Magnetic-Amplifier Circuits

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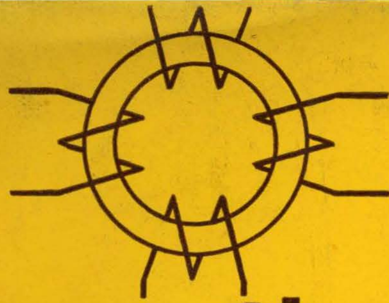

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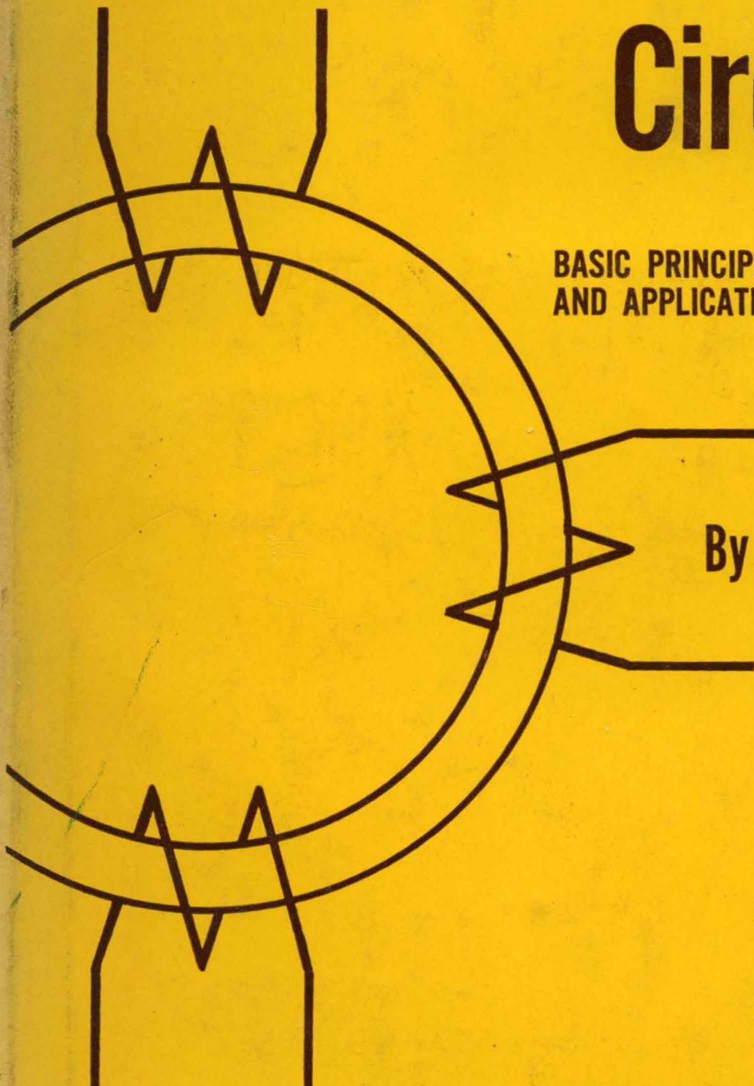
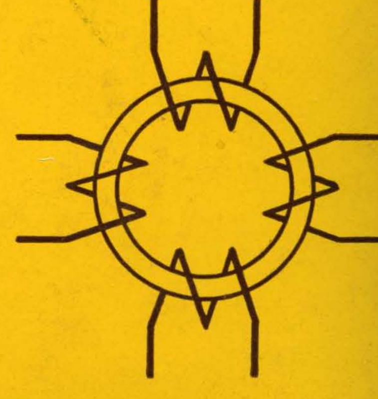


BASIC PRINCIPLES, CHARACTERISTICS, AND APPLICATIONS

By William A. Geyger



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BASIC PRINCIPLES
CHARACTERISTICS
AND APPLICATIONS

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Warning

A considerable portion of the circuit arrangements, constructions, and combinations illustrated and described in this book are covered by letters patent. Designers, manufacturers, and commercial institutions are warned to check the patent situation thoroughly before incorporating or exploiting any of the embodiments described herewith into their designs, constructions, or manufacturing programs.

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This book is dedicated to Harold B. Rex

Preface

Everything in the world may be observed, judged, and classified from different standpoints. Recognition of this fact is particularly important for someone who plans to write a good technical book; and the author, when outlining his approach to the subject of magnetic amplifiers, was fully aware of the many problems involved by his acceptance of this reality.

Since their inception at the beginning of this century, saturable-core reactor devices have proved to be very useful components for numerous fields of electrical engineering. Originally used to control theater-lighting loads, to modulate high-frequency currents in radio transmitters, and to measure large direct currents, they have now developed into extremely valuable arrangements which are applied, for example, as magnetic servo amplifiers, magnetometers, and special components of electronic devices.

During the past fifty years a great deal of information on saturable-core reactor devices and magnetic amplifiers has been accumulated, both in this country and abroad, in the form of numerous patents, scientific papers, and technical reports. There are many theoretical studies on the fundamental mode of operation of the transductor, which represents the basic element of certain groups of magnetic amplifier circuits, and there are also several publications giving some general ideas and interesting results of various experimental investigations. Other published works are devoted to descriptions of particular applications of transductors and magnetic amplifiers.

With regard to the bulk of purely theoretical studies, it should be borne in mind that the basis of operation of magnetic amplifiers is the nonlinear characteristic of the core material and that it is this nonlinearity which causes many difficulties in mathematical analysis. Other difficulties are introduced by the nonlinear characteristics of the dry-disk rectifiers which are generally used in connection with the transductor elements to produce positive feedback. Thus, any detailed analysis requires certain simplifying assumptions regarding the magnetization curve of the core material and the performance characteristics of the rectifier elements. Consequently, because the authors of such publications have been using different assumptions, the results and practical conclusions of different studies appear to be in conflict in many instances; and the reader must appreciate the actual point of view of the particular author.

A review of the literature of experimental investigations and other practical information, as disclosed in American and European patents, shows that numerous valuable contributions, particularly those concerning special magnetic-amplifier circuitry, are written in German, French, or Swedish and that complete files of such technical publications have been available only in some few libraries and in the U. S. Patent Office. Therefore, it is not easy for practitioners in the magnetic-amplifier art to get reliable information about the solution of their special problems from this diversity of sources using different terminologies, showing various kinds of graphical symbols, and sometimes written in foreign languages. Furthermore, the newcomers in this field will find it very hard to get a clear picture of the fundamental principles, performance characteristics, and practical applications of magnetic amplifiers.

This book, designed to be as simple as possible in treatment, is intended as a practical exposition of the fundamental principles and applications of magnetic amplifiers with special reference to magnetic servo amplifiers. It develops logically the various kinds of basic and more complicated circuit arrangements and represents a source of detailed information for engineers, teachers, and students. Because as much emphasis as feasible is placed on experimentally observed phenomena, extended mathematical considerations and cumbersome proofs are avoided. Thus, descriptive and graphical methods are used to give a qualitative and quantitative interpretation of the essential facts.

Chapter 1 gives a classification of saturable-core devices and reviews the history of magnetic amplifiers. Chapter 2 describes magnetic core materials, different kinds of core construction, winding arrangements, and various types of rectifier devices, as used in magnetic-amplifier circuits. In Chapters 3 to 16, the book presents a systematic description of the numerous types of magnetic-amplifier circuits: nonfeedback circuits, single-stage and multistage circuitry utilizing external or internal feedback, and special circuits using critical regeneration or derivative feedback, etc.

Chapters 17 and 18 discuss the technical properties, transient response, and typical applications of magnetic amplifiers, particularly those in instrumentation, in servomechanisms, in regulators and automatic-control devices, and in electronic arrangements.

The references listed at the ends of chapters contain a large number of important European patents which have not been discussed in previous presentations of the magnetic-amplifier art. Using these special references when describing the various types of circuits enables the author to give credit to those investigators who have contributed to the general progress in this field and who have not had the opportunity to publish the results of their pioneer work in technical papers.

The author believes that this book, which contains the newest development in magnetic-amplifier techniques, will be of value to engineers in the solution of their special problems. Newcomers in this field will get an understanding of the fundamental principles and applications of magnetic amplifiers and a foundation for more advanced thinking. Students will be interested to read the book as an introduction to a very interesting field of electrical engineering which has seen most of its growth during the past fifteen years. Finally, those readers interested in teaching or in the interpretation of patents will appreciate the author's standpoint in outlining his treatment of magnetic-amplifier circuitry with special reference to performance characteristics.

The author wishes to express his appreciation to his many friends and associates, both in the U.S. Naval Ordnance Laboratory and elsewhere, for their helpful cooperation in encouraging him to write this book.

William A. Geyger

Contents

Introduction

General Comments on Amplifiers. The American Standards Association definition of an amplifier is " . . . a device for increasing the power associated with a phenomenon without appreciably altering its quality, through control by the amplifier input of a larger amount of power supplied by a local source to the amplifier output." In a simpler way it can be said that an amplifier is a device for reproducing a given signal at an increased intensity by controlling the supply of additional energy from some suitable source.

There is a large variety of devices satisfying this description, and there are various possibilities of proper classification. For example, classification of amplifiers may be based upon whether or not any movable parts are involved in the operating principle. Another classification may stress whether or not any vacuum-tube elements are employed in the fundamental operation.

Of course, it will also be possible in some cases to base classification of amplifiers upon actual performance characteristics and different practical applications, because the question whether or not a certain type of amplifier can meet special technical requirements will be of paramount importance. For example, some types of amplifiers, used in connection with electrical recorders making a permanent record of very small direct voltages created by thermocouples, need not have a high speed of response, since the temperatures to be recorded will change slowly, as a result of the actual operating conditions of a large furnace. On the other hand, some types of servo amplifiers, as used in high-performance servomechanisms, must have a sufficiently high speed of response to avoid hunting or undesired overshooting.

