kc/s, the phenomenon known as "cavitation" occurs, that is to say vapour-filled cavities are continuously formed and collapse violently, thus breaking up the layer of oxide. No further oxygen can reach the metal because the surface is covered by the molten solder, so that the aluminium cannot re-oxidise and its surface is efficiently "wetted" by the solder — in other words, the aluminium surface is tinned.

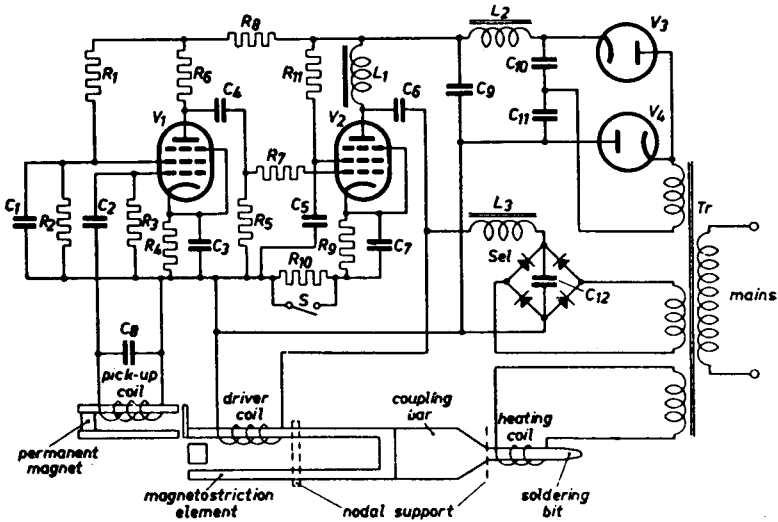


Fig. 20-4. Diagram of an ultrasonic soldering iron.

An ultrasonic soldering iron with its ancillary apparatus is illustrated schematically in fig. 20-4. The body of the iron contains a transducer consisting of a magnetostriction element operated at half-wavelength resonance. At the centre of this U-shaped element, which is composed of cobalt alloy laminations, appears a vibration node, and the element is supported at this point so that the two ends are allowed to vibrate longitudinally. On one leg of the element is mounted the driver coil, which is energised by the output of a two-stage tube oscillator comprising driver tube V_1 and class-C operated power tube V_2 . In order to ensure that the element always vibrates at its resonant frequency, the first stage of the oscillator is excited not by a separate crystal oscillator but by an alternating voltage induced in the coil of a small pick-up mounted at one end of the magnetostriction element. The system is thus self-driving. The other end of the unit is connected to a half-wavelength coupling bar which in turn is connected to the electrically-heated soldering bit.

The magnetic circuit of the U-shaped transducer is completed by a small block of magnetic alloy mounted between the ends of the two limbs. One limb also carries a small piece of alloy which faces the poles of the pick-up. The air-gap can be accurately adjusted. The value of capacitor C_8 is so chosen that, together

with the pick-up coil, it forms an oscillatory circuit tuned to the natural frequency of the transducer. Thus, stable operation at this frequency is assured and excitation at harmonic frequencies is avoided.

When switch S is opened the negative grid bias of tube V_2 is greatly increased and the tube is cut off. Thus the power for the driver coil can be controlled. Selenium rectifier Sel produces a voltage in the order of 10 or 12 volts for polarising the magnetostriction element. Another winding on mains transformer Tr supplies a low voltage current for the heating element of the soldering bit which consists of a coil of resistance wire. The high voltage supply unit for V_1 and V_2 consists of a voltage-doubler rectifying circuit employing tubes V_3 and V_4 .

Photo-Electric Revolution Counter

With most of the customary revolution counters it is impossible to obtain an accurate result without loading to a certain extent the device that is to be measured. This may give rise to errors, particularly when measuring high speed,

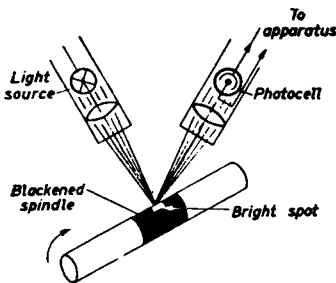


Fig. 20-5. Optical measurement of rotational speeds by discontinuous reflection.

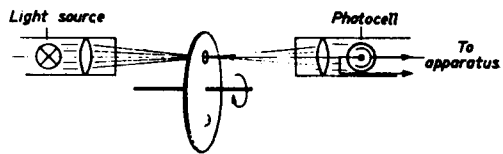


Fig. 20-6. Optical measurement of rotational speeds by interruption of a beam of light.

owing to the additional friction-losses and resonance phenomena caused by the coupling elements. These disadvantages are avoided by the photo-electric revolution counter described below, the device to be measured being scanned by a beam of light. A further advantage of this method is the possibility of counting revolutions at places which are not easily accessible, for example those of spindles in internal grinding and spinning machines and in high-speed centrifugal machines, and so forth.

Internal grinding machines may have a speed of the order of 120 000 r.p.m. The measuring equipment described below was therefore designed for measuring speeds ranging from 1000 to 180 000 r.p.m. (corresponding to 3000 c/s).

The measurement is carried out by mounting a photocell and a light source so that the object to be measured produces variations in the intensity of the light beam impinging on the photocell. This can, for example, be achieved by discontinuous reflection from a spindle or a flywheel, by regular interception of the light beam or by some such method, as shown diagrammatically in figs 20-5 and 20-6. Such a variation of the light beam results in the production of a current pulse by the photocell.

As shown by the circuit of Fig. 20-7, this current pulse results in a voltage pulse being produced across the anode resistor R_1 . This voltage pulse is am-

plified by the tube V_1 (EF 40), and subsequently shaped by V_2 (EF 40) in such a way that, regardless of the duration of the initial current pulse, a sharp voltage pulse is obtained.

The pulses are counted by means of the gas-filled triode V_3 (AC 50) in combination with an RC -network and a moving-coil meter. When the AC 50 is ignited by a pulse, the capacitor C_{17-21} will be discharged fairly rapidly across the resistor R_{18} ; during the intervals between the pulses this capacitor is slowly recharged via the resistor R_{21} . The arithmetical average value of the charging current is proportional to the repetition frequency of the pulses, that is to the speed, so that the instrument gives a direct reading of the frequency.

The RC -combination $R_{21}-C_{17-21}$ should be so dimensioned that the capacitor switched-on is almost fully charged during the intervals. The switch C_1 has therefore been provided with 5 positions, so that a next range can be selected when the reading of the instrument becomes too small.

Since the instrument is controlled by voltage pulses, it can be used not only for measuring rotational speeds, but also for determining the frequency of alternating voltages or of pulsating direct voltages. The instrument has therefore been provided with two input terminals. The input A_1 is intended for the photo-cell; the anode voltage required for this cell is taken from the voltage divider R_3R_4 . For determining the frequency of a voltage, the input A_2 is used; the form of the pulses applied is not critical in this case. Here too, the upper limit of the measuring range is 3000 c/s.

COMPONENT VALUES OF FIG. 20-7.

$R_1 = 1 \text{ M}\Omega, 0.5 \text{ W}$	$R_{27} = 0.1 \text{ M}\Omega, 0.5 \text{ W}$	$C_{12} = 1000 \text{ pF}, 250/750 \text{ V}$
$R_2 = 2 \text{ M}\Omega, 0.5 \text{ W}$	$R_{28} = 0.1 \text{ M}\Omega, 0.5 \text{ W}$	$C_{13} = 400 \text{ pF}, 250/750 \text{ V}$
$R_3 = 0.1 \text{ M}\Omega, 0.5 \text{ W}$	$R_{29} = 2 \text{ k}\Omega, 0.5 \text{ W}$	$C_{14} = 50 \text{ pF}, 250/750 \text{ V}$
$R_4 = 0.2 \text{ M}\Omega, 0.5 \text{ W}$	$R_{30} = 40 \text{ k}\Omega, 0.5 \text{ W}$	$C_{15} = 20 \text{ pF}, 250/750 \text{ V}$
$R_5 = 1 \text{ k}\Omega, 0.5 \text{ W}$	$R_{31} = 0.1 \text{ M}\Omega, 0.5 \text{ W}$	$C_{16} = 50 \text{ }\mu\text{F}, 12/150 \text{ V}$
$R_6 = 35 \text{ k}\Omega, 0.5 \text{ W}$	$R_{32} = 10 \text{ k}\Omega, 0.5 \text{ W}$	$C_{17} = 101000 \text{ pF}$
$R_7 = 0.5 \text{ M}\Omega, 0.5 \text{ W}$	$R_{33} = 1.2 \text{ M}\Omega, 0.5 \text{ W}$	$C_{18} = 31400 \text{ pF}$
$R_8 = 0.1 \text{ M}\Omega, 0.5 \text{ W}$	$R_{34} = 0.5 \text{ M}\Omega$	$C_{19} = 10000 \text{ pF}$
$R_9 = 10 \text{ k}\Omega, 0.5 \text{ W}$	$R_{35} = 10 \text{ k}\Omega, 0.5 \text{ W}$	$C_{20} = 3470 \text{ pF}$
$R_{10} = 0.5 \text{ M}\Omega$	$R_{37} = 100 \text{ }\Omega, 2 \text{ W}$	$C_{21} = 950 \text{ pF}$
$R_{11} = 1 \text{ k}\Omega, 0.5 \text{ W}$	$R_{38} = 8 \text{ k}\Omega, 12 \text{ W}$	$C_{22} = 25000 \text{ pF}$
$R_{12} = 35 \text{ k}\Omega, 0.5 \text{ W}$	$R_{39} = 0.5 \text{ M}\Omega, 0.5 \text{ W}$	$C_{23} = 41000 \text{ pF}$
$R_{13} = 0.5 \text{ M}\Omega, 0.5 \text{ W}$	$R_{40} = 0.5 \text{ M}\Omega, 0.5 \text{ W}$	$C_{24} = 10000 \text{ pF}$
$R_{14} = 0.1 \text{ M}\Omega, 0.5 \text{ W}$	$R_{41} = 0.5 \text{ M}\Omega, 0.5 \text{ W}$	$C_{25} = 8 \text{ }\mu\text{F}, 6/9 \text{ V}$
$R_{15} = 10 \text{ k}\Omega, 0.5 \text{ W}$		$C_{26} = 0.25 \text{ }\mu\text{F}, 250/750 \text{ V}$
$R_{16} = 0.1 \text{ M}\Omega, 0.5 \text{ W}$	$C_1 = 0.1 \text{ }\mu\text{F}, 250/750 \text{ V}$	$C_{27} = 500 \text{ pF}, 250/750 \text{ V}$
$R_{17} = 6 \text{ k}\Omega, 0.5 \text{ W}$	$C_2 = 0.1 \text{ }\mu\text{F}, 250/750 \text{ V}$	$C_{28} = 1000 \text{ pF}, 250/750 \text{ V}$
$R_{18} = 1 \text{ k}\Omega, 0.5 \text{ W}$	$C_3 = 2.0 \text{ }\mu\text{F}, 250/750 \text{ V}$	$C_{29} = 0.1 \text{ }\mu\text{F}$
$R_{19} = 30 \text{ k}\Omega, 0.5 \text{ W}$	$C_4 = 50 \text{ }\mu\text{F}, 6/9 \text{ V}$	$C_{30} = 4 \text{ }\mu\text{F}$
$R_{20} = 5 \text{ k}\Omega, 0.5 \text{ W}$	$C_5 = 0.5 \text{ }\mu\text{F}, 250/750 \text{ V}$	$C_{31} = 0.5 \text{ }\mu\text{F}, 500/1500 \text{ V}$
$R_{21} = 67 \text{ k}\Omega, 1 \text{ W}$	$C_6 = 2 \text{ }\mu\text{F}, 250/750 \text{ V}$	$C_{32} = 4 \text{ }\mu\text{F}$
$R_{22} = 10 \text{ k}\Omega$	$C_7 = 0.05 \text{ }\mu\text{F}, 250/750 \text{ V}$	$C_{33} = 1000 \text{ }\mu\text{F}$
$R_{23} = 1 \text{ M}\Omega, 0.5 \text{ W}$	$C_8 = 50 \text{ }\mu\text{F}, 6/9 \text{ V}$	$C_{35} = 16 \text{ }\mu\text{F}, 450/550 \text{ V}$
$R_{24} = 0.1 \text{ M}\Omega, 0.5 \text{ W}$	$C_9 = 0.5 \text{ }\mu\text{F}, 250/750 \text{ V}$	$C_{36} = 16 \text{ }\mu\text{F}, 450/550 \text{ V}$
$R_{25} = 10 \text{ k}\Omega, 0.5 \text{ W}$	$C_{10} = 2.0 \text{ }\mu\text{F}, 250/750 \text{ V}$	$C_{37} = 16 \text{ }\mu\text{F}, 450/550 \text{ V}$
$R_{26} = 150 \text{ k}\Omega, 0.5 \text{ W}$	$C_{11} = 2000 \text{ pF}, 250/750 \text{ V}$	$C_{38} = 2 \times 500 \text{ }\mu\text{F}, 3/9 \text{ V}$

Transformer Tr_1 :

primary: 3300 turns, 0.1 mm \varnothing ;
secondary: 1640 turns, 0.1 mm \varnothing
tapped at 1000 turns.