Furthermore, there are problems where certain unique features of the amplifier are required, *e.g.*, special transfer characteristics, inherent constant-current characteristics, dynamic-braking properties, substantially constant gain, action as a d-c/a-c or a-c/d-c converter, low-impedance or high-impedance output, special feedback characteristics, etc. Such features may be obtained by applying modified circuitry or by proper rating of different circuit components.

Whichever form of amplifier is required, its design has to be based upon the various technical requirements of the actual problem to be solved.

Indeed, successful design of amplifiers to be used in the techniques of instrumentation, telemetering, automatic control, and military equipment involves not only the amplifier itself but also requires an understanding of d-c and a-c circuits and electronics, including those branches which deal with the large variety of servomechanisms and automatic-control systems.

The various types of amplifiers may be classified in the following way:

1. Amplifiers having movable parts
   a. Dynamoelectric amplifiers (standard separately excited d-c generator, Amplidyne, Rototrol, Regulex Exciter, and VSA Regulator)
   b. Galvanometer amplifiers (using mechanical devices, bolometer elements, photoelectric cells, variable-induction coil devices, or variable liquid resistors)
2. Amplifiers having no movable parts
   a. Electronic amplifiers (using vacuum tubes, thyratrons, or ignitrons)
   b. Magnetic amplifiers (saturable-reactor devices, saturable-transformer arrangements, and magnetic modulators with negative resistance characteristics)
   c. Semiconductor amplifiers (using transistors)
   d. Dielectric amplifiers (using d-c voltage-controlled barrier-layer devices or electrolytic condensers)

It is to be noted that special combinations of different types of these amplifiers are used in many cases to combine the advantages due to certain inherent properties of the amplifier components.

The dynamoelectric amplifiers[1]* make it possible to apply many useful characteristics of electronic amplifiers to rotating machinery on an economic basis. They are mainly used in connection with closed-loop control circuits. There are certain kinds of current, voltage, and power regulators which involve intermediate electronic amplifiers and control circuits, the final stage being a dynamoelectric amplifier for the purpose of energizing the drive motor.

The galvanometer amplifiers[2] are based upon the common principle that a high-sensitivity moving-coil galvanometer controls periodically acting mechanical elements (*e.g.*, levers with electrical contacts, or a purely mechanical type of servomechanism) or the electric current supply from a d-c or a-c source (*e.g.*, through control of bolometer elements, photoelectric cells, variable-induction coil devices, or variable liquid resistors). Galvanometer amplifiers are extensively applied to the fields of instrumentation, telemetering, and automatic control. Such

* Superscript numbers refer to the references at the end of this introduction.

devices, as in the self-balancing potentiometer recorders, are used in various ways for measuring, recording, and controlling of nonelectric quantities by means of thermocouples, radiation pyrometers, resistance thermometers, and numerous other types of transducers.

The enormous development in electrical engineering which has taken place during the past 30 years has resulted from the introduction of electronic amplifiers, especially in communication techniques and automatic-control equipment. Application of vacuum-tube amplifiers in the field of electrical measurements and control problems made possible an efficient amplification of very small voltages, such as those produced by photoelectric cells, thermocouples, and similar devices. High-speed recording potentiometers and self-balancing bridge circuits using electronic servo amplifiers have been constructed. Various kinds of electronic devices have been used for military purposes, *e.g.*, in gun-control servomechanisms, in automatic aircraft pilots, and as radio aids to air and marine navigation. The Navy, especially, has numerous technical problems in which electronic amplifiers represent a very important part.

However, two serious disadvantages of electronic amplifiers must be considered in practical applications. First, the life of vacuum tubes used in electronic equipment is limited, and, second, shocks or vibrations are dangerous to such tubes. Indeed, perfect operation of important military equipment can be hindered by the sudden failure of a vacuum tube.

Furthermore, it is significant that in the field of d-c amplifier problems the electronic-amplifier tube represents a d-c measuring instrument of a very high resistance, especially suitable for the measurement or amplification of direct voltages in the order of 0.1 to several volts. This method, however, is unsuitable when only very small direct voltages of about 1 to 10 mv are available.

In order to employ the various advantages introduced by suitable use of electronic amplifiers and to avoid, on the other hand, the disadvantages mentioned above, a tubeless amplifier has been developed in different countries, the magnetic amplifier, also called the saturable reactor or transductor. This amplifier consists of electric and saturable magnetic circuits so interlinked that the reactance of an a-c circuit is controlled by an independently applied magnetization.

The magnetic amplifiers belong to the large group of magnetically controlled saturable-reactor devices which are based upon the process of magnetizing a core by a periodically varying magnetomotive force and by a unidirectional magnetomotive force applied independently and simultaneously.

The semiconductor amplifiers using transistors[3,4] are based upon the possibility of controlling the resistances of thin semiconductor layers by the application of electric fields strong enough to penetrate their surfaces. When two point contacts are placed close together on the surface of a small block of germanium and proper bias potentials are applied, the current-voltage relationship of one contact ("collector") can be altered by the passage of current through the other contact ("emitter"). The extent of the interaction between the two contacts is such that it is possible to use the device to amplify electric currents. The transistors, which are still in early-stage development, have been successfully demonstrated in radio-frequency and audio-frequency amplifiers, oscillators, pulse generators, etc.

The dielectric amplifiers[5,6] are based upon the fact that the capacitance of various barrier-layer devices and some electrolytic types of condensers is a function of the applied d-c voltage. Such devices, which preferably are used for the tuning of oscillator circuits, are still in the experimental stage.[7]

Types of Magnetically Controlled Saturable-reactor Devices. It is practical to classify the various types of these devices according to the following scheme:

1. Saturable-reactor devices which are controlled by means of a *constant* magnetomotive force
   a. Using a constant *magnetic field* (frequency transformers and constant-current devices)
   b. Using a constant *electric current* (frequency transformers and constant-current devices)
2. Saturable-reactor devices which are controlled by means of a *variable* magnetomotive force
   a. Using a variable *magnetic field* (flux-gate magnetometers and telemeter arrangements)
   b. Using a variable *electric current* (arrangements for measurement and control of currents and voltages and magnetic amplifiers)

There is obviously a very close linkage between the various types of magnetically controlled saturable-reactor devices, because similar magnetic-core arrangements and similar circuit techniques are used, although the practical applications may be quite different. For example, the essential difference between flux-gate magnetometers and magnetic amplifiers consists merely in the fact that in the first case the saturable-reactor device is controlled by the magnetic field to be measured, whereas in the second case this device is controlled by the electric current or voltage to be amplified. Thus, when outlining a treatise on magnetic-amplifier circuits, it is important to consider this fact, and it will be necessary to include also those arrangements which represent modifications of the original magnetic-amplifier circuits.

The very large number of basic circuits, as used in the techniques of magnetic amplifiers and flux-gate magnetometers, can best be treated according to the differentiation as to whether or not feedback effects are involved in the fundamental principle. Furthermore, it is advisable to base detailed classification upon the different performance characteristics (nonpolarized, polarized, duodirectional circuits), single-stage and multistage arrangements, and special circuits to be used in connection with electronic devices, etc. The wide variety of typical magnetic-amplifier circuits, especially those types of push-pull circuits to be used in connection with high-performance servomechanisms, makes difficult their extensive illustration. Therefore, in the following treatise, each group of these circuits will be described by means of typical examples, suitably selected, with special reference to the original literature. Using this method of presentation and providing simplified diagrams with uniform graphical symbols will make it possible to recognize similar or even equivalent principles and forms involved in those circuits, which might look quite different had symbols and arrangement of the drawings been shown in a different way.

The main objective of this book is to indicate a method of advanced thinking that can give a clear and comparatively simple picture of magnetic-amplifier circuits rather than just to describe the multitude of specific arrangements, as disclosed in the bulk of technical papers, patents, and other publications.

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CHAPTER 1

Historical Development of Magnetic-amplifier Circuits

1.1. General. Reference to the literature shows that considerable effort has been expended in the development of magnetic-amplifier and flux-gate-magnetometer circuits. The comprehensive "Bibliography of Magnetic Amplifier Devices and the Saturable Reactor Art," recently

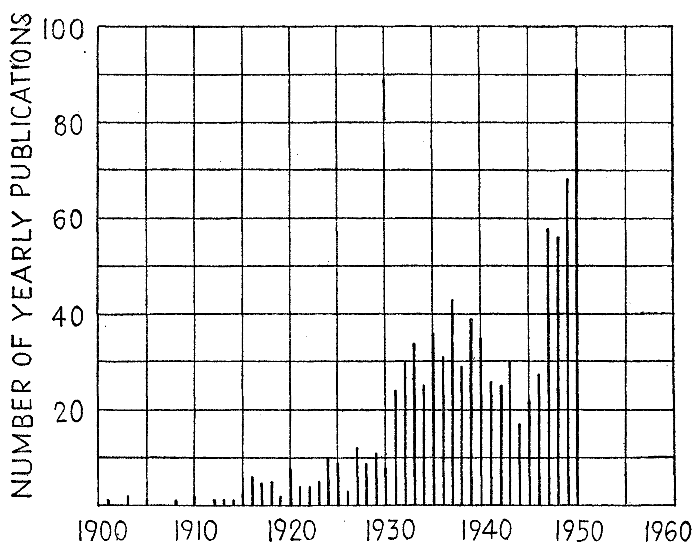

FIG. 1.1. Number of scientific papers, technical reports, and United States patents being published each year during the period from 1900 to mid-1951, corresponding to J. G. Miles's, "Bibliography of Magnetic Amplifier Devices and the Saturable Reactor Art."

presented by J. G. Miles,[1*] makes it evident that the number of scientific papers, technical reports, and United States patents being published each year during the period from 1900 to mid-1951 has steadily increased. A corresponding graphical presentation, as shown in Fig. 1.1, illustrates the recent development in a dramatic way. It is to be noted, however,

* Superscript numbers refer to the references at the end of each chapter.

that Miles's bibliography and the corresponding diagram (Fig. 1.1) do not include the large number of European patents pertaining to this special field of electrical engineering.

1.2. Historical Survey. The history of saturable-reactor devices and magnetic amplifiers is about 50 years old and begins in 1901 when C. F. Burgess and B. Frankenfield[2] disclosed the early use of various forms of d-c-controlled saturable reactors for the "regulation of electric circuits" and described different types of control-circuit decoupling means, as follows: series impedance, three-legged and four-legged cores, hollow annular cores, and magnetic "cross-valve" core arrangements. This disclosure, however, makes no claims regarding amplifying properties.

In 1901, R. A. Fessenden[3] described the basic idea of using a saturable reactor in a "wireless signaling" system as a magnetic modulator in such a way that when the current in the microphone circuit "is modified or changed by speaking into the transmitter, the permeability of the core is correspondingly changed or modified, thereby producing a corresponding change or modification in the self-inductance and a change in the frequency of the natural period of vibration of the sending-conductor . . . ."

At the same time, H. J. Ryan[4] devised a method for the measurement of large direct currents that employs a saturable transformer for comparing the values of large and small direct currents in terms of the ratio of transformation and the reading of the instrument used for measuring the small current. This method is based upon the following principle: Where two coils of similar dimensions carrying direct current are placed side by side and linked with a closed magnetic core, the permeability of the core for alternating magnetic flux will be a maximum when the ampere-turns of the two coils are equal and opposite. By using this null method, it was possible to measure large direct currents in the range from 100 to 1,000 amp with an accuracy of about ± 0.5 per cent.

In 1913 an important advance was made by E. Besag,[5] who introduced the combination of a twin-type saturable-reactor arrangement with direct-indicating instruments and recorders for measuring large direct currents by means of the magnetomotive force around the d-c conductor, without using giant-size shunt resistors, which are very expensive and difficult to cool. The a-c windings of the two saturable reactors are connected in series and to an a-c power supply, and the alternating current flowing through these windings is measured with an ammeter, which is calibrated in d-c units, to indicate directly the direct current to be measured.

Very important development started in the period 1912 to 1918 when E. F. W. Alexanderson[6] devised a method of modulating high-frequency alternating currents from one of his alternators so that its current could be used for transatlantic radio telephony. The corresponding disclosures comprise:

1. Early illustration of external-feedback means
2. Control-circuit decoupling means employing two cores, also four-legged and eight-legged cores
3. Early claim of off-resonant principle for increased gain
4. Frequency-discriminating means in a-c power-supply and control circuits
5. Early illustration of cascaded magnetic-amplifier stages
6. Means for utilizing a single winding for the combined functions of saturation control, d-c bias, and as the reactance winding of a d-c saturable reactor

Independently, in 1913, special forms of magnetic modulators were developed in Germany by L. Kühn,[7] M. Osnos,[8] and L. Pungs.[9] In 1915, G. W. Elmen[10] described (1) an early Wheatstone-bridge circuit employing a signal-controlled saturable reactor in one arm, (2) novel control means by controlling the amount of agitating magnetomotive force to obtain variable hysteresis-effect reduction, and (3) a hollow annular magnetic core for circuit decoupling. In 1916 and 1917, R. V. L. Hartley[11] disclosed a fundamental push-pull magnetic modulator and a d-c saturable transformer with ultra-carrier-frequency magnetomotive-force means for hysteresis-effect reduction.

An early full-wave d-c load saturable-reactor circuit with internal feedback (self-saturating circuit) and control by d-c saturating windings was described in 1919 by J. Jonas.[12] In 1928 to 1930, various forms of magnetic amplifiers with external-feedback circuits and some types of balanced circuits (zero output for zero control signal) were disclosed by P. H. Dowling.[13] The fundamental negative-resistance magnetic amplifier was introduced in different forms by K. Heegner[14] (1925) and E. Peterson[15] (1930).

In 1930, A. J. Sorensen[16] and P. H. Dowling[17] introduced (1) an early Wheatstone-bridge-type magnetic amplifier (the four reactance arms of the bridge are the four reactance windings on two three-legged reactors) and (2) a group of interesting push-pull (balanced) magnetic-amplifier circuits.

Reference to Fig. 1.1 shows that general activity in the field of magnetic amplifiers and the saturable-reactor art has greatly increased during the past 20 years. It is, of course, impossible to give acknowledgment to all the investigators who have contributed to the enormous progress which has been achieved during this period. For example, much of the extensive German development work in the fields of flux-gate magnetometers and magnetic servo amplifiers for various types of military equipment appeared originally in restricted documents. Some of it has since been published in European patents, and some has not.

The recent historical development is considered to be the practical result of combined efforts and achievements in the following fields of electrical engineering:

1. The introduction of high-permeability nickel-iron-alloy core materials
2. The possibility of manufacturing efficient and reliable dry-disk rectifiers
3. The development of special push-pull feedback circuits for flux-gate magnetometers and magnetic servo amplifiers

The excellent results which have been obtained by the application of such core materials[18] in the development of high-performance magnetic amplifiers and similar devices can rightly be described as revolutionary. The considerable improvement in the manufacturing process of dry-disk rectifiers was also of paramount importance for this development, because it is the combination of the saturable reactor and the rectifier that leads to perfect solutions of numerous technical problems. Referring to the development of special circuits for magnetic amplifiers and flux-gate magnetometers, it can be said that their mechanism is in many cases similar or even analogous to those circuits which are successfully used in the fields of tube voltmeters, electronic servo amplifiers, and some special types of automatic-control devices.

Since 1930, this development has been greatly increased, both in the United States and abroad. Some of the most important advances will be characterized in the following historical survey comprising the period from 1930 up to date.

A. Boyajian[19] and C. G. Suits[20] disclosed various improvements (nonlinear novel external-feedback circuits, etc.). P. Thomas[21] and A. S. FitzGerald[22] devised various single-stage and multistage types of magnetic amplifiers (neutral type and push-pull type) to be used in connection with relay devices or reversible-motor arrangements. F. G. Logan[23] and M. A. Edwards[24] disclosed various types of self-saturating circuits with a-c or d-c load. E. T. Burton[25] and C. W. La Pierre[26] described several "second-harmonic" types of magnetic amplifiers and flip-flop circuits. A. L. Whiteley and L. C. Ludbrook[27] described several half-wave self-saturating push-pull circuits to be applied with relays or thyratrons.

In 1937, W. Krämer[28] revealed that properly rated saturable reactors with nickel-iron-alloy cores, preferably with rectangular-hysteresis-loop core material, have inherent current-transformer characteristics and that the ratio between the "primary" direct current to be measured and the average value of the rectified "secondary" current is equal to the ratio between the secondary and primary number of turns on the cores. The current-transformer character of this arrangement is to a certain extent independent of variations in magnitude, frequency, and waveshape of the a-c source. Furthermore, it has been proved by Krämer[29] that by suitably designing the arrangement, the rectified average value of the alternating current is not only proportional to the direct current but to some extent will even have a similar oscillographic variation when the direct current varies, owing to a superimposed a-c component.

These arrangements, which are primarily intended for measuring large direct currents in the range from 100 to 10,000 amp, may equally well be used for measuring high direct voltages (W. Krämer, 1938).[30] The only difference will be that the direct voltage to be measured must first be translated, by means of a series resistor, into a direct current proportional to the voltage. This current will be of a magnitude which necessitates employing d-c windings with a large number of turns.

Krämer's fundamental development work attracted the attention of numerous investigators, resulting in further research and improvement. Pioneer work on saturable-reactor arrangements, without or with feedback, was carried out by T. Buchhold,[31] A. U. Lamm,[32,33] S. E. Hedström,[34] B. Nordfeldt,[35] K. M. H. Forssell,[36] and U. Krabbe.[37] In the Scandinavian countries the rather long expression "saturable reactor" has been replaced by the word "transductor." Many techniques involving special combinations of saturable reactors and dry-disk rectifiers for supplying controllable d-c power from single-, three-, and six-phase networks in place of the conventional mercury-arc-rectifier systems have been developed by Allmänna Svenska Elektriska Aktiebolaget (ASEA) personnel. These techniques are particularly suitable for installations with large power outputs.

The extensive development of magnetic-amplifier techniques in Germany during the period from 1937 to 1945 was based upon the fact that there are certain military applications for amplifiers of electric signals which require that these amplifiers possess unusual durability, simplicity, and reliability. Pioneer work on the single-phase push-pull type of saturable-reactor circuit as a magnetic amplifier for various servo applications (*e.g.*, in autopilots, blind-approach systems, remote-control positional servomechanisms, and computer circuits) and as a flux-gate magnetometer was carried out by G. Barth[38] and his staff. These devices were designed to operate at ambient temperatures between +70 and −60°C for supply voltages of 30 to 40 volts and frequency ranges of 400 to 600 cps. Both naval and air-force equipment was produced by Siemens Apparate & Maschinen (SAM) personnel. Of course, disclosures concerning this wartime development were limited to internal reports, and no theoretical design principles were published. This restriction of German publications in the field of magnetic amplifiers led some to believe that, although excellent results were obtained, design and development progressed almost entirely by trial and error.[39]

In the author's experience, the magnetic amplifier made its first appearance in special instrumentation work for replacing the mechanical d-c/a-c converter and the electronic amplifier, as used in high-speed recording potentiometers, by a high-performance magnetic servo amplifier operated from a 50-cps power supply. This magnetic amplifier of the balance-detector type was designed to have duodirectional transfer characteristics, high sensitivity (large power gain), and an extremely small drift rate.[40] The output power of this push-pull-type amplifier was sufficient for the control of a small reversible motor, which may be either a separately excited a-c motor of the induction-meter type or a d-c motor of the moving-coil-meter type. According to different special problems concerning d-c potentiometers or d-c bridge networks making available different d-c input-power values corresponding to full-scale deflection of the recording mechanism, two different types of magnetic servo amplifiers have been developed, which are described in several papers.[40]

Early in 1944, C. S. Hudson[41–43] initiated work on magnetic amplifiers at the Royal Aircraft Establishment, which has been concerned mainly with amplifiers designed to operate from a 400-cps power supply. This work and detailed examination of German designs[41] may have reawakened interest in magnetic amplifiers in England.