Transformers Tr_2 and Tr_3 :

primary: 200 turns 0.2 mm \varnothing ;
secondary: 200 turns, 0.2 mm \varnothing .

In view of the fact that too large or too small pulses may result in misfiring of the AC 50 and counting errors, the first amplifying tube is followed by a volume control. The resistor R_{20} included in the cathode circuit of V_3 prevents this tube from being ignited twice by one pulse; during the interval in which the capacitor is charged, this resistor ensures that an additional negative grid bias is applied to the tube.

Whereas it is possible on the one hand to measure frequencies with this apparatus, it is also possible to calibrate the apparatus with known frequencies; the accuracy of the calibration is then determined exclusively by the constancy of the calibration frequency and by the accuracy of the measuring instrument.

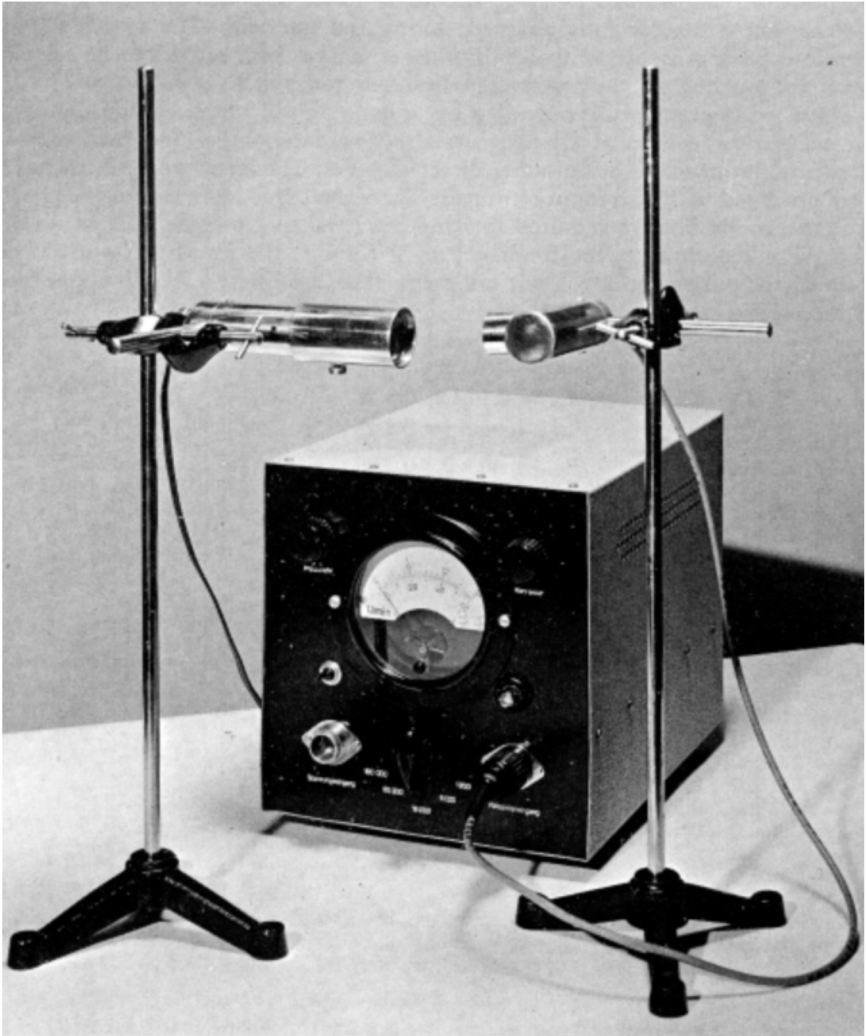


Fig. 20-8. Laboratory model of an apparatus according to the circuit of fig. 20-7.

The apparatus described above was designed for measuring purposes in a laboratory and was equipped with a very accurate calibration oscillator. For other purposes it might be possible to simplify the circuit to some extent. It would, for example, be possible to omit the calibration device, and to calibrate the amplifier by applying a voltage of known frequency (for example the mains frequency) to the input A_2 .

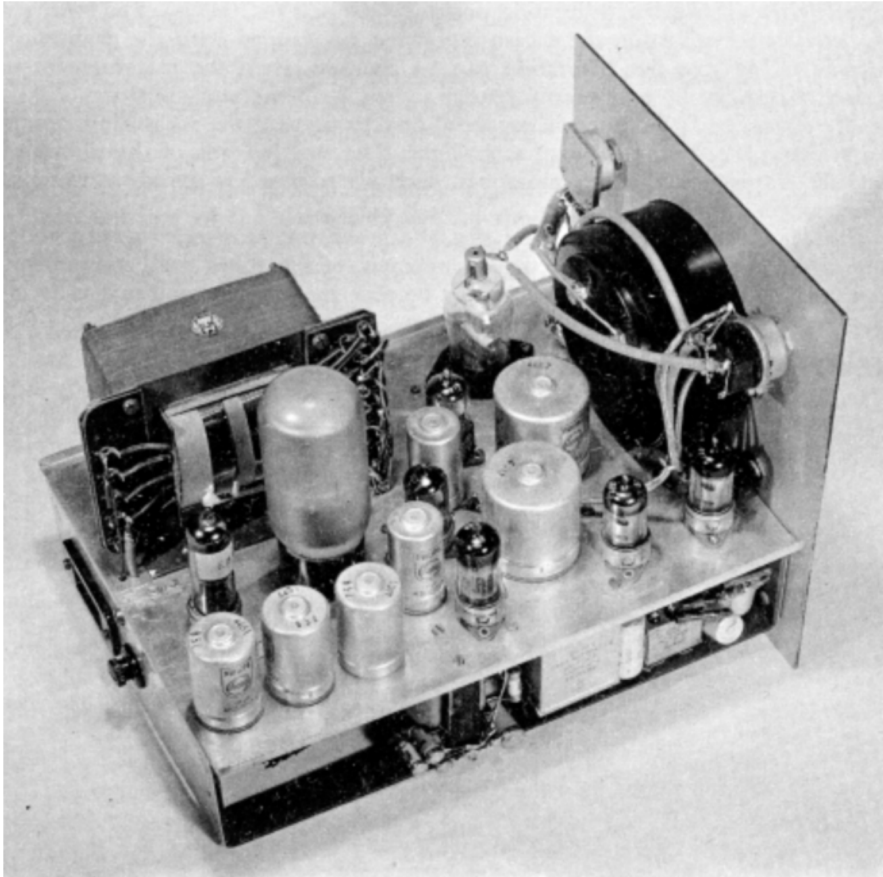


Fig. 20-9. Chassis of the apparatus according to the circuit of fig. 20-7.

The calibration oscillator is equipped with the tubes V_4 , V_5 and V_6 . The left triode section of V_4 (ECC 40) operates as a 1000 c/s oscillator; the right section, with the RC -network included in its anode circuit, converts the oscillations into pulses in a similar way as is done in the measuring apparatus. These pulses are applied to the calibration switch S_{1c} via the capacitor C_{27} , and also to the control grid of the tube V_5 (EF 42), which is connected as a blocking oscillator. The synchronisation is adjusted by means of the resistor R_{28} so that oscillations with a frequency of 200 c/s are produced by this tube. These oscillations are used

the thyratrons T_1 and T_2 . As soon as the latter have reached their operating temperature, the contact of the bi-metal relay Rel_1 will be closed so that the direct current will flow from the positive terminal of the d.c. mains via the choke L to the centre tapping of the primary of transformer Tr . The two extremities of the transformer primary are connected to the negative terminal of the d.c. mains via thyratrons. To obtain an alternating current the direct current must be made to flow alternately through the left and the right half of the primary, so that alternating voltages are induced in the secondary windings. This alternate commutation is achieved by controlling the grids of the thyratrons. As soon as the one thyatron is ignited, the other thyatron is extinguished by means of the commutating capacitor C_1 , which is connected between the two anodes and the operation of which corresponds to that of the capacitor C_3 in the circuit of fig. 12-8. The corresponding thyatron is extinguished in this case due to the effective anode voltage temporarily dropping below the arc voltage.

The frequency of the alternating current thus generated depends on that of the alternating grid voltage, which — owing to the coupling between the grid and anode circuit — depends in turn on the frequency in the anode circuit. For this reason the transformer Tr has been provided with an additional winding w_{s2} , which forms a phase-shifting network in combination with C_2 and R_4 . To ensure self-excitation, the phase relation between the anode voltage and the grid voltage should have a definite value; consequently, the frequency is determined by the time constant of the RC -network. When the resistance is modified, the frequency will also vary. The waveform of the generated alternating voltage

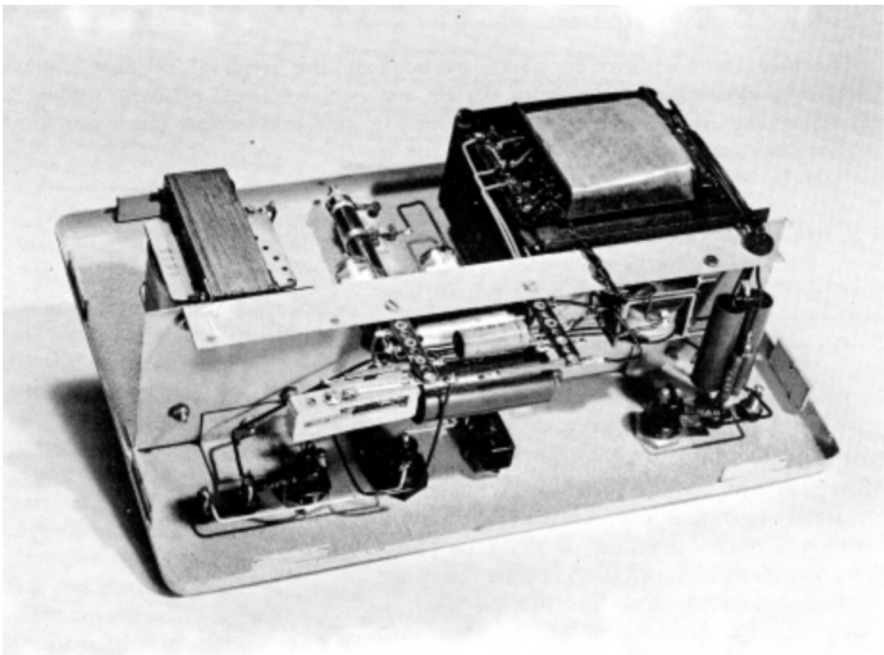


Fig. 20-11. Inverter according to the circuit of fig. 20-10.

and the efficiency are, on the other hand, determined exclusively by the component values in the anode circuit, mainly by the value of the commutating capacitor.

To obtain a high efficiency the losses in the transformer and choke (IR -drop and iron losses) should be kept as small as possible. The efficiency is moreover reduced by the voltage drop (arc voltage) in the thyratrons.