Theoretical and experimental studies of the series-connected magnetic amplifier with external feedback have been published by S. E. Tweedy,[44,45] A. G. Milnes,[46,47] and by H. M. Gale and P. D. Atkinson.[48] Furthermore, several studies in this field have been presented by E. H. Frost-Smith,[49,50] J. H. Reyner,[51] and F. E. Butcher and R. Willheim.[52]

One reason for the increased interest in magnetic amplifiers in this country was the successful German development work for various military applications, especially for naval fire-control systems, as used on the German heavy cruiser "Prinz Eugen."[53] Reference to Fig. 1.1 shows that the activity in the United States has rapidly increased during the past six years. The results of recent theoretical and experimental investigations on magnetic amplifiers have been presented in numerous technical papers and United States patents. Many of the contributors are specifically mentioned in later chapters.

Finally, it is interesting to note that standardizing groups of the American Institute of Electrical Engineers are now working on definitions and methods of expressing the performance characteristics of magnetic amplifiers.[54]

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CHAPTER 2

Magnetic-amplifier Elements

2.1. Introduction. Generally, magnetic amplifiers comprise combinations of saturable reactors, dry-disk rectifiers, fixed and adjustable resistors, and conventional-type transformers. The fact that no moving parts and no vacuum-tube elements are involved is one of the most important advantages of these amplifiers. In special cases, however, it is practical to replace the dry-disk rectifiers by vacuum tubes (diodes) or mechanical devices with rotating or vibrating contacts.

An obvious advantage of magnetic amplification is its freedom from deterioration, since there is no part that cannot be so designed as to be practically everlasting. Magnetic amplifiers can therefore be contained in a sealed case, a practice employed in connection with certain automatic regulators of the ordinary electromagnetic pattern.

The performance of magnetic amplifiers depends on the properties of the components which are used. Therefore, it is necessary to describe briefly the elements of these amplifiers and to illustrate the importance of different factors which are involved in their practical application in various circuits.

This chapter gives at first a summary of the fundamental considerations concerning the construction of saturable reactors, which are closely linked with the properties of special magnetic core materials. Then, some considerations of the explicit characteristics desired of dry-disk rectifiers used in magnetic-amplifier circuits, particularly those utilizing external or internal feedback, will be taken up.

2.2. Core Materials for Saturable Reactors. For optimum performance, the core material should have the following magnetic properties:

1. Hysteresis and eddy-current losses a minimum (high resistivity, low coercive force, and ability to be made in thin laminations or tapes)
2. High saturation flux density, to obtain large power-handling capacity of a given weight of core material
3. The general shape of the magnetization curve as close as possible to the rectangular hysteresis loop, as shown in Fig. 2.1

4. Stability of magnetic characteristics under changing temperature and mechanical strain and shock conditions

Additionally, with regard to the manufacturing process of magnetic amplifiers, the core material should have properties which are reproducible as between different batches and are the same for all the material in each batch.

Of the various types of commercially available core materials which may be used for saturable-reactor devices, nickel-iron alloys, particu-

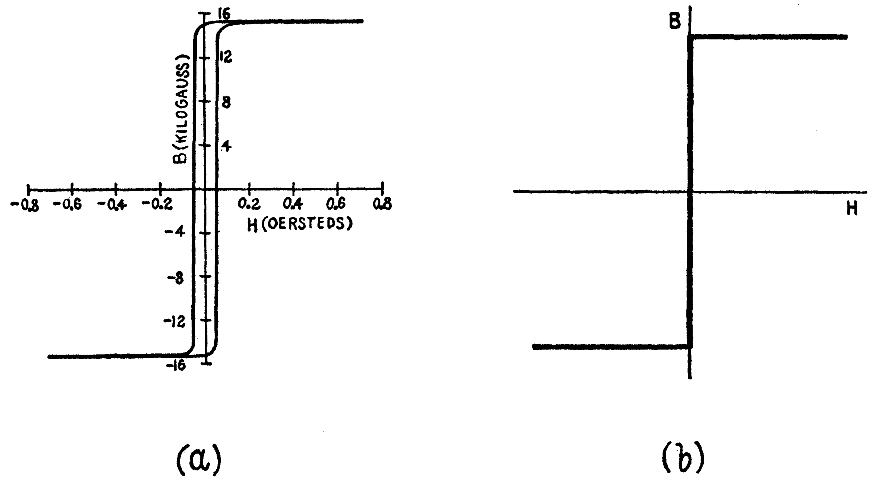

FIG. 2.1. Rectangular hysteresis loops: (*a*) actual magnetization curve of Orthonol; (*b*) idealized magnetization curve.

larly the grain-oriented rectangular-hysteresis-loop materials (Permenorm 5000-Z alloy, Orthonol, Deltamax, etc.), have the most suitable properties. Cores of cold-rolled silicon-steel tape may also be suitable in many cases, especially for higher power outputs (100 watts and above at 60 cps). It is to be noted, however, that the saturation-flux density values of the high-permeability materials (Permalloy C, Mumetal, 1040 alloy, etc.), as extensively used in low-level input-stage circuits, are much lower than those of the rectangular-hysteresis-loop materials, which have proved to be most suitable for output-stage circuits and in which maximum permeability occurs close to saturation flux density, thus increasing substantially the power-handling capacity of a given weight of core material.

2.3. Core Construction of Saturable Reactors. For optimum performance, the structure of magnetic-amplifier cores should have the following characteristics:

1. Effective air gap as small as possible, in order not to shear over the hysteresis loop
2. Spirally wound tape-core construction with loose-fitting insulated container to support the multilayer toroidal windings, when grain-oriented rectangular-hysteresis-loop materials are used
3. Very thin laminations or tape (*e.g.*, 0.001 to 0.003 in. only, or even less), to reduce eddy-current effects to a minimum

At first, it is important to stress that the usual interleaving technique adopted in conventional transformer-core construction, with a stack of thin laminations of the required form, is not adequate for saturable reactors in magnetic amplifiers, owing to the presence of air gaps at the joints. Staggering of the joints (Fig. 2.2*a*) does not overcome the fundamental difficulty, because the effective air gap is still appreciable. Evidently, some of the magnetic flux still crosses the air gap, while at high flux densities the concentration of flux occurring in the isthmus portion of the core introduces a double knee in the magnetization curve, thus reducing the sharpness of the bend in the magnetization curve and also increasing the leakage flux. Furthermore, when such a core construction, with staggered butt joints between the laminations, is used, the effect of these air gaps depends on their actual size. Thus, consistent results with similar cores depend on great care being taken to ensure uniformity in assembly.

Considerable improvement over the characteristics of the butt-joint assembly can be obtained by overlapping the laminations over a substantial part of the magnetic path. U-type laminations, as shown in Fig. 2.2*b*, have been used in some German magnetic amplifiers to maintain a uniform cross-sectional area of magnetic material throughout the core. Similar C- and E-type laminations have been standardized recently in this country[1,2] and in England.[3] The main advantages of this "sideways-air-gap" core are that the isthmus is removed, allowing the full core area to be used to saturation, and that the windings can be applied in the usual way, without using toroidal winding machines or other special winding procedures which are based upon the use of split bobbins.

An obvious way of overcoming difficulties due to the presence of air gaps in laminated types of core construction is to employ single-piece stampings (Fig. 2.2*c*) and a winding method in which a split bobbin with a split gear is placed on the core, the bobbin is rotated, and the wire is wound onto the bobbin, after which the temporary gear is taken off. Such cores, however, do not give the possibility of taking full advantage of the special magnetic properties of grain-oriented core materials.

Because of difficulties encountered in handling punchings of very thin material, spirally wound tape cores will be preferred in many cases.[4] It is to be noted that the magnetizing force should be uniform over the entire cross section of the core, to obtain a sharp knee in the magnetiza-

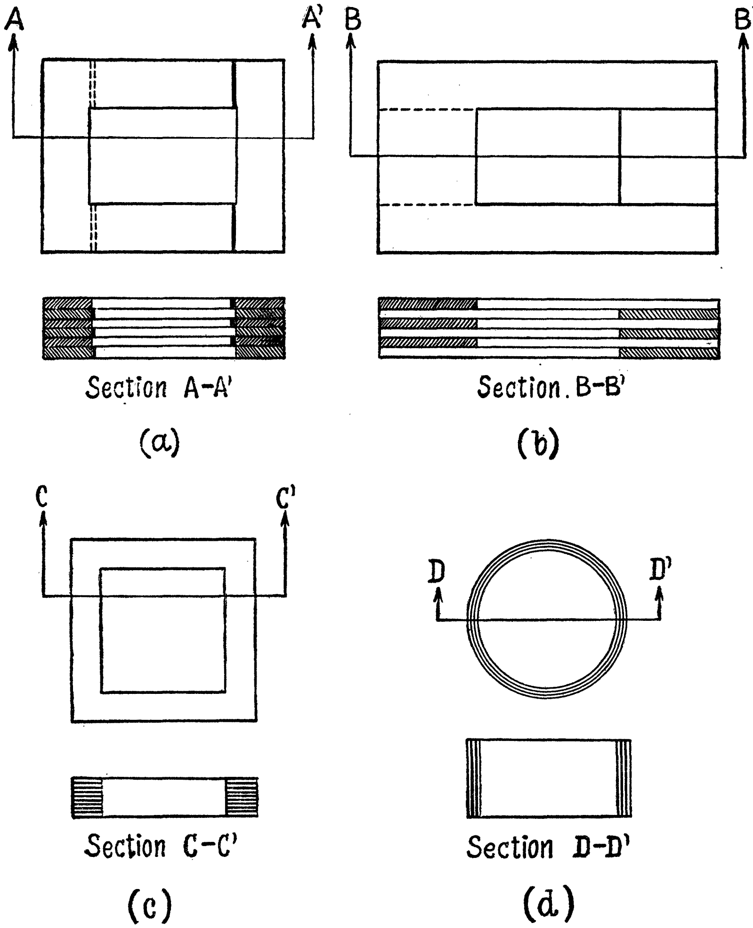

FIG. 2.2. Various core constructions of saturable reactors: (*a*) conventional transformer-core construction with staggered butt joints between the laminations; (*b*) the "sideways-air-gap" core with U-type laminations which are overlapping over a substantial part of the magnetic path; (*c*) laminated type of rectangular core construction having single-piece stampings without an air gap; (*d*) spirally wound tape core with a relatively narrow annular width.

tion curve. This implies a relatively narrow annular width (Fig. 2.2*d*). Precaution should also be taken to prevent any mechanical strains after the final heat-treatment, as this is known to change the magnetic characteristic considerably. The tape cores therefore are mounted in loose-fitting, insulated containers to support the multilayer toroidal windings. These windings are applied by using toroidal winding machines. The tape cores may have the same forms as the afore-mentioned laminated cores which are composed of single-piece stampings (Fig. 2.2c), and the special winding procedure, which uses split bobbins, may be applied, when rectangular forms of tape cores are provided.

2.4. Arrangement of Windings. The performance of magnetically controlled saturable-reactor devices depends to a considerable extent on the arrangement of the various windings operating as load windings, control windings, bias windings, feedback windings, or as additional windings for special purposes. It is generally desirable that all the windings be placed on the core construction in such a way that leakage effects are reduced to a minimum. Another most important requirement consists in preventing the circulation of disturbing alternating currents of fundamental frequency in d-c control, bias, and feedback circuits. Such currents may be induced in these circuits by transformer action from a-c components of currents flowing through the load windings.

Circulation of fundamental-frequency currents in the control circuit can be prevented in two ways:

1. Connecting a sufficiently high impedance in series with the control winding so that the fundamental-frequency voltage induced from the load winding into the control winding is unable to produce excessively large currents in the control-circuit loop
2. Using a special arrangement of windings so that no net voltage of fundamental frequency appears across the control-circuit terminals

The first method is extensively used in those types of very sensitive magnetic-amplifier circuits which utilize external or internal feedback, so that the ratio N_C/N_L between the number of turns N_C of the control windings and the number of turns N_L of the load windings can be made very small (N_C/N_L about 0.01 to 0.1). This method is particularly useful when the control windings of the magnetic amplifier are supplied with pure alternating current or unidirectional current (*e.g.*, half-cycle pulses), and when special push-pull circuits having only two saturable-reactor elements are used.

Referring to the second method, it is to be noted that there are two possibilities for designing the arrangement of windings so that no net voltage of fundamental frequency will appear at the terminals of the control circuit:

1. Providing two separate saturable-reactor elements, for example, with series-aiding-connected d-c control windings N'_C, N''_C and series-opposing-connected a-c load windings N'_L, N''_L, as shown in Fig.

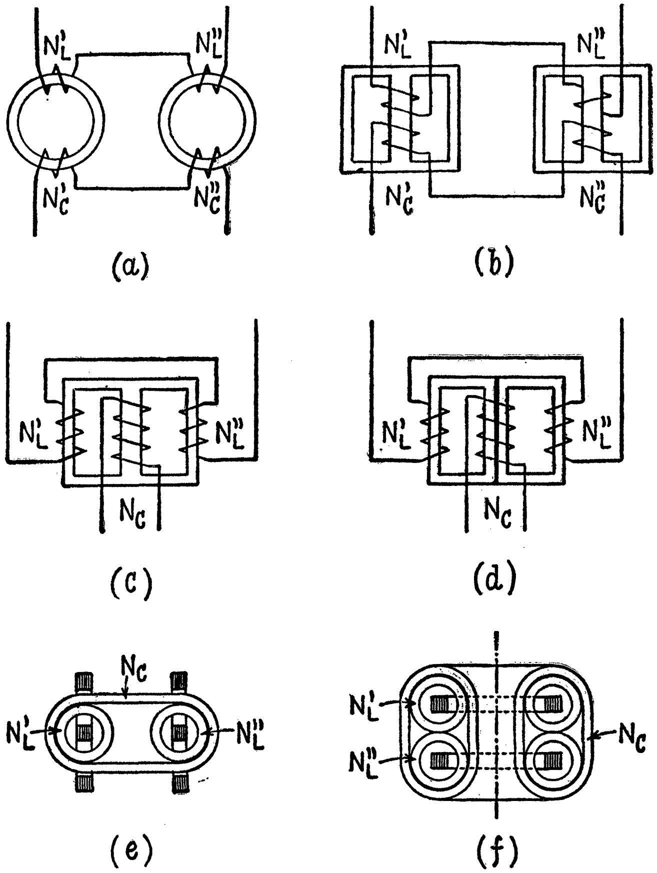

FIG. 2.3. Various arrangements of windings on saturable reactors: (*a*) two separate ring-core saturable-reactor elements; (*b*) two separate three-legged-core saturable-reactor elements; (*c*) a three-legged-core saturable-reactor element, the center leg of which is not split. This arrangement may cause disturbing hysteresis effects because there is substantially no a-c flux flowing through the center leg of the core; (*d*) a three-legged-core saturable-reactor element, the center leg of which is split lengthwise to eliminate disturbing hysteresis effects; (*e*) two separate three-legged cores having one common d-c control winding (N_C) linking both center legs and their a-c load windings (N'_L and N''_L); (*f*) toroidal arrangement with one common d-c control winding (N_C) linking both tape cores and their a-c load windings (N'_L and N''_L).

2.3*a* and *b*. The two fundamental-frequency voltages which are induced in the control windings N'_C, N''_C cancel each other out, and there is no resultant voltage of fundamental frequency in the d-c control circuit. However, it is important to note that actual voltage across each individual winding will be very large (of the order of 1,000 volts) when the turn ratio $N'_C/N'_L = N''_C/N''_L$ is large, necessitating a high degree of insulation and, therefore, requiring comparatively large space occupied by the insulating material. This arrangement will successfully be used in connection with magnetic-amplifier feedback circuits having a small turn ratio N_C/N_L, as described in later chapters.

2. Providing one of the single-core or twin-core arrangements, as shown in Fig. 2.3*c* to *f*. Each of these arrangements has only one d-c control winding, which embraces both magnetic circuits, so that there is no resultant a-c flux through this winding and no net voltage of fundamental frequency can appear across the control-circuit terminals.

Referring to the three-legged core with one d-c control winding on the center leg and two parts of the a-c load windings on the other legs (Fig. 2.3*c* and *d*), it is necessary that the center leg be split lengthwise (Fig. 2.3*d*), to eliminate disturbing hysteresis effects. If the center leg of the core is not split (Fig. 2.3*c*), then the a-c flux components produced by the load windings N'_L and N''_L tend to flow around the circumference of the core and not to flow through the center leg at all. Thus, only unidirectional flux will flow through the center leg, and this may cause disturbing hysteresis effects, particularly when such an arrangement is used for amplification of very feeble input signals. By splitting the center leg lengthwise (Fig. 2.3*d*), a narrow air gap is provided between the two cores, which prevents the flux from following the circumference path. Therefore, alternating flux components as well as unidirectional flux components must flow through the center leg of the core, and disturbing hysteresis effects will be eliminated in this way.

The arrangement with two separate three-legged cores (Fig. 2.3*e*) has one common d-c control winding linking both center legs and their a-c load windings. When toroidal arrangements are used, it is sometimes possible to put the a-c load windings on the two separate cores and wind the d-c windings over the two toroids together (Fig. 2.3*f*).

Reference to the literature shows that various kinds of graphical symbols for saturable-reactor elements are used in diagrams of magnetic-amplifier circuits. Figure 2.4*a* to *c* show three different ways for illustrating the basic arrangement of Fig. 2.3*a* and *b*, which has series-aiding-connected d-c control windings N'_C, N''_C, and series-opposing-connected a-c load windings N'_L, N''_L. Other graphical symbols for "transductors," as used in Swedish publications, are shown in Fig. 2.4*d* and *e*. Diagrams with uniform graphical symbols similar to those of Fig. 2.3*a* and *d* will be used in this book for illustrating the various types of magnetic-amplifier circuits.

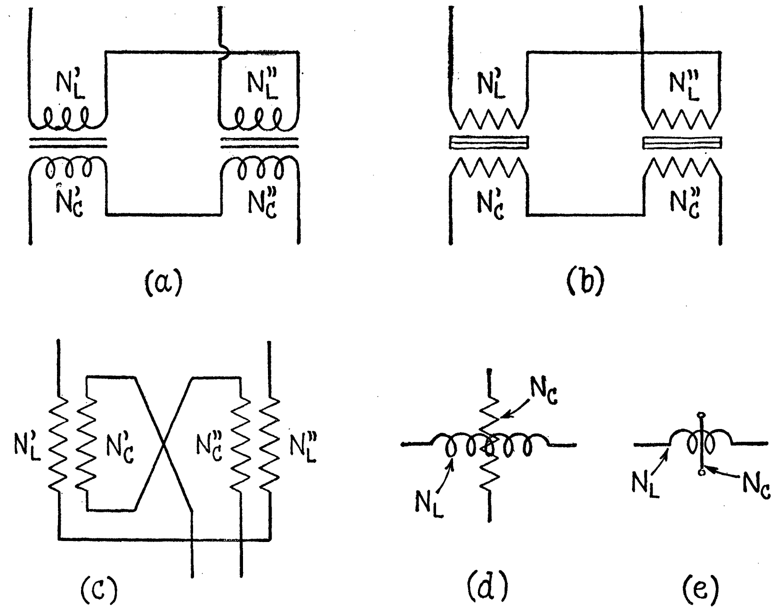

FIG. 2.4. Various kinds of graphical symbols for saturable-reactor elements, as used in the literature concerning magnetic-amplifier circuits.