As soon as the inverter starts to operate, and a voltage is induced in the secondary winding w_{s1} , a voltage will also be induced in the winding w_{s3} , so that the relay Rel_2 is energised. The filament circuit of the thyratrons is then disconnected from the d.c. mains, and the filament current supply is taken over by the winding w_{s3} . In this way no energy is wasted in the dropping resistor R_2 .

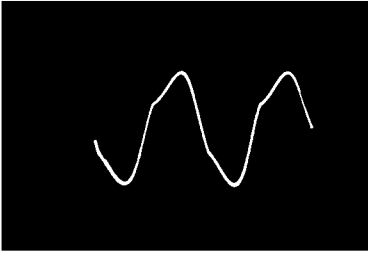


Fig. 20-12. Oscillogram of the output voltage.

When two thyratrons PL 2D 21 are used, the output power will be approximately 40 W at an efficiency of approximately 75%. Roughly 10 W are lost in the filament supply and the excitation of the relays, so that a useful power of approximately 30 W is available.

Fig. 20-11 shows the construction of such an inverter and fig. 20-12 an oscillogram of the output voltage at 50 c/s.

Electronic Dust Precipitators

The problem of purifying gases, particularly the removal of dust, fly-ash and similar floating particles from the air, has become a real problem owing to the increasing industrialisation and traffic. In this connection electronic dust precipitators deserve special attention, for it is only by means of these devices that the smallest particles, with a diameter of less than 1μ , can be captured. The electrostatic filter used for this purpose can, moreover, also precipitate fluid particles, such as tar, oil and vaporised acids.

The principle of the electrostatic dust precipitator was made known 35 years ago, when this device was invented by the American Cottrell. The same principle is still applied and is based on the schematical representation shown in fig. 20-13. Two metal plates are mounted opposite a row of stretched tungsten wires, and have a potential of 30 kV to 75 kV with respect to these wires. A corona discharge is thus produced, and this ionises the gas molecules, i.e. the molecules are split into electrons, negative and positive ions. Owing to the minute diameter of the wires, the

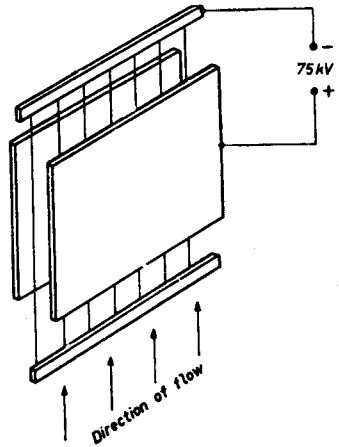


Fig. 20-13. Electrostatic dust precipitator according to Cottrell.

electrostatic field is highly concentrated in their immediate vicinity, so that they strongly attract the positive ions, which are thus removed from the discharge space. The negative ions and the electrons on the other hand travel towards the plates, colliding against the particles floating in the gas, so that these are now charged negatively, the charge being proportional to the square of their radii. These particles are then in turn attracted by the positive plates, so that they are precipitated on their surface and subsequently drop into a dust retainer underneath the plates.

This system is still being used with excellent results for precipitating soot, fly-ash, particles of tar, and so forth, notwithstanding the inherent disadvantages of requiring an extremely high direct voltage and a fairly high current due to the continually maintained corona discharge. A further drawback is the considerable amount of ozone produced by the discharge, and in view of the unpleasant smell, this procedure can be used only for purifying air that is circulated. For this reason a new system, based on the investigations of Penney, was developed, and this has none of these disadvantages, so that it is not only suitable for industrial purposes but also for purifying air in buildings.

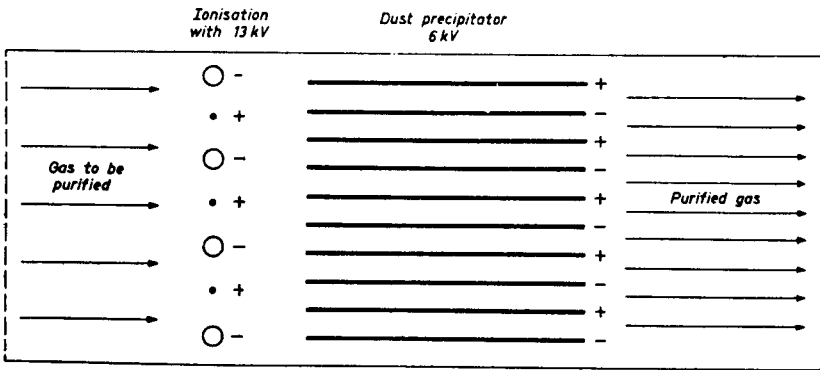


Fig. 20-14. Basic principle of an installation for purifying gases with separate dust precipitator (according to Westinghouse).

Fig. 20-14 represents the principle of a similar device for purifying air. The air first flows along a system of round metal rods between which tungsten wires are stretched. The latter have a potential of approximately +13 kV with respect to the rods, so that a non-uniform electrostatic field is produced which ionises the air molecules. As in the Cottrell system, the negative ions are attracted by the wires, whilst the floating particles are charged by the positive ions. These particles are now, however, unable to travel towards the rods, but are taken along with the air current to a separate dust precipitator consisting of several parallel metal plates with a mutual potential difference of 6 kV. Under the influence of the transversely directed electrostatic field the positively charged particles will travel to the negative plates, where they are precipitated, so that the evacuated air is completely dust-free.

By introducing a few simplifications and assuming the dust particles to be uniform, the efficiency of this procedure may be set equal to:

$$\eta = \left(1 - e^{-C \cdot \frac{F}{\Phi} \cdot v} \right) \cdot 100\%, \tag{20.1}$$

in which F denotes the effective surface area of the collecting electrodes and Φ the strength of the gas or air current. C is a proportionality factor. The velocity v of the charged particles in the transversely directed field is:

$$v = \frac{rE^2}{2\pi c}, \tag{20.2}$$

where E represents the effective electric field strength in the dust precipitator, r the radius of the particles and c the viscosity of the gas.

A number of important conclusions can be drawn from these expressions. The efficiency appears to increase with the effective surface area of the dust precipitator and with the square of the effective field strength (or of the voltage between the plates), and to decrease as the air stream increases. The expressions, moreover, reveal that the efficiency is independent of the concentration of the floating particles. It is finally seen that the efficiency increases with the radius of the particles. It is true that in practice the larger particles may again be torn away from the electrodes, particularly at high velocities of the gas flowing through the precipitator, and this obviously reduces the efficiency. By giving the electrodes a suitable shape, this effect can however be reduced to such an extent that it may be disregarded.

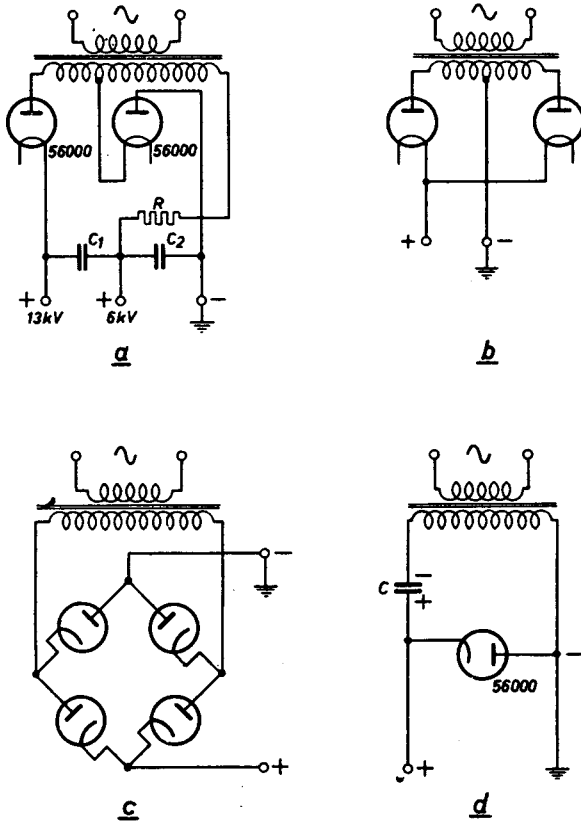


Fig. 20-15. Rectifying circuits for generating high voltages.

which are normally used for this purpose are shown in fig. 20-15.

In circuit *a* two high-vacuum, single anode rectifying tubes type 56 000 are used. In a voltage doubler circuit one direct voltage of 6 kV and one of 13 kV can thus be obtained. The current amounts to approximately 100 mA, which is sufficient for small and medium dust precipitators, as used for dwelling-houses and individual machines in factories. The resistor R has been provided to limit

The voltages required for ionising and capturing the particles are supplied by an H.T. rectifier. The circuits

the current and thus to safeguard the rectifier against flash-overs in the dust precipitator.

Circuit *b* represents a customary two-phase, half-wave rectifier and is used for medium-sized and larger installations according to Cottrell's principle.

Circuit *c* is a single-phase, full-wave rectifier. The advantage of this circuit is that, the tubes need only be designed for half the inverse voltage, but on the other hand four tubes are required, and it is a distinct disadvantage that the filament windings (not indicated) are at different potentials.

Circuit *d* represents a single-phase half-wave rectifier in which the inverse voltage across the tube is used as the output voltage. The transformer winding can in this way be designed for approximately half the desired output voltage. Moreover, there is no risk of the tube being damaged by a flash-over in the dust precipitator. The output voltage, however, drops to zero every cycle and the output current is fairly small, so that this circuit can be used only for small-sized installations.

The construction of an air-purifying installation will always have to be adapted to the individual requirements of each case. In principle, one differentiates between installations in which the gas is conducted through the dust precipitator in a horizontal direction and those using a vertical direction. The former type is suitable both for small installations to purify air in dwelling-houses and for factory installations where large quantities of dust (e.g. fly-ash) have to be removed and where the emptying of the dust container should take up little time. For individual machines, e.g. in mills, or for precipitating oil vapour of fast machine tools, or lacquer in spraying installations, preference is, however, given to an installation in which the air is evacuated vertically.

The effective surface area of the electrodes in the dust precipitator must be made as large as possible to obtain a high efficiency. Since the clearance between the plates may not be made too small on account of the risk of flash-overs (approx. 8 mm at 6 kV), the physical dimensions of the dust precipitator are usually quite considerable. The electrical power consumption is, however, small; for purifying 1000 litres per minute the required power is only approximately 0.5 W to 1 W.

Since the precipitated dust must be removed from the electrodes at regular intervals (roughly every three to six weeks), the dust precipitator is usually combined with a washing device to rinse away the dust. In installations made by Westinghouse in the U.S.A., which are intended to be combined with air conditioning plants in dwelling-houses and offices, rinsing is carried out automatically when a switch is operated. After the dust precipitator has been cleaned, the installation is switched on again automatically by means of a timer.

In larger installations for use in factories, the dust precipitator consists of several units which can be put out of operation individually for cleaning, so that the purifying of the air need not be interrupted.

The efficiency of the installations may even exceed 90%, more than 99.5% of the weight of the floating particles being captured. For purifying the air in dwelling-houses containing from 6 to 8 rooms, an installation with a capacity of 35 m³ to 45 m³ per minute will be sufficient, the total consumption being less than 60 W. The dimensions of such an installation are roughly 31½" × 28" × 44"; the cross-section of the air duct may be 20" × 24".

21. TRANSISTORS

The transistor is a semi-conducting circuit element, invented less than ten years ago. As regards build and design it went through a period of radical and rapid development, which even to-day has by no means come to an end. Like the electron tube, it is a circuit element suitable for amplifying input signals, but it differs from it by a number of essential properties. It merely depends on the task it has to fulfil whether these properties justify its application in preference to electron tubes or not.

Its most important characteristics are the following :

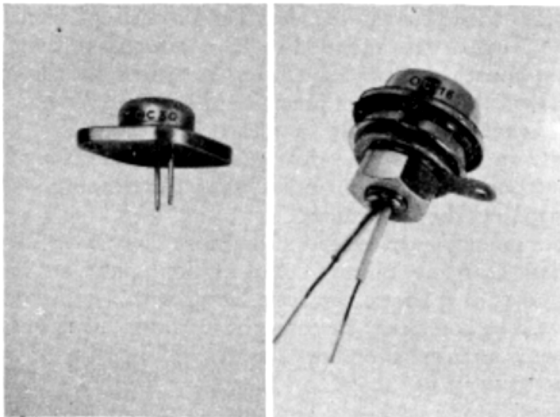
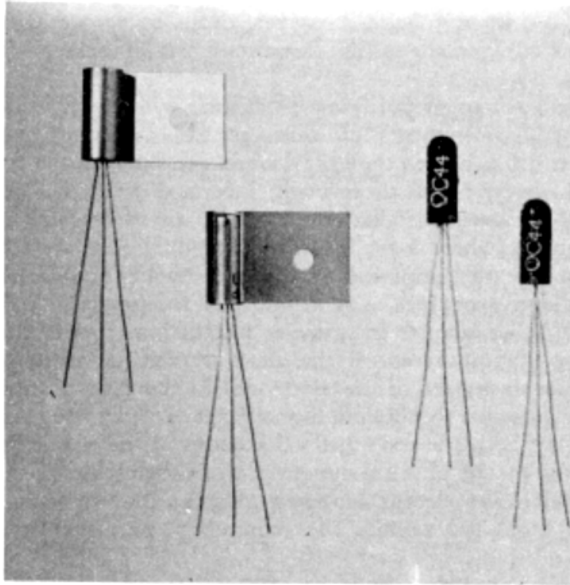


Fig. 21-1. Various designs of commercially available transistor: a) low-power transistors for input stages, b) output transistors.

- 1) The transistor does not require any heating current for its operation; moreover, its efficiency is relatively high, i.e. the d.c. power to be applied is less than for an equivalent tube circuit. This means that also the loss in heat to be dissipated is smaller.
- 2) Low tensions of the order of a few volts only are required to operate a transistor. Direct battery supply, without the aid of a direct-voltage converter, is therefore possible.
- 3) Unlike most tube circuits, a transistor amplifying stage requires a certain control power.
- 4) The electrical properties of the transistor are highly temperature-dependent; they also show more variance than those of electron tubes. Special devices to compensate for these must as a rule be incorporated in the circuit.
- 5) For the time being at least, there are certain limits to the application of transistors where very high frequencies or the handling of large powers is concerned.
- 6) The working life of transistors is undoubtedly very long. Should any defect occur, however, remedying it will be more difficult and take more time, since transistors are as a rule soldered into the wiring.

Fig. 21-1 shows various designs of commercially available transistors. For industrial electronic equipment the transistor should first of all be considered for use as low-frequency amplifier and as controlled switching element. In the following brief study, therefore, these applications in particular will be dealt with.

General Design and Working Principle

The p-n-p junction transistor, which is most generally used at present, consists of a semi-conductor crystal (mostly germanium), which has been divided by a specific process into three adjoining regions (see fig. 21-2). The thin, central region, the base, being "contaminated" by a specific dose of foreign atoms, has a preference for conducting electrons (n-layer), whilst the two outer regions have a shortage of electrons in the crystal lattice, which renders them conducting for virtual carriers of positive charge (the so-called "holes"). The p-n-p transistor is always circuited such that the emitter is positive with respect to the base, whereas the collector c is given a negative potential with respect to the base (see fig. 21-3). The collector-base region thus corresponds to a semi-conductor diode with a bias in the reverse direction, i.e. in spite of a relatively high supply voltage, only a small current is passed. Owing to the positive potential applied to the emitter, "holes" are transferred from emitter into base. Since this current flows in the forward direction of the semi-conductor diode formed by the emitter-base system, the current is large in comparison with the voltage applied. The amplifying effect of the transistor is attributable

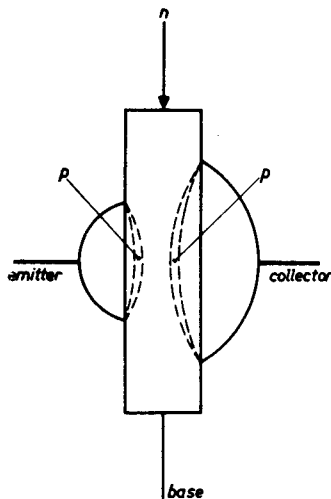


Fig. 21-2. Schematic representation of a p-n-p transistor.

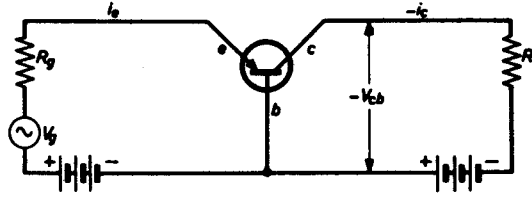


Fig. 21-3. Common-base configuration.

to the fact that virtually all the current from the emitter is passed via the base into the collector. Since the collector circuit is in fact more high-ohmic than the emitter circuit, power gain is realized. The current amplification, i.e. the ratio of collector current to emitter current, however, is in this circuit slightly less than unity, even if the load resistance R_l is dynamically short-circuited. In such a case the expression

$$a = \left[\frac{-I_c}{I_e} \right]_{-V_{cb} = 0} \quad (21.1)$$

represents the short-circuit current amplification factor. This is only valid for alternating currents; the so-called “static a ” for direct current is always smaller. If an a.c. input source V_g is incorporated in the emitter circuit, then these voltage variations, amplified by a factor of roughly R_l/R_g will appear across the load resistance R_l , since nearly the full emitter current is transferred to the collector circuit.

More customary than the “configuration with common base” or “grounded base circuit” shown in fig. 21-3 is the “configuration with common emitter” or “grounded emitter circuit” shown in fig. 21-4, since it is capable of producing substantially higher values for both current amplification and power gain.

Taking into account the relation

$$I_e + I_c + I_b = 0 \quad (21.2)$$

and the expression (21.1), we find

$$\left[\frac{I_c}{-I_b} \right]_{-V_{cb} = 0} = \frac{a}{1 - a} \quad (21.3)$$

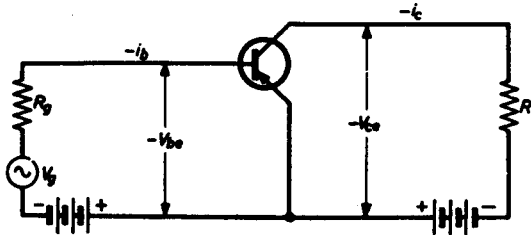


Fig. 21-4. Common-emitter configuration.

for the short-circuit current amplification.

Without any actual change we may substitute $-V_{ce} = 0$ for $-V_{cb} = 0$, so that we obtain

$$\left[\begin{matrix} -I_c \\ -I_b \end{matrix} \right]_{-V_{ce} = 0} = a' \approx \frac{a}{1 - a} \tag{21.4}$$

for the grounded emitter circuit. a' may attain fairly large values.

If, for instance, $a = 0.98$, then $a' = 49$.

Since, furthermore, the power gain is proportional to the ratio of load resistance R_l to input resistance, it will be clear that the total power gain that can be obtained with the grounded emitter circuit is considerable.

For RC-coupling the power gain for each stage may lie between 10^3 and 10^4 ; for transformer coupling even higher. This is one of the main reasons why the grounded emitter circuit, which is more or less comparable to the familiar grounded cathode circuit of an amplifying tube, is most frequently applied. In the following, therefore, we shall deal with this circuit only.

Fig. 21-5 shows the characteristics of a p-n-p transistor in grounded emitter circuit. They comprise four sets of curves in four quadrants, the first (right-top) representing the relations at the output side, and the third (left below) those at the input side, whilst the second, and the fourth represent the relationship between output and input.

The first quadrant contains the $I_c - V_{ce}$ characteristics with $-I_b$ as parameter, which greatly resemble the $I_a - V_c$ characteristics of a pentode, with the difference, however, that here the knee, and hence the knee voltage, is appreciably nearer to the ordinate axis than in the case of a normal pentode. The internal resistance is accordingly very low, which means that the transistor can be effectively used as a switching element. If a load resistance R_c is incorporated in the collector circuit, then a straight load line is described in the $I_c - V_{ce}$ diagram, corresponding to the value of the negative base current between the points A and B, the latter representing the conducting and the blocked

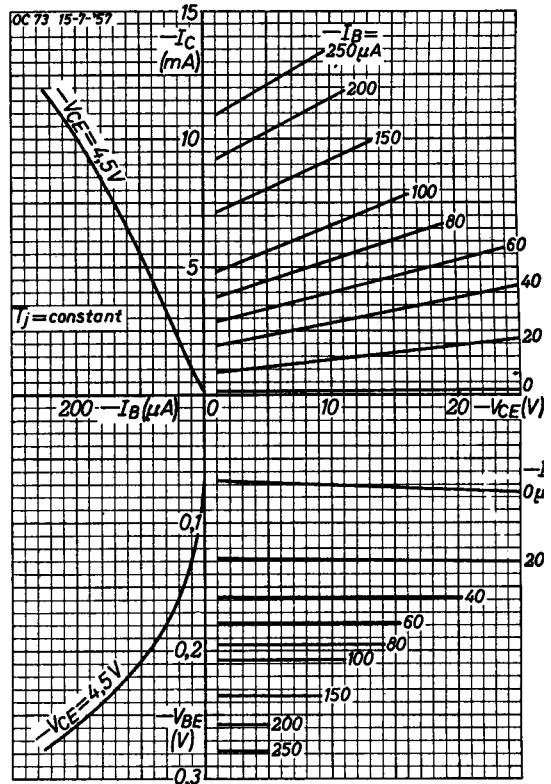


Fig. 21-5. Characteristics of a p-n-p transistor in common-emitter configuration.

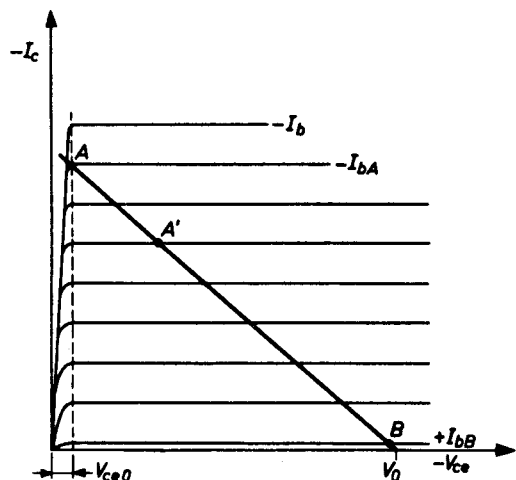


Fig. 21-6. Illustration of the operation of the transistor as a switching element (family of $I_c - V_{ce}$ characteristics with load line).

condition (see fig. 21-6). V_{ce0} represents the low residual voltage remaining in the conducting condition. V_{ce0} lies somewhere between 0.4 and 0.8 V.

The third quadrant in fig. 21-5 (left below) shows the input characteristic (base voltage versus base current). This line is not straight, but has a certain curvature, which means that an alternating voltage ΔV_{be} applied at the input will produce a distorted base current $-I_b$. The $I_b - I_c$ characteristic in the second quadrant may be considered as virtually straight where small signals are concerned. To avoid distortion it is therefore recommended, instead of applying an alternating voltage at the input, to control the transistor via the base current, viz. by using a control-voltage source with the highest possible dynamic internal resistance. In this case the part played by the shape of the $I_b - V_{be}$ characteristic may be practically disregarded, as the base current is directly influenced, the latter controlling the collector current in accordance with the $I_b - I_c$ characteristic shown in the second quadrant. Where large signals are concerned, the curve of the $I_b - V_{be}$ characteristic will indeed cause some distortion, but this will be approximately compensated by the curvature in the $I_b - I_c$ characteristic, which is then no longer straight, thanks to the fact that the two curvatures are opposed. For large amplitudes of the input signal, voltage control may therefore be preferable.

For small collector currents the $I_b - I_c$ characteristic is practically straight; the tangent of its slope directly represents the current amplification α' .

The fourth quadrant (right below) represents the "reverse voltage amplification factor", i.e. the feedback effect of a change in collector voltage on the base voltage. This is evidently only very small and may be generally disregarded.