2.5. Application of Rectifiers. There are different reasons for using rectifiers in magnetic-amplifier circuits:

1. To take advantage of the current-transformer character of saturable reactors, it is necessary that the alternating current flowing through the load windings (N_L' and N_L'' in Fig. 2.3) be rectified so that it becomes of the same form as the direct current which flows through the control windings (N_C' and N_C'' in Fig. 2.3). It has already been mentioned that with suitable design of the arrangement the rectified average value of this alternating current is not only proportional to the direct current but to some extent will even have a similar oscillographic variation when the direct current varies, owing to a superimposed a-c component.
2. Simple arrangements like those shown in Fig. 2.3 represent "nonpolarized" magnetic amplifiers, which are unable to discriminate between positive and negative currents in the d-c control circuit. It is possible to obtain a "polarized" magnetic amplifier, which will be influenced by changes in direction of the control current, by using a d-c bias magnetization, produced preferably by means of an additional rectifier circuit which is supplied from the a-c power supply.
3. In magnetic amplifiers utilizing feedback effects, rectifiers are used for producing a d-c component, proportional to the alternating load

current, which cooperates with the control current. Additional rectifiers may be used in such feedback circuits to bring the "quiescent current" (the load current with no-signal conditions) to the desired value.

The rectifier, therefore, will be considered as an essential component of magnetic amplifiers. The following types of dry-disk rectifiers are commonly used for the afore-mentioned purposes: (1) selenium rectifiers, (2) copper-oxide rectifiers, and (3) germanium rectifiers. In special cases, vacuum-tube rectifiers (diodes) or mechanical rectifiers with rotating or vibrating contacts will be used. Detailed information concerning these types of rectifiers is given in the literature.

The performance of the rectifiers used in magnetic-amplifier circuits, particularly circuits utilizing external or internal feedback, affects the characteristics of the amplifier, because the actual feedback factor is a function of the "current ratio" (ratio between forward current and reverse current) of the rectifier elements in the load circuit. The following explicit characteristics are desired of dry-disk rectifiers for magnetic-amplifier applications:

1. The lower the reverse current of the rectifier, the smaller is the demagnetization of the magnetic core during the nonconducting portion of the cycle. Thus, high inverse resistance is very important.
2. In order to get improved transfer characteristics (having increased slope) of magnetic amplifiers utilizing a comparatively high amount of regenerative (positive) feedback, it is advisable to decrease the voltage drop across each rectifier cell (*e.g.*, 8 to 10 volts instead of 17 to 25 volts). Therefore, low forward resistance is desirable, to make the total forward resistance of the rectifier set (containing, for example, 15 cells) as small as possible.
3. It is desirable that the current ratio of the dry-disk rectifier (the ratio between forward current and reverse current) be as high as possible.
4. The dry-disk rectifiers should have the maximum value of inverse voltage rating consistent with a high value of forward- vs. reverse-current ratio. This is desirable for obtaining both light weight and high sensitivity of the magnetic amplifier.
5. Age and changes in ambient temperature should not considerably alter the electrical properties of the rectifiers.
6. Dry-disk rectifiers for magnetic-amplifier applications should have reproducible characteristics. This is particularly important for small low-level input-stage units, which must have a very small drift rate.
7. Rectifiers for such applications should not change when not in use or, at least, should recover rapidly from nonuse.

8. It is desirable that the two parts of a half-wave three-terminal-type rectifier set be as well matched as possible to get symmetrical forward-resistance conditions in the corresponding two branches of a self-saturating circuit.
9. The capacitance of the rectifier cells should be as small as possible, particularly when higher power-supply frequencies (*e.g.*, 400 cps) are used.

Vacuum-tube rectifiers having an extremely high reverse resistance are very suitable for experimental investigations on self-saturating magnetic-amplifier circuits under zero-reverse-current conditions. Mechanical

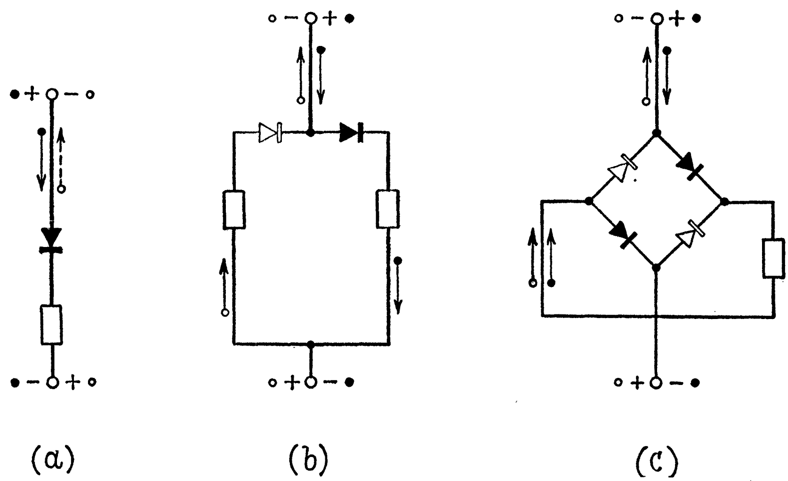

FIG. 2.5. Graphical symbols used in this book for rectifier arrangements in magnetic-amplifier circuits.

rectifiers with rotating or vibrating contacts can be considered as "ideal" rectifiers having zero forward resistance and infinite reverse resistance.

Figure 2.5 shows graphical symbols which will be used in this book for illustrating the mode of operation of rectifier arrangements in magnetic-amplifier circuits.

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2. Catalogue, "Magnetic Engineering Practice," Magnetic Metals Company, Camden, N.J.
3. J. H. Reyner, "The Magnetic Amplifier," pp. 112–113, Stuart & Richards, London, 1950.
4. G. W. Elmen and E. A. Gaugler, "Special Magnetic Alloys and Applications," *Electrical Engineering*, Vol. 67, No. 9, pp. 843–845, 1948. Also E. A. Gaugler, "Method of Producing Grain Oriented Ferro-magnetic Alloys," United States patent 2,569,468, issued October 2, 1951 (application filed June 16, 1948).

CHAPTER 3

Basic Nonfeedback Types of Magnetic-amplifier Circuits

3.1. Introduction. Generally, the various types of single-stage and multistage magnetic-amplifier circuits, as used in connection with high-performance servomechanisms, automatic-control devices, and similar applications, utilize large amounts of positive feedback to obtain high power amplification. All these circuits, however, are derived from some basic saturable-reactor circuits which operate without feedback and represent the basis of the design procedure of feedback arrangements.

Furthermore, there are numerous nonfeedback types of magnetic-amplifier circuits which offer valuable possibilities in the field of electrical measurements, although comparatively low values of power amplification are obtained with such devices. Finally, the nonfeedback circuits have proved to be very useful for testing of magnetic core materials and matching of saturable-reactor twin elements, as used in symmetrical push-pull circuits of magnetic servo amplifiers.

This chapter, therefore, describes the basic saturable-reactor circuit, its performance characteristics, and some typical applications.

3.2. The Basic Saturable-reactor Circuit. In 1901, C. F. Burgess and B. Frankenfield[1] described the basic saturable-reactor circuit, its fundamental mode of operation, and some possibilities of practical application. One of the simplest arrangements (Fig. 3.1*a*) has a single saturable-core reactor with two windings N_L and N_C. The load winding N_L is connected to a source of alternating current, through a load of some sort, represented by the load resistor R_L (*e.g.*, a bank of lamps) carrying the load current I_L. The control winding N_C is supplied with direct current I_C from a battery through a variable resistor R_C and an iron-cored choke coil L_C. This coil presents a very high impedance to alternating current so that the voltage induced from the load winding N_L into the control winding N_C is unable to produce excessively large circulating currents in the control-circuit loop.

Evidently, actual magnitudes of load current I_L and load voltage I_LR_L will be determined by the combined effect of the load resistor R_L and the saturable reactor acting as a d-c-controlled inductive series impedance. The operation of this circuit is based upon the fact that the actual impedance of the saturable reactor is a function of the direct current I_C, which can be controlled by varying the value of the series resistor R_C. If $I_C = 0$ ($R_C = \infty$), then the inductive impedance of the saturable reactor has its maximum value, and the load current I_L will be reduced to the minimum. However, when a direct current I_C is supplied to the control winding N_C, then this impedance will vary inversely with the

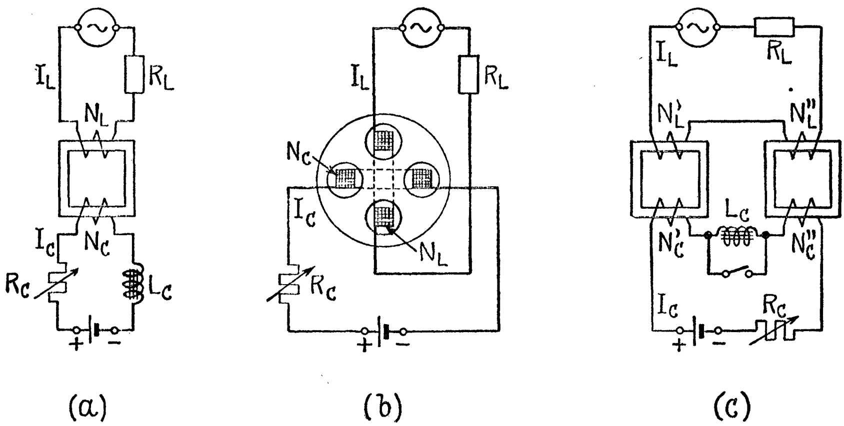

FIG. 3.1. Basic saturable-reactor circuits (*a*) with a single core element (*Burgess and Frankenfield,* 1901); (*b*) with a special core construction (*Burgess and Frankenfield,* 1901); (*c*) with two separate core elements (*Alexanderson,* 1912).

amount of d-c magnetization, any increase in d-c magnetization decreasing the impedance, and vice versa. If the core of the saturable reactor is saturated, then the impedance of the a-c circuit is reduced to the minimum, and the load current I_L will have its highest value.