Since in transistor circuits we are concerned with four variables, namely voltages and currents at both input and output, the transistor can be regarded and treated as a four-terminal network, or fourpole. Only the theory of fourpole networks, in fact, will provide the quantitative relationship between input and output power and the appropriate matching conditions. Entering into this subject matter, however, would go beyond the scope of this book, and the reader is referred to the relevant specialized literature.

Multivibrator Circuit with Transistors

Fig. 21-7 shows the circuit of a bistable multivibrator. The base electrodes of the two transistors are given a certain negative bias from two junctions of the voltage divider R_c , R_m , R_b , and are furthermore a.c. coupled via the capacitors C_m , each with the collector of the other transistor. If, for instance, the collector current of the right-hand transistor increases, the negative potential of point A will drop, and the current through the base of the left-hand transistor becomes less negative. This means that the collector current of the left-hand transistor decreases, which in its turn raises the negative potential of point B. This increases the collector current of the right-hand transistor even further, and the process continues until ultimately the right-hand transistor is completely conducting and the left-hand transistor is completely blocked.

The base current of the latter has then approached zero or even become slightly positive (cf. fig. 21-6). If a negative pulse is now applied to the input of the circuit, there is a momentary drop of the emitter potential of the transistors, which has the same effect as if a positive pulse were applied to the base electrodes. This directly blocks the right-hand transistor, the negative potential of A suddenly increases, and this negative pulse is transferred via C_m to the base of the left-hand transistor. This causes the left-hand transistor to carry collector current, and the change-over into the other stable position has been initiated. Any sudden change in collector current of the

right-hand transistor produces, at the secondary of the transformer Tr_2 , a voltage pulse of alternating polarity, which may be employed for controlling a following multivibrator stage.

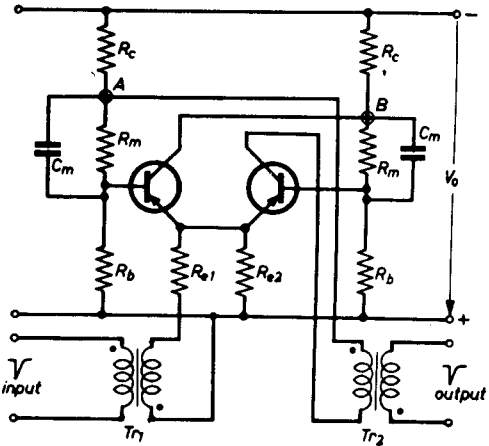


Fig. 21-7. Circuit of a bistable multivibrator with transistors.

Periodic Switch

Fig. 21-8 shows a power switch which is alternately opened and closed by an astable multivibrator (blinker-signal circuit). The load is incorporated in the collector circuit of an output transistor OC 16. The latter will be conducting when the negative base current is sufficiently large, i.e. when control transistor OC 72 is conducting. The value of the negative base current of the power transistor, and accordingly the maximum output current, can be adjusted by means of potentiometer R_p in the collector circuits of the OC 72. Control transistor OC 72 is itself controlled by a third transistor, which is combined with a fourth transistor into an astable multivibrator arrangement. Let us assume that the left-hand transistor is blocked and the right-hand transistor is conducting. The collector of the left-hand transistor is then nearly at the potential of the supply voltage, and the

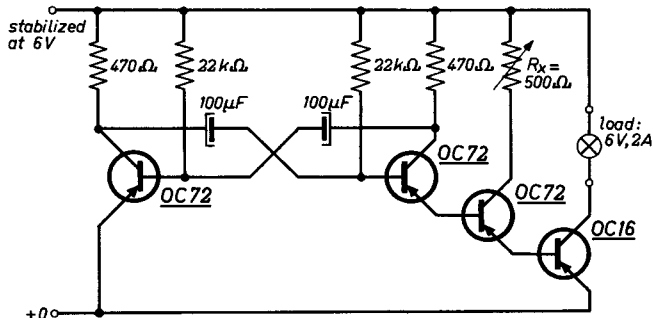


Fig. 21-8. Circuit of an astable multivibrator with output stage.

left-hand coupling capacitor is charged. The base current of the left-hand transistor is becoming increasingly negative owing to the gradual discharging of the right-hand coupling capacitor, which causes the left-hand transistor to become conducting, whilst its negative collector potential is dropping.

Because of charging of the left-hand coupling capacitor, the base current of the right-hand transistor approaches zero, thus blocking it. The collector potential approaches that of the supply voltage, and the right-hand coupling capacitor is charged. As soon as the left-hand capacitor has been sufficiently discharged, the right-hand transistor becomes conducting again and the left-hand one blocked. The resulting multivibrator frequency depends upon the capacitance of the coupling capacitors and the value of the base resistors.

Phase-sensitive Switch

Another example of electronic control by means of transistors is shown in fig. 21-9, representing the circuit of a phase-sensitive power switch. This circuit can be used in conjunction with a regulating device for influencing control devices with two discrete conditions, e.g. for operating a magnetic clutch. The magnet coils of the clutch for clockwise or anti-clockwise rotation must then be inserted instead of the pilot lamps La_1 and La_2 shown in the circuit. Instead of pilot lamp La_2 , a relay may be incorporated for switching on the electric motor driving the clutch. In the example considered here we see to the left of the dotted line two a.c. fed voltage dividers, either of which can be manually circuited, each voltage divider forming a bridge circuit in combination with the secondary winding of the mains transformer. When the bridge equilibrium is passed, as is known, the phase of the bridge output voltage changes by 180° , whilst the amplitude is a measure of the extent of the unbalance. In one case the bridge is unbalanced by adjusting the potentiometer, in the other by resistance variation of the cadmium sulphide photocell ORP 30. In practice there may be to the left of the dotted line e.g. the output of a measuring device followed by an amplifier.

A transistor OC 76 is used here for the input stage, and a type OC 30 for the output stage. The OC 76 is used in grounded-collector arrangement, in order to present a low matching impedance to the following OC 30. This circuit resembles the cathode-follower circuit with high-vacuum tubes described elsewhere in this book. The collectors of both transistors are fed by a pulsating direct voltage.

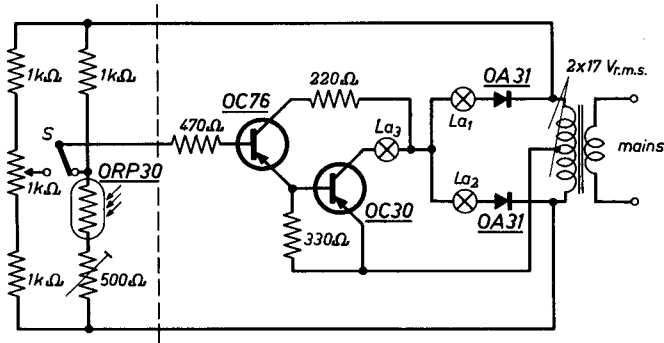


Fig. 21-9. Circuit of a phase-sensitive switch.

As soon as the bridge unbalance has risen to a certain value, current is supplied to La_3 (or the driving electric motor is switched on). In this case the polarity of the unbalance, i.e. the phase of the bridge output voltage, is immaterial. The two other pilot lamps La_1 and La_2 are each connected in series with a diode OA 31, and each receives only one half-wave of the applied alternating voltage. These half-waves are in opposite phase, since the emitter of transistor OC 30 is connected to the centre tapping of the main transformer. According to the phase of the a.c. signal controlling the base of transistor OC 76, either La_1 or La_2 will light up in combination with La_3 .

The output power amounts to about 8 W, the peak collector current of the OC 30 being 1.3 A. The magnetizing windings of the magnetic clutch to be incorporated instead of La_1 and La_2 , may be designed accordingly.

Phase-independent Switch

Fig. 21 - 10 shows the circuit of a simple switch employing transistor OC 76 and having a switching capacity of 3.7 W. The output stage is controlled by a pulse shaper with two transistors OC 71 which has the object of converting an input voltage of any shape into rectangular pulses. By means of pulses of this kind, the output transistor is very rapidly changed from the blocked into the conducting state. This is essential for a high switching capacity, as during a more gradual change-over the transistor would be impermissible overloaded.

The operation of the Schmitt trigger circuit is based on the fact that the emitter circuits of the two transistors OC 71 are coupled via the common $100\ \Omega$ resistance, whilst also the collector of the left-hand transistor is coupled to the base of the right-hand one. When an input signal causes the negative base current of the left-hand transistor to increase, its collector potential becomes less negative, and the same happens to the base potential of the right-hand transistor. At the same time, however, the emitter potential is changing in the negative direction. These two changes in potential have the effect of rapidly blocking the right-hand OC 71, with the result that output transistor OC 76 becomes suddenly conducting. When the circuit is switched on, a current of 125 mA is flowing through the $240\ \Omega$ load resistance, the residual voltage at the transistor amounting to 0.3 V.

The transistor can remain in this condition for any period of time without being

overloaded, as long as a base current of approx. 3.5 mA is being applied to its input. This current is provided by the Schmitt trigger, which in its turn requires a tension of only 20 mV applied to its input (approx. 5 kΩ) to be completely controlled.

With the device switched off, the residual current through the load resistance is about 0.03 mA. In this condition too, the switching device can remain for any desired period. The collector voltage is then approx. 30 V.

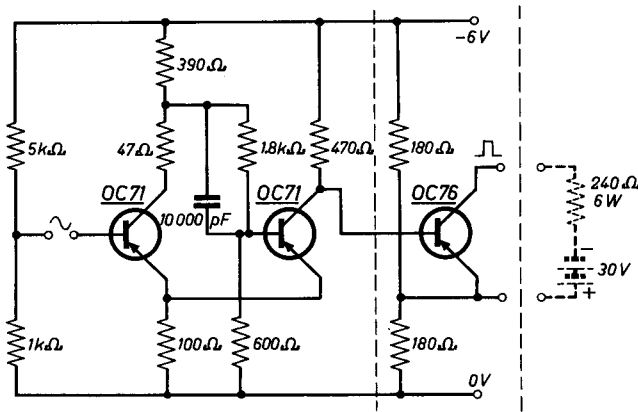


Fig. 21-10. Simple power switch with pulse shaper circuit.

Apart from being very slightly influenced by stray pulses, the operation of the switch is stable up to a temperature of 45°C without requiring any additional compensating devices.

Photoelectric Relay

The circuit of fig. 21-11 is a photoelectric relay employing three transistors in the oscillator, amplifying and relay stages. One transistor OC 71 operates as self-excited oscillator; its frequency (approx. 10 kc/s) is determined by the oscillating circuit comprising capacitor C_6 and the central winding of transformer Tr_2 . Feedback is effected via winding n_2 , whilst an alternating voltage of 0.7 $V_{r.m.s.}$ developed across n_3 is applied to a bridge circuit comprising R_2 and two germanium diodes. The output side of the bridge contains a photocell, which produces a certain e.m.f. under the influence of light. During the negative half-cycles of the 10 kc/s voltage produced across n_3 , the diodes are blocked, so that the e.m.f. of the photocell is connected, in series with the high inverse resistance of the diodes, to the input resistance R_3 of the transistor amplifier, where it produces virtually no effect. During the positive half-cycles, however, the low forward resistances of the diodes are in series with the e.m.f. of the photocell, and to the transistor OC 71, pulses are applied which, after being amplified and applied to the base of the transistor OC

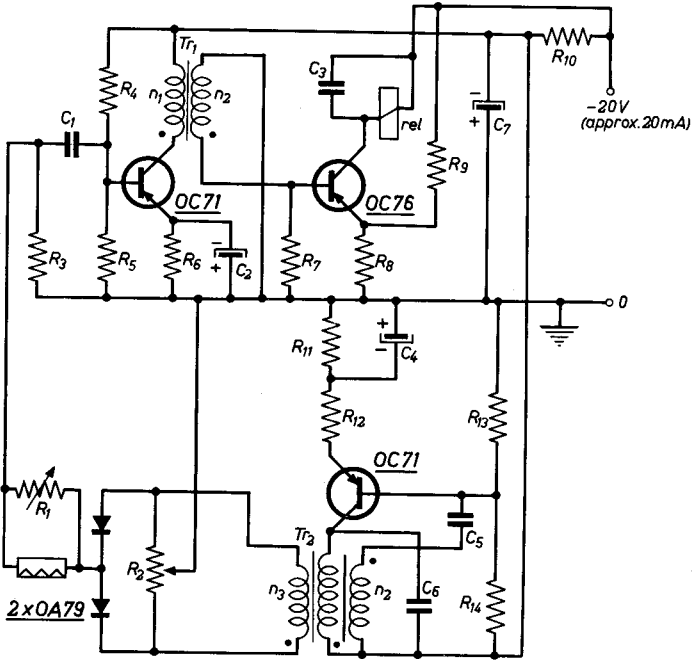


Fig. 21-11. Photoelectric relay

76, produce a corresponding collector current. When the photocell receives no light, no pulses are fed to the transistor amplifier, and no collector current will flow, since the emitter is given a negative bias via the voltage divider R_8, R_9 .

Data of the circuit of fig. 21-11

$R_1 = 10 \text{ k}\Omega$	$R_8 = 18 \Omega$
$R_2 = 200 \Omega$	$R_9 = 1.8 \text{ k}\Omega (1/2 \text{ W})$
$R_3 = 2.2 \text{ k}\Omega$	$R_{10} = 1 \text{ k}\Omega$
$R_4 = 15 \text{ k}\Omega$	$R_{11} = 2.7 \text{ k}\Omega$
$R_5 = 10 \text{ k}\Omega$	$R_{12} = 120 \Omega$
$R_6 = 10 \text{ k}\Omega$	$R_{13} = 5.6 \text{ k}\Omega$
$R_7 = 1 \text{ k}\Omega$	$R_{14} = 4.7 \text{ k}\Omega$
$C_1 = 0.1 \mu\text{F}$	$C_5 = 0.1 \mu\text{F}$
$C_2 = 2.5 \mu\text{F}$	$C_6 = 0.05 \mu\text{F}$
$C_3 = 0.1 \mu\text{F}$	$C_7 = 25 \mu\text{F}$
$C_4 = 2.5 \mu\text{F}$	

Transformer Tr_1 :

$n_1 = 700$ turns 0.1 mm Cu ; $L_1 = 280$ mH

$n_2 = 110$ turns 0.2 mm Cu

Core: Ferroxcube pot core D 18/12-00 — IIIB, Philips

Transformer $T r_2$:

$n_1 = 220$ turns 0.1 mm Cu ; $L_1 = 5.3$ mH

$n_2 = 80$ turns 0.1 mm Cu

$n_3 = 50$ turns 0.1 mm Cu

Core: Ferroxcube pot core D 18/12-05 — IIIB, Philips

Stabilized Power-supply Unit for Transistor Circuits

For control devices that have to meet high demands as regards precisions, fluctuations of the mains, such caused by load fluctuations when large consumers are switched on the power line, are most objectionable. In industrial practice, sensitive measuring and control apparatus are often connected to the same mains system as large consumers of power. For such equipment a stabilized supply is frequently required. The use of transistors, which require low supply voltages, makes it necessary to employ low-ohmic power packs which should accordingly also be transistor-operated.

The circuit shown in fig. 21-12 will produce a stable output voltage even under greatly fluctuating load conditions. This circuit has been so designed that its output voltage, derived from two-phase rectification of 2×15 V a.c., remains constant within 0.3% during current fluctuations from 0 to 2 A and mains-voltage fluctuations of 10%.

The direct current obtained from the two germanium diodes OA 31 is smoothed and applied to a transistor cascade. The base potential of the first OC 72 is taken from the output-voltage divider and compared with the voltage of the emitter circuit, the latter being kept constant by two standard cells. A change in collector current causes a shift in the base potential of the second OC 72 of the transistor cascade; this again changes the resistance of the collector-emitter region

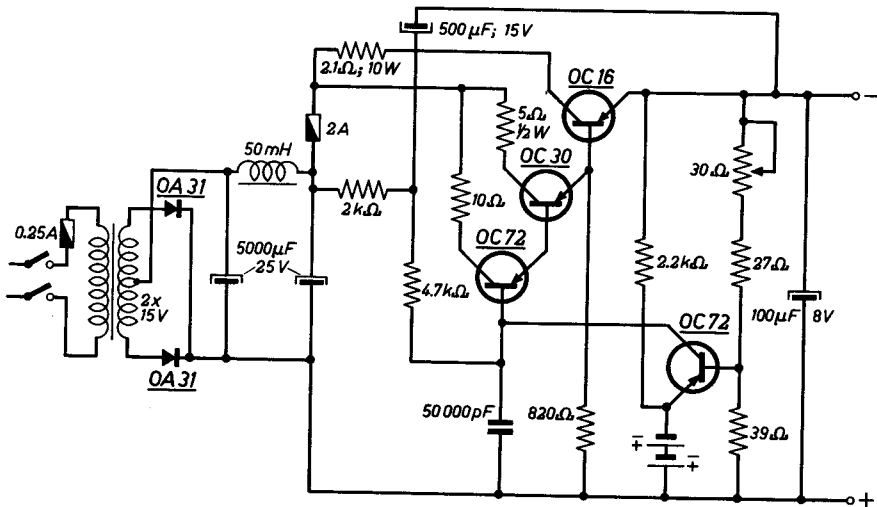


Fig. 21-12. Stabilized power-supply unit for transistor circuits.

of the output transistor OC 16. Owing to the high gain of about $1.5 \cdot 10^6$ of the four transistors engaged in the regulating process, the stabilizing effect is most satisfactory. The circuit is characterized by the following data:

output voltage adjustable between 5.4 and 7.5 V

output current: 0 2A

internal resistance: approx. 0.025 Ω

hum voltage at any adjustment and load: approx. 20 μ V

10 % mains-voltage fluctuation produces max. 0.3% output-voltage fluctuation.

Mains transformer:

Core: EI 78; primary: 1300 turns, 0.25 mm \emptyset Cu

secondary: 2 \times 89 turns, 0.8 mm \emptyset Cu

Choke coil:

92 turns, 1.50 mm \emptyset Cu on EI 78 core without air-gap.

2 Neumann standard cells, 5 mA reactive current, type 1.5/10

If not stated otherwise, all resistors are rated for 0.25 W.

CONCLUSION

The purpose of this book has been to indicate the wide field of application of electron devices in industry, and to describe the principles upon which a large number of typical equipments operate. It would, of course, have been impossible in a single volume of reasonable size to treat this vast subject exhaustively. Special problems, as they arise, often call for individual solution, and valuable guidance can be obtained from the extensive literature available, a selection of which is noted in the Bibliography.

It will readily be understood that the initial cost of an electronic device is sometimes greater than that of the more conventional installation which it is intended to replace. In most cases, however, the lower production costs, both direct in increased output and indirect by way of improved quality, rapidly recoup the initial capital expenditure, and this must be borne in mind when deciding whether or to what extent the use of electronic methods, with possibly increased initial expenditure, can be justified. Of course there will always be cases in which the higher capital cost does not bring any economic advantages and it is no part of the Author's policy to suggest that electronics always provides the most favourable solution even though it may be the most perfect from the technical view point.

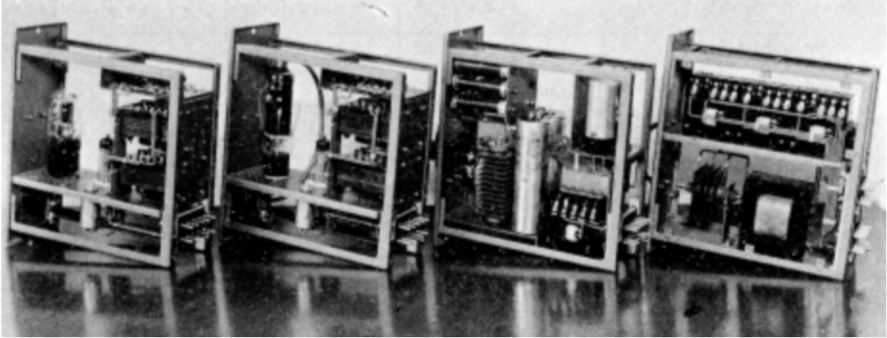
It is not advisable that the construction of electronic equipment should be undertaken by the user, unless he has a special knowledge of electronics, for this can only lead to disappointment and loss. It is for this reason that component values are specified in the book only for the simpler circuits which can be easily constructed by those possessing a reasonable amount of technical knowledge and having available at least the indispensable minimum of measuring and testing instruments.

It should also be pointed out that description of any circuit in this book does not imply permission to use any device which may be patented.

Something must now be said concerning the reliability of electronic equipment. Occasionally doubts are expressed by industrial engineers as to the reliability of apparatus and tubes, particularly under severe working conditions. It can be stated at the outset that recent developments in tube design and manufacture have resulted in the production of tubes which are no less reliable than other components used in industrial equipment — for example electro-mechanical contactors — provided the maximum permissible ratings published by the manufacturers are not exceeded. There is no reason, therefore, to doubt the reliability of modern electronic equipment if it is given the reasonable amount of care and maintenance which would be given to any other pieces of industrial apparatus or machinery. It is, however, very essential that facilities for any necessary repair work should be available on the spot in order to avoid hold-up of the manufacturing processes. At least one spare tube of each of the types

employed should be kept at hand, in the same way that every engineering shop keeps a sufficient reserve of twist drills, milling cutters, saw blades, etc.

It is also of importance that the construction and layout of the apparatus is simple and accessible so that individual components, or even complete assemblies, can be easily replaced. As can be seen from the illustrations, manufacturers of electronic equipment facilitate repair work by constructing all major equipment as a number of units which can easily be put together or disconnected (see fig.).



Electronic installation comprising several units.

Finally, the importance of thorough training of personnel in charge of the operation or maintenance of equipment cannot be too strongly emphasised. It is certain that much of the hesitancy in adopting electronic methods occasionally encountered in industry stems from doubts as to whether the installation staff, maintenance staff, and even the engineer in charge, possesses the required special technical knowledge. It is obvious that there is a great need of further schemes for industrial training in electronics, and it is hoped that this book may be some contribution to this end.

Principal Data of some Preferred Types of Philips Tubes

PHILIPS INDUSTRIAL RECTIFYING TUBES

Type	Double anode tubes								
	328	1010	1725A	1110	1119	1710	367	1048	1039
Filament voltage (V)	1.9	1.9	1.9	1.9	1.9	1.9	1.9	1.9	1.9
Filament current (A)	3.0	3.5	3.5	3.5	5.8	7	8	7	20
Max. a.c. anode voltage (r.m.s. value) (V)	28	60	150	60	45	150	45	60	60
Min. a.c. anode voltage (r.m.s. value) (V)	16	16	22	16	12	25	16	10	10
d.c. output current per tube (A)	1.3	1.3	1.3	1.7	3	3	6	6	15
Peak value of the anode current (A)	4	4	4	5	9	9	18	18	45
Min. total anode resistance (Ω)	3	10	5	4	1.8	2.5	1	1.75	0.75
Peak value of the inverse anode voltage (V)	90	185	470	185	140	470	140	185	185
Ignition voltage (V)	16	16	22	16	16	22	16	16	16
Arc voltage (V)	7	9	10	9	9	10	9	9	9
Max. horizontal width (mm)	33	37	71	39	71	69.5	81	81	94
Max. total height (mm)	112	120	135	131	142	205	170	170	264

INDUSTRIAL RECTIFYING TUBES (continued)

Type	Double anode tubes					Single anode tubes			
	1838	1049	1849	1859	1069K	1173	1174	1176	1177
Filament voltage (V)	1.9	1.9	1.9	1.9	3.25	1.9	1.9	1.9	1.9
Filament current (A)	21.5	28.5	29	60	70	13	12	28	60
d.c. output current per tube (A)	15	25	25	50	60*)	4	6	15	25
Peak value of the anode current (A)	45	75	75	150	200	20	30	75	135
Min. total anode resistance (Ω)	0.25	0.3	0.2	0.1	0.12	0.75	0.5	0.1	0.1
Peak value of the inverse anode voltage (V)	360	185	360	360	170	850	850	850	850
Ignition voltage (V)	22	16	22	28	16	22	22	22	28
Arc voltage (V)	10	9	10	12	10	12	12	12	12
Max. horizontal width (mm)	97	101	106	143	114	62	77	92	128
Max. total height (mm)	262	280	294	436	365	189	218	301	362

*) With fan cooling.