Figure 3.2 gives a reproduction of the original diagrams,* as presented by Burgess and Frankenfield, for illustrating the fundamental mode of operation of the basic saturable-reactor circuit:

"When a piece of magnetic material is magnetized by an alternating current, the well-known hysteresis loop (shown in Fig. 3.2*a*) is described during each cycle. Let OC represent the maximum instantaneous value of this current, and BB' the total change in magnetization caused by variation of the current from the maximum positive value OC to an equal and maximum negative value OC'. Now suppose that the same magnetic material is initially magnetized by a direct current to a point O' (Fig. 3.2*b*) on the magnetization curve, which is represented by the dotted curve $OO'B$. The point O' then represents an amount of magnetization

* In the following quotation of the original text of United States patent 743,444 explaining "Fig. 1" (Fig. 3.2*a*) and "Fig. 2" (Fig. 3.2*b*), only the reference numbers of these figures have been changed.

CO' produced by a current OC. If an alternating current of the same value as before be used to magnetize the material in the same linear direction as the direct current, Cc and Cc' will represent its equal maximum instantaneous values, and the total change in magnetism caused by the variation from the maximum positive value Cc to an equal maximum negative value Cc' of the alternating current will be represented by the distance BB'. The same number of magnetizing turns are assumed for the alternating current in Fig. 3.2*a* and *b* and for the direct current in Fig. 3.2*b*, and the magnetism is taken in the same direction and in the

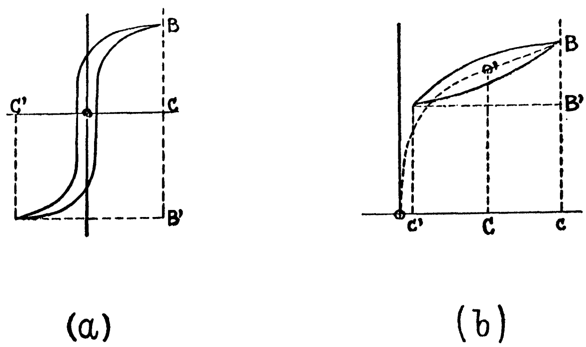

FIG. 3.2. Reproduction of the original diagrams in United States patent 743,444 (*Burgess and Frankenfield,* 1903) illustrating the fundamental mode of operation of saturable-reactor circuits: (*a*) dynamic hysteresis loop without d-c magnetization; (*b*) dynamic minor hysteresis loop with d-c magnetization.

same plane in both figures. It will be noticed that the range of variation BB' is greater in Fig. 3.2*a* than in Fig. 3.2*b*. In other words, the self-inductance for a complete cycle will be greater for the conditions represented by Fig. 3.2*a* than in Fig. 3.2*b*. When the self-inductance is varied by this means, the two magnetizations being in the same plane and in the same direction, the alternating current induces in the d-c circuit an alternating electric pressure after the manner of the action of the primary of a transformer on its secondary. In many practical applications it is desirable to avoid this transformer action, and to do so, the two magnetizing coils must be so placed that their mutual inductance is zero. It is, furthermore, necessary that the variation of magnetism represented by the distances BB' in Fig. 3.2*a* and *b* be regulable by change of the direct magnetization with this arrangement of the coils."

Burgess and Frankenfield have described different types of saturable-reactor devices for reducing or practically eliminating the mutual inductance between the a-c load circuit and the d-c control circuit as follows: Three-legged and four-legged cores, hollow annular cores, and magnetic "cross-valve" arrangements. Figure 3.1*b* shows as an interesting example* a special core construction in which the plane of the load winding N_L is substantially at an angle of 90° to the plane of the control winding N_C so that the mutual inductance between these windings is practically zero.

Another method (E. F. W. Alexanderson, 1912[2]) for eliminating undesired effects of transformer action between load winding and control winding consists in using two separate and equally rated saturable-core-reactor elements (twin-core arrangement) with series-aiding-connected d-c control windings N'_C, N''_C and series-opposing-connected a-c load windings N'_L, N''_L, as shown in Fig. 3.1*c*. In this case, the fundamental-frequency voltage induced in the first control winding N'_C will be canceled by the equal and opposite voltage induced in the second control winding N''_C. Because no net voltage of fundamental frequency will appear at the terminals of the d-c control circuit, it is not necessary to provide a choke coil (L_C) in the control circuit. However, incorporation of such a choke coil may be valuable in special cases to prevent the circulation of even-harmonic-frequency currents in the control-circuit loop.

The original purpose of the basic arrangements, as disclosed by Burgess and Frankenfield,[3] is "to provide a method of regulating and controlling the current and potential in an electric circuit or system of circuits whereby the use of heavy movable parts and expensive apparatus is avoided and simple and effective means are provided in place thereof and smooth and uniform variations in the circuit are obtained for the purposes of regulation without the use of a multiplicity of contacts carrying load currents." This disclosure, however, makes no claims regarding amplifying properties.

It is to be noted that moderate power amplification can be obtained with such simple arrangements when the load resistance R_L is matched to the impedance values of the saturable reactors in the output circuit so that the "output power" $P_L = I_L{}^2R_L$ is greater than the "input power" $P_C = I_C{}^2R_C$ representing the actual losses in the d-c control circuit. The actual value of the power gain $K_P = P_L/P_C$ (about 10 to 100) depends on the design and size of the saturable-core reactor.

The simple saturable-reactor circuit is useful for light control in theaters and large public rooms where it is pleasant to the eye that the light is put on and off slowly. Of course, it can also be used to adjust the lighting to different intensities without moving parts or contact devices requiring supervision and maintenance. The inherent amplifying properties of the circuit make it possible to control large power outputs by means of small devices having low power consumption.

In 1913, E. Besag[4] introduced the combination of a twin-type saturable-reactor arrangement (Fig. 3.1*c*) with direct-indicating instruments and

* Corresponding to Fig. 8 of United States patent 720,884.

recorders for measuring large direct currents (I_C) by means of the magnetomotive force around the d-c conductor (see also Fig. 3.3*a*), without using extremely large-sized shunt resistors, which are very expensive and difficult to cool. The a-c windings (N'_L, N''_L) of the two saturable-reactor elements are connected in series and to an a-c power supply; the alternating current (I_L) flowing through these windings is measured with an iron-vane-type ammeter (R_L), which is calibrated in d-c units, to indicate directly the direct current in the bar.

3.3. Performance of the Basic Saturable-reactor Circuit with Rectangular-hysteresis-loop Core Material. In 1937, W. Krämer[5] revealed that the basic saturable-reactor circuit with twin-core elements (Fig. 3.1*c*) can be considerably improved in the following way (Fig. 3.3*a*):

1. Using high-permeability nickel-iron-alloy twin-core elements, preferably with rectangular-hysteresis-loop core material[6]
2. Measuring the average value of the alternating current $I_{2\sim}$ flowing through the load windings N'_L, N''_L by means of the combination of a full-wave-rectifier circuit with a moving-coil-type d-c ammeter (or an integrating d-c meter) A
3. Providing an inductance L_A for smoothing the direct current I_{2-}, which is a function of the direct current I_{1-} in the bar

Krämer has shown that such an arrangement has inherent "current-transformer characteristics": The ratio between the "primary" direct current I_{1-} to be measured and the average value of the rectified "secondary" current I_{2-} is equal to the ratio between the secondary ($N'_L = N''_L$) and primary ($N'_C = N''_C$) number of turns on the cores. With $N'_C = N''_C = 1$, this turn ratio is $N'_L = N''_L$. The current-transformer character of this arrangement is to a certain extent independent of variations in magnitude, frequency, and waveshape of the a-c power-supply voltage E_P.

In order to understand the mode of operation of this arrangement, it is important to note the following facts (Fig. 3.3*a*):

1. The a-c windings N'_L, N''_L are connected in series in such a way that, when the current $I_{2\sim}$ assists the magnetomotive force of the direct current I_{1-} in one core, it opposes that magnetomotive force in the other core.
2. Actually, the performance of rectangular-hysteresis-loop core materials is such that the permeability is very high and nearly constant up to saturation, at which point further increases in magnetomotive force will produce only very small increases in magnetic flux (Fig. 2.1).
3. The resistances of the a-c windings and the rectifier-ammeter circuit are assumed to be negligible.

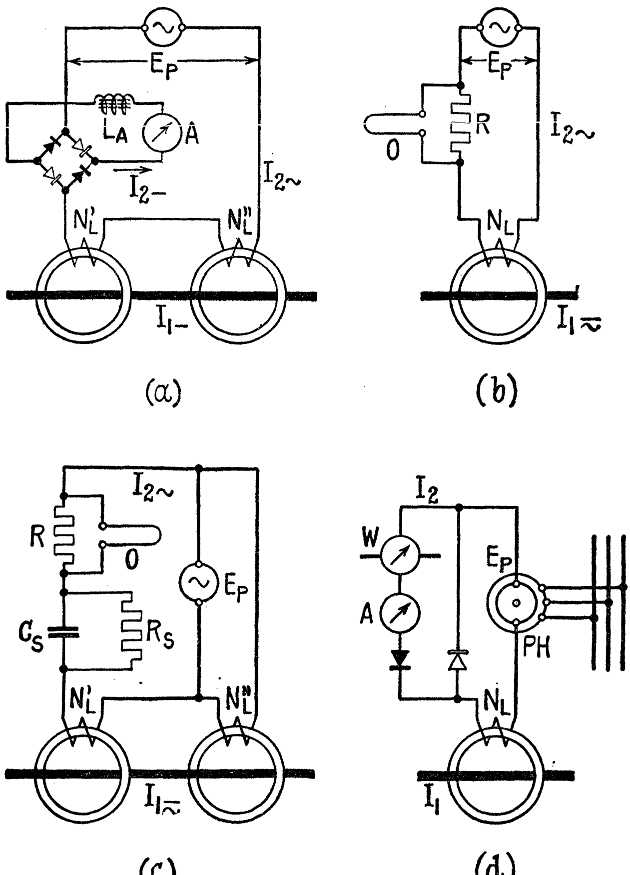

FIG. 3.3. Saturable-reactor circuits with rectangular-hysteresis-loop core material for d-c measurements. (*a*) D-c transductor applied to measurement of very large d-c bus-bar currents, using series-connected a-c windings. (*Krämer*, 1937.) (*b*) Basic circuit of a universal type of current transformer for oscillographic measurements on large pulsating currents having superimposed d-c and a-c components. (*Krämer*, 1939.) (*c*) Improved circuit for oscillographic measurements containing a damping resistor R_S with shunt condenser C_S. (*Krämer*, 1939.) (*d*) Simple one-core circuit for special measurements on large mercury-arc-rectifier installations. (*Krämer*, 1939.)