PHILIPS THYRATRONS

	PL 5557	PL2D21	PL 5559	PL 105	PL 150	PL 1807	PL 5544	PL 255
Number of electrodes	3	4	3	4	3	4	3	3
Filling	Hg-vapour	Xenon	Hg-vapour	Hg-vapour	Hg-vapour	Xenon	Xenon	Hg-vapour
Cathode:								
Heating	dir.	ind.	ind.	ind.	dir.	dir.	dir.	ind.
Heater voltage (V)	2.5	6.3	5	5	1.9	2	2.5	5
Heater current (A)	5.0	0.6	4.5	10	26	2.6	12	14
Heating time (typical)	5 sec	10 sec	5 min	5 min	1 min	30 sec	1 min	5 min
Arc voltage drop (V)	12	8	12	12	12	15	12	10
Control characteristic	neg.	neg.	neg.	neg./pos.	neg./pos.	neg.	neg./pos.	neg./pos.
Deionization time (μ sec)	1000	75	1000	1000	1000	500	neg./pos. 40 ^{*)} , 400 ^{*)}	1000
Maximum peak anode voltage:								
Inverse (V)	5000	1300	1000	2500	500	650	1500	2500
Forward (V)	2500	650	1000	2500	240	650	1500	1500
Maximum negative control-grid voltage:								
Before conduction (V)	500	100	500	1000	150 ¹⁾	100	250	300
During conduction (V)	10	10	10	10	50 ²⁾	10	10	10
Maximum negative screen-grid voltage:								
Before conduction (V)	—	100	—	500	—	100	—	—
During conduction (V)	—	10	—	10	—	10	—	—
Maximum anode current:								
Instantaneous, 25 cycles and above (A)	2	0.5 ⁵⁾	15 ⁵⁾	40	90	2	40 ⁵⁾	80 ⁵⁾
Instantaneous, below 25 cycles (A)	1	—	5 ⁵⁾	12.8	—	1	—	—
Average (A)	0.5	0.1 ⁵⁾	2.5 ⁵⁾	6.4	15	0.5	3.2 ⁵⁾	12.5 ⁵⁾
Maximum control-grid current:								
Instantaneous (A)	—	—	—	1	0.1	0.25	—	1
Average (A)	0.05	0.01	0.25	0.25	—	0.05	0.2 ⁵⁾	0.25 ²⁾
Maximum screen-grid current:								
Instantaneous (A)	—	—	—	2	—	0.25	—	—
Average (A)	—	0.01	—	0.5	—	0.05	—	—
Maximum averaging time of currents (sec)	15	30	15	15	15	15	15	15
Temperature limits (°C)	35 to 80 ^{*)}	-75 to +90 ^{**)}	40 to 80 ^{*)}	40 to 80 ^{*)}	40 to 80 ^{*)}	-75 to +90 ^{**)}	-55 to +70 ^{**)}	35 to 75 ^{*)}
Maximum resistance between control grid and cathode (M Ω)	50	20	60	60	60 to 70	20	20	60
	0.01 to 0.1	1 to 10	0.01 to 0.1	0.1	0.03	0.1 to 5	0.1	0.05

*) Temperature of condensed mercury
 **) Ambient temperature

1) V_g negative
 2) V_g positive
 3) at $V_g = -250$ V
 4) at $V_g = -12$ V
 5) Cathode current
 6) Averaging time max. 1 cycle

PHILIPS IGNITRONS

Welder control service; two tubes in inverse parallel					Three phase welding							
Type ↓	r.m.s. anode supply voltage (V)	Max. demand (kVA)	Max. average anode current (A)	Max. aver- aging time (sec)	Type ↓	Max. peak forward anode voltage (V)	Max. peak anode current (A)	Average anode current (A)	Max. average anode current (A)	Max. aver- aging time (sec)		
PL 5551	220	530	30.2	18	PL 5822	1500	1200	16	—	6.25		
		180	56				336	—	56			
	250	600	30.2	18		1200	1500	20	—		—	
		200	56				420	—	70			
	600	600	30.2	7.5		Continuous rectifier service						
		200	56									
PL 5552	220	1060	75.6	14	Type ↓	Max. peak forward anode voltage (V)	Max. peak inverse anode voltage (V)	Max. peak anode current (A)	Max. average anode current (A)	Max. surge current (A)		
		350	140									
	250	1200	75.6	14								
		400	140									
	600	1200	75.6	5.8								
		400	140									
PL 5555	2400	2400	135	1.66	PL 5555	900	900	1800	200	12000		
		1105	207			2100	2100	1200	150	9000		

PHILIPS BI-METAL RELAY 4152

Heater current (mA)	Resistance of thermo- element (Ω)	Delay time at 92 mA (sec)	Maximum current during switching on		Maximum current during switching off		Max. diam. (mm)	Max. length (mm)
			at 220 V \cong (A)	at 380 V \sim (A)	at 220 V \cong (mA)	at 380 V \sim (mA)		
92 \pm 13%	340- 372	30 *)	1.5	0.7	250	75	35	146

*) The delay time is independent of the ambient temperature. At 105 mA it is reduced to approx. 60 sec. whereas at 80 mA it is increased to approx. 110 sec.

PHILIPS TRANSMITTING TUBES

Type	Filament voltage (V)	Filament current (A)	Anode voltage (kV) max.	Anode dissipation max. (kW)	Industrial operating conditions	
					Transformer voltage (kV _{rms})	Output power approx. (kW)
TB 2,5/300	6.3	5.4	2.5	0.135	2.2 *)	0.29
TB 3/750	5	14.1	3	0.25	2.5 *)	0.49
TB 4/1250	10	9.9	4	0.45	3.35 *)	1.04
TBL 6/5000 } TBW 6/6000 }	12.6	33	6	5 6	4.4 **)	4.5
TBL 12/100 } TBW 12/100 }	17.5	196	15	45 50	9.5 **)	90

*) Two-phase, half-wave rectification (voltage of each phase).

***) Three-phase, full-wave rectification (voltage between phases).

PHILIPS HIGH-VOLTAGE RECTIFYING TUBES

Type	Filament voltage (V)	Filament current (A)	Peak inverse voltage max. (kV)	Anode current max. (A)	Total height max. (mm)	Diameter max. (mm)
DCG 1/250	4	2.5	3	0.25	115	48
DCG 4/1000 G	2.5	4.8	10	0.25	157	49
DCG 5/5000 GS	5	7	13	1.5	222	59
DCG 12/30	5	13.5	27	2.5	384	120
DCG 7/100	5	20	15	10	417	117
DCX 4/1000	2.5	5	10	0.25	156	53
DCX 4/5000	5	7.1	10	1.25	216	59

PHILIPS PHOTOTUBES

Type	58CG	59CV	90AG	90AV	90CG	90CV	3545	3546	3554	
Vacuum or gas-filled	g	v	g	v	g	v	v	g	g	
Radiation sensitivity	red	red	blue	blue	red	red	red	red	red	
Projected sensitive cathode area (cm ²)	1.1	1.1	4	4	2.4	2.4	0.9	0.9	5.2	
Operating characteristics	Supply voltage (V)	85	50	85	85	85	50	90	90	90
	Dark current max. (μA)	0.1	0.05	0.1	0.05	0.1	0.05	0.05	0.1	0.1
	Sensitivity *) (μA/l)	108	20	130	45	125	20	25	150	150
	Anode load resistance (MΩ)	1	1	1	1	1	1	1	1	1
Max. ratings	Supply voltage (V)	90	250	90	100	90	250	250	90	90
	Cathode current (μA/mm ²)	0.015	0.03	0.006	0.0125	0.007	0.03	0.05	0.02	0.02
	Ambient temperature (°C)	100	100	70	70	100	100	100	100	100
Anode-to-cathode capacitance (pF)	3.0	3.0	0.7	0.7	0.7	1.1	2	2	3.4	
Tot. height max. (mm)	33	33	54	54	54	54	73**)	73**)	103	
Diam. max. (mm)	16	16	19	19	19	19	16.5	16.5	30	

*) Measured with a lamp of colour temperature 2700 °K.

***) 64 mm when provided with Pee-Wee base instead of 2-pin base.

SUBJECT MATTER INDEX

	page		page
A.C. resistance	60	Braking, electrical	222
Action, inverter.	176, 222	Burned igniting electrodes	58
Air leakage	57	Caesium	65
Alternating voltage, stabilised	268	Caesium-antimony cathode	67
Amplification factor	8	Calibration oscillator	275
—, gas	68	Capacitive heating	245
Amplifier, class A	12	— igniter	46
—, pre-	70, 87	Capacitor, reservoir	17
Amplifying stage	7	Cathode	5
—, tube	5	Cathode, caesium-antimony	67
Angle, extinction	132	—, cold	59, 71
—, extinguishing	21	—, mercury pool	50
—, firing	21, 35, 130	—, oxyde-coated	19
—, loss	246	—, photo-emissive	65
— of lag	36	Cathode-ray oscilloscope	79
—, phase	44	— tube	75
Anode	5	Cathode spot	46
—, auxiliary	46, 52, 59, 155	Cavitation	270
— current	5	Changer, frequency	206
— resistance, optimum	241	Characteristic, constant current	12
—, zero-point	149	—, control	29
Apertured disc	87	—, dynamic	7
Apparent power	128	—, $I_a - V_a$	68
Arc-back	29, 33	—, igniting	34
Arc voltage	20, 51	—, speed/torque	211
Armature voltage	209	Charge, space	19
Asymmetrical deflection	78	Choke, balance	44
Auxiliary anode	46, 52, 59, 155	Circuit, counting	95, 204
— rectifier	187	—, current-limiting	213
Average current	54	—, Eccles-Jordan	98
Averaging time	30, 54	—, firing	49
Back-to-back	53	—, flip-flop multivibrator	98
Balance choke	44	—, gate	107, 113
Barium	5	—, multiple stabiliser	63
Base, time-	78	—, multiplying	112
Battery charger, two-phase	142	—, NEMA weld timer	122
Battery-charging equipment	144	—, oscillatory	9, 87, 240
Beam, ribbon-shaped electron	102	—, reduction	95
Bias, grid	11	—, ring counter	74, 96
Bi-metal relay	144	—, scale-of-two	96
Bi-stable multivibrator	113	—, stage lighting	166
Blocking oscillator	275	—, supervisory	91
Boiler, control of oil-fired	91	—, tank	9
Braking	221	—, time delay	91
—, dynamic	221	—, voltage doubler	280

	page		page
Class A amplifier	12	Current, average	54
Class C operation.	11	—, dark	67
Clean-up, gas.	27, 57	—, eddy	236
Coaxial resonator	266	—-limiting circuit.	213
Coefficient, coupling.	250	—-limiting resistor	143
Cog-wheels, hardening.	243	—, peak.	129
Coil, inductor	242	—, saturation.	241
—, load	236	—, transformer.	213, 262
Cold cathode.	59, 71	Cycle, duty.	190
— resistance	58	—, welding	193
Colour filter	86	Dark current.	67
Colpitts oscillator.	240	D.C. shunt-wound motor.	208
Commutation period	26	Dead zone.	173
— time	127	Decade counter tube	101
Compensation, IR	213	Deflection, asymmetrical.	78
Concentrator	242	De-ionisation time	30
Condition, no load	15	Delay, phase	56
Conductance, mutual	223	—, time	86
Conducting time	23, 42	Demand, kV_a	54
Constant current characteristic	12	Detector, electronic smoke.	93
Constant, dielectric	246	Device, recording.	115
Constant output voltage.	144	Diagram, $I_a - V_a$	6
Constant-power drive	220	—, $I_a - V_a$	5
Constant, time	116	Dielectric constant	246
Constant-torque drive	219	Dimmer	167
Continuously-acting timer	120	Diming of lamps	157
Control characteristic	29	Diodes.	17
—, electronic motor.	208	Disc, apertured.	87
— grid	34	Discharge, corona.	279
—, grid	139	—, glow.	68
—, heat	193	Dissipation, power	15
—, horizontal.	35, 140	Distortion, trapezoidal.	78
—, lateral register.	173	Dividing.	115
Controlled rectification.	137	Doubler circuit, voltage.	280
— rectifier.	42	—, frequency.	256
— two-phase half-wave rectification.	130	Drive, constant-power.	220
Controlling fluorescent lamps.	158	—, constant-torque	219
Control of oil-fired boiler	91	—, multi-motor.	219
— resistance welding.	185	—, variable high-speed	177
— seam welding.	193	Driving power	12
—, paper register.	168	— stage, crystal-controlled.	254
— range.	209	Dust precipitators, electronic.	278
—, register.	170	Duty cycle.	190
—, reversing.	94	— factor.	55
—, temperature.	180	Dynamic braking	221
—, vertical.	37, 139	— characteristic.	7
Cooling water	52	Dynamo, tacho-	212
Core reactor, saturable.	146	Eccles-Jordan circuit	98
Corona discharge	279	Eddy current.	236
Correcting motor.	171	Effect, skin.	236
Cottrell	278	Efficiency	237, 279
Counter-electromotive force	20	Electrical braking.	222
Counter, predetermined	109	Electrodes, burned igniting	58
—, revolution	271	Electromotive force	208
— tube, decade	101	Electron beam, ribbon-shaped	102
Counting circuit	95, 204	Electronic brain.	208
Coupling coefficient.	250	— dust precipitators	278
Crystal-controlled driving stage.	254	— motor control.	208
Curie point.	237		
Current, anode	5		

	page		page
Electronic regulator	64	Gate circuit	107, 113
— smoke detector	93	Glow discharge	68
— switch	112	Glueing, plywood	249
— timer	197	Grid	5
— welding contactor	186	— bias	11
Electron gun	76	—, control	34
Electron lens	76	— control	139
Emission, field	46	— emission	44
Emission, grid	44	—, screen	8, 31
—, secondary	9	—, suppressor	9
Enneode	260	Gun, electron	76
Equipment, battery-charging	144	Half-wave rectification	17
—, three-phase welding	204	Hardening cog-wheels	243
—, welding	51	Hard-starting	58
Examination of food cans	94	Harmonics	12
Excitation, field	208	Heat control	193
Excitron	59	Heater voltage	17
Extinguishing angle	132	Heating, capacitive	245
Extinguishing angle	21	—, inductive	236
Factor, amplification	8	Heat, specific	246
—, duty	55	Hold time	122, 190
—, phase	26	Horizontal control	35, 140
—, power	181	Hysteresis loss	236
—, quality	250	$I_a - V_a$ characteristic	68
Feed-back	10	$I_a - V_a$ diagram	6
Feeler	226	$I_a - V_a$ diagram	5
Field emission	46	Igniter, capacitive	46
— excitation	208	Igniting characteristic	34
— regulation	210	Ignition voltage	19, 59
Filter, colour	86	Ignitron, rectifier	155
—, smoothing	18	Ignitrons	50
Firing angle	21, 35, 130	Impedance, surge	251
— circuit	49	Inductive heating	236
Flash-tube	48	Inductor coil	242
Flip-flop multivibrator circuit	98	Inert gas	31
Fluorescent lamps, controlling	158	Interference	163
Flux, magnetic	208	Intermittent light source	87
Flyback	82	Internal resistance	7
Food cans, examination of	94	Intervals, timer giving two successive	119
Force, counter-electromotive	20	Inverse voltage	26, 33, 129, 281
—, electromotive	208	Inverter action	176, 222
Formula, Thomson	10	—, two-phase	276
Four-phase rectifier	144	IR compensation	213
Four-terminal network	250	Iron, ultrasonic soldering	269
Four-time weld timer	123	kVA demand	54
Frequency changer	206	Lag, angle of	36
— doubler	256	Lamps, dimming of	157
— generator, variable	179	Lateral register control	173
—, resonant	10, 179	Leakage, air	57
Full-wave rectification	18	Lens, electron	76
Function, work	67	Light-controlled relay	69, 73
Fundamental wave	13	Line, load	7
Gain	7	—, quarter-wave	251
Gas amplification	68	—, transmission	251
— clean-up	27, 57	Load circuit tuning device	258
Gas-filled photocell	68	— coil	236
Gas, inert	31		
— triode	30, 70		

	page		page
Load line	7	Phase angle	44
— resistance	17	— delay	56
Loss angle	246	— factor	26
—, hysteresis	236	— -shifting network	39, 140
Magnetic flux	208	— -shifting transformer	155
Magnetisation, remanent	198	Photocell	65
Magnetostriction	270	—, gas-filled	68
Matching transformer	242	Photo-electric relay	85
Measuring shutter speed	107	Photo-emissive cathode	65
Mercury pool cathode	50	Photographic enlarger, timer for	117
Monostable multivibrator	104	— timer, single-stroke	120
Motor control, electronic	208	Plywood, glueing	249
—, correcting	171	Point, Curie	237
—, d.c. shunt-wound	208	—, working	7, 12
—, servo-	173	Power, apparent	128
Multi-motor drive	219	— dissipation	15
Multiple stabiliser circuit	63	—, driving	12
Multiplying circuit	112	— factor	181
Multivibrator, bi-stable	113	—, nominal	129
—, monostable	104	—, output	128
Mutual conductance	223	Pre-amplifier	70, 87
NEMA weld timer circuit	122	Precipitators, electronic dust	278
Network, four-terminal	250	Predetermined counter	109
—, phase-shifting	39, 140	Pressure roll	220
—, <i>RL</i>	45	—, vapour	52
Neutralisation	256	Protection, overload	233
No-load condition	15	Protective resistor	41
Nominal power	129	Pulsation welding	201
Operating conditions, over-excited	241	Pulse shaper	105
— voltage	60	Quality	240
Operation, class C	11	— factor	250
—, single stroke	116	Quarter-wave line	251
Optimum anode resistance	241	Quenching spray	243
Oscillation, saw-tooth	30, 63	Quick reverse	233
—, self-excited	11	Range, control	209
Oscillator, blocking	275	Ratio, transformer	128
—, calibration	275	—, voltage	21, 132
—, Colpitts	240	<i>RC</i> oscillator	177
—, <i>RC</i>	177	Recording device	115
—, relaxation	48, 120	Rectification, controlled	137
Oscillatory circuit	9, 87, 240	—, — two-phase half-wave	130
Oscilloscope, cathode-ray	79	—, full-wave	18
Output power	128	—, half-wave	17
Over-excited operating conditions	241	—, three-phase	135
Overload protection	233	—, two-phase half-wave	127
Oxyde-coated cathode	19	Rectifier, auxiliary	187
Paper register control	168	—, controlled	42
Peak current	129	—, four-phase	144
Peaking transformer	35	—, ignitron	155
Penetration	236	—, single-phase	17
Penny	279	—, transformerless 3-phase	151
Pentode	5	Rectifying tube	17
Pentode, variable- μ	223	Recuperation	228, 233
Period, commutation	26	Reduction circuit	95
Permeability	147, 237	Register control	170
Persistent screen	79	Regulation, field	210
		—, temperature	183
		Regulator, electronic	64

	page		page
Relaxation oscillator	48, 120	Speed, measuring shutter	107
Relay- bi-metal	144	— stabilisation	213
—, light-controlled	69, 73	Speed/torque characteristic	211
—, photo-electric	85	Spot, cathode	46
Reliability	282	Spot-welder, timer for small electric .	121
Remanent magnetisation	198	Spot-welding	47
Reservoir capacitor	17	Spray, quenching	243
Reset time	104	Squeeze time	122, 190
Resistance, a.c.	60	Stabilisation	266
—, cold	58	—, speed	213
—, internal	7	Stabilised alternating voltage	268
—, load	17	— supply unit	267
—, optimum anode	241	Stabiliser circuit, multiple	63
—, welding	47	Stage, amplifying	7
— welding, control of	185	— lighting circuit	166
Resistivity, volume	237	Starter	71
Resistor, current limiting	143	Starting, hard-	58
—, protective	41	Strontium	5
Resonant frequency	10, 179	Subtracting	115
Resonator, coaxial	266	Supervisory circuit	91
Response, spectral	66, 79	Supply unit, stabilised	267
Reversal	221	Suppressor grid	9
Reverse, quick	233	Surge impedance	251
Reversing control	94	Switch, electronic	112
Revolution counter	271	Synchronisation	81
Ribbon-shaped electron beam	102	Synchronous timing	185
Ring counter circuit	74, 96		
Ripple	267	Tacho-dynamo	212
— voltage	18, 154	Tank circuit	9
RL networks	45	Temperature control	180
Roll, pressure	220	— regulation	183
		Tetrode	8, 86
Saturable core reactor	146	Thermostat	92
Saturation current	241	Thomson formula	10
Saw-tooth oscillation	30, 63	Three-phase rectification	135
— voltage	76, 81	Three-phase welding equipment	204
Scale-of-two circuit	96	Thyratron	29
Scaler	95	Time, averaging	30, 54
Scott transformer	144	— base	78
Screen grid	8, 31	—, commutation	127
—, persistent	79	—, conducting	23, 42
Seam welding, control of	193	— constant	116
Secondary emission	9	—, de-ionisation	30
Self-excited oscillation	11	— delay	86
Senditron	46	— delay circuit	91
Sensitivity	67	—, hold	122, 190
Servo-motor	173	—, reset	104
Shaper, pulse	105	Timer	48, 116
Shutter speed, measuring	107	—, continuously acting	120
Single-phase rectifier	17	—, electronic	197
— -stroke operation	116	Timer for photographic enlarger	117
— -stroke photographic timer	120	— — small electric spot-welder	121
Skin effect	236	—, four-time weld	123
Slope	6	— giving two successive intervals	119
Smoothing filter	18	—, single stroke photographic	120
Soldering iron, ultrasonic	269	Time, squeeze	122, 190
Source, intermittent light	87	—, weld	122, 190
Space charge	19	Timing, synchronous	185
Specific heat	246	Torque	209
Spectral response	66, 79	Transducer	270

Index

	page		page
Transformer, current	213, 262	Voltage, arc	20, 51
Transformerless 3-phase rectifier	151	—, armature	209
Transformer, matching	242	—, constant output	144
—, peaking	35	—, doubler circuit	280
—, phase-shifting	155	—, heater	17
— ratio	128	—, ignition	19, 59
—, Scott	144	—, inverse	26, 33, 129, 281
Transmission line	251	—, operating	60
Transmitting tube	5	— ratio	21, 132
Trapezoidal distortion	78	— reference tube	59
Trigger tube	71	—, ripple	18, 154
Triode	5	—, saw-tooth	76, 81
—, gas	30, 70	—, stabilised alternating	268
Two-phase battery charger	142	— stabilising tube	59
— half-wave rectification	127	Volume resistivity	237
— inverter	276	Ward-Leonard	233
Tube, amplifying	5	Warming-up	33
—, cathode-ray	75	Water, cooling	52
—, decade counter	101	Wave, fundamental	13
—, flash	48	Welding contactor, electronic	186
—, rectifying	17	— cycle	193
Tubes connected "back-to-back"	38	— equipment	51
Tube, transmitting	5	— —, three-phase	204
—, trigger	71	—, pulsation	201
—, voltage reference	59	—, resistance	47
—, voltage stabilising	59	—, spot-	47
Tuning device, load circuit	258	Weld time	122, 190
Ultrasonic soldering iron	269	Westinghouse	281
Vapour pressure	52	Work function	67
Variable frequency generator	179	Working point	7, 12
— high-speed drive	177	Zero-point anode	149
— mu pentode	223	Zig-zag connection	216
Vertical control	37, 139	Zone, dead	173

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