4. According to the practical working conditions, it can also be assumed that the saturable-reactor device has a negligible reaction on the bus-bar current I_{1-}, which comes from a nearly infinite reservoir.
5. Because current I_{1-} exerts a positive magnetomotive force in one core and a negative magnetomotive force in the other core, the algebraic sum of the fluxes in each core must be such as to give the flux pulsation demanded by the applied a-c voltage E_P.

Figure 3.4*a* shows that, as a result of these special working conditions of the circuit, the alternating current $I_{2\sim}$ has a nearly rectangular waveshape. The actual ampere-turns of this current are equal to the ampere-turns produced by the "primary" direct current I_{1-}, independent of the magnitude of the applied a-c voltage E_P; and the "secondary" current

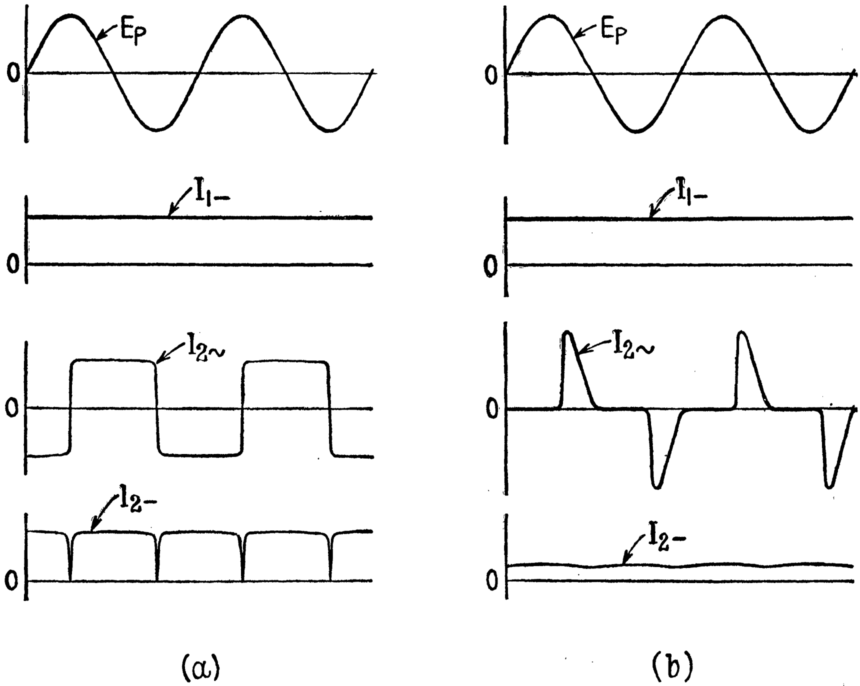

FIG. 3.4. Waveshapes of a-c supply-voltage E_P, "primary" direct current I_{1-} to be measured, "secondary" alternating current $I_{2\sim}$, and rectified current I_{2-} in the ammeter *A*. (*a*) Measurement of large d-c bus-bar currents (Fig. 3.3*a*). The alternating current $I_{2\sim}$ has a nearly rectangular waveshape. (*b*) Measurement of high d-c voltages (Fig. 3.5). The alternating current $I_{2\sim}$ has a typical sharp-pulse waveshape.

I_{2-} in the d-c ammeter *A* is the rectangular current with every other half-cycle wave reversed. Thus, I_{2-} is a smooth direct current, which is proportional to I_{1-} and is substantially independent of the a-c voltage E_P. Evidently, such an arrangement acts to a certain extent in a similar way as an a-c current transformer.

Krämer has also shown[7] that the current-transformer action of saturable-reactor circuits, as shown in Fig. 3.3*a* to *d*, extends further than to the average values, if a rectangular-hysteresis-loop core material is used. Indeed, the alternating current $I_{2\sim}$ is not only proportional to the direct current I_{1-} but to some extent will even have a similar oscillographic variation when I_{1-} varies, owing to a superimposed a-c component.

Figure 3.3*b* illustrates the basic circuit of a universal type of current transformer which makes it possible to perform oscillographic measurements on large pulsating currents having superimposed d-c and a-c components. The a-c winding N_L of this one-core arrangement is connected to the auxiliary a-c voltage E_P through a mirror-type oscillograph O with shunt resistor R. Because high-permeability core material is used, the actual exciting current flowing through N_L, R, and O is extremely small and has substantially no influence on the oscillographic picture. When the primary current I_1 to be investigated is a pure alternating current (having no d-c component at all), this current will be transferred on the secondary circuit N_L, R, O in a similar way as with a conventional a-c current transformer. This is true because the internal resistance of the a-c source does not represent an appreciable load with respect to this transformer action. The only difference consists in the fact that a large a-c flux, corresponding to the a-c voltage E_P, will be superimposed upon the a-c flux, which is produced by the primary a-c current I_1. This, however, does not affect the performance of the current transformer in an unfavorable way, because the superposition of the two a-c fluxes happens in the range of high permeability. The effect of the auxiliary a-c voltage E_P starts when the primary current I_1 to be investigated contains a d-c component. It has been proved[8] that this arrangement gives a true oscillographic picture even when the superimposed d-c component of I_1 is very large.

Figure 3.3*c* shows an improved circuit with twin cores containing a "damping resistor" R_S with shunt condenser C_S to eliminate undesired transient effects. Figure 3.3*d* shows a simple one-core circuit containing (1) the a-c winding N_L, (2) an ammeter A and a wattmeter W with series-connected half-wave rectifier, (3) a second half-wave rectifier in shunt connection, and (4) a phase shifter PH. This arrangement makes it possible to change the phase relationship between the auxiliary a-c voltage E_P and the synchronous a-c component of the current I_1 so that the most interesting parts of I_1 will appear inside the oscillographic picture. Furthermore, it is possible to measure the half-wave current I_2, corresponding in every respect to the current I_1 to be investigated, by means of ammeter A and wattmeter W. This circuit is particularly suitable for special measurements on large mercury-arc-rectifier installations. The arc losses, the anode current, and the average value of the arc voltage can be correctly measured in this way. By selecting a suitable turn ratio between the primary and secondary windings of the saturable-reactor elements, it is possible to represent the pulsating current I_1 on any desired scale. It is to be noted that these windings can be made highly insulated from each other, as in the case of an ordinary a-c transformer.

Arrangements like those shown in Fig. 3.3 are primarily intended for measuring large direct currents in the range from 100 to 10,000 amp. However, they may equally well be used for measuring high direct voltages. Figure 3.5 shows a suitable circuit for this purpose (W. Krämer[9]). The only difference will be that the direct voltage E_X to be measured

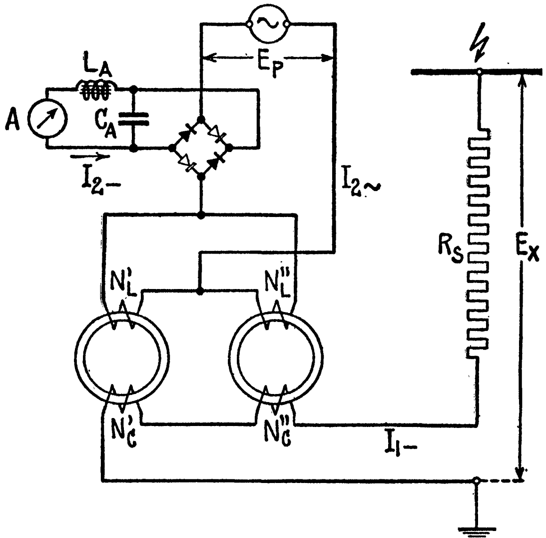

FIG. 3.5. Saturable-reactor circuit with rectangular-hysteresis-loop core material for measurement of high d-c voltages, using parallel-connected a-c windings. (*Krämer*, 1938.)

must first be translated into a direct current proportional to the voltage E_X by means of a series resistor R_S. This current I_{1-} will be of a magnitude which necessitates employing d-c control windings N'_C, N''_C with a large number of turns.

In this case, however, it is important to consider the fact that, with *series*-connected a-c windings N'_L, N''_L, second-harmonic-frequency voltages will be induced in the control windings N'_C, N''_C, as a result of the superposition of the a-c and d-c fluxes in the magnetic cores. The performance of the circuit with *series*-connected a-c windings, as shown in Fig. 3.3*a*, is substantially not influenced by this fact because only one "primary" turn ($N'_C = N''_C = 1$) is used and because the induced voltages of second-harmonic frequency are of the order of 0.07 to 0.1 volt.

If control windings N'_C, N''_C with a large number of turns are provided, as in the circuit of Fig. 3.5, then second-harmonic-frequency voltages of several hundred volts may be induced in the control windings. For this reason, *parallel*-connected a-c load windings, representing a short-circuit path for the induced second-harmonic-frequency currents, are used in this arrangement. These currents will circulate in the closed-loop circuit of the windings N'_L, N''_L, and not more than a very small fraction will appear in the control circuit.

Krämer has shown that the basic saturable-reactor circuit* with *parallel*-connected a-c windings also has current-transformer properties,

1. If high-permeability core material, preferably having a rectangular hysteresis loop, is used
2. If the actual average value of the total alternating current $I_{2\sim}$ is measured by means of a full-wave rectifier with a d-c instrument (ammeter or wattmeter)
3. If a properly rated filter circuit L_A, C_A is provided to smooth the "secondary" direct current I_{2-} flowing through the d-c instrument

This arrangement has proved to be suitable for the measurement of high d-c voltages up to about 10,000 volts. The direct control current I_{1-} is about 3 to 30 ma. Variations in the a-c supply voltage E_P within the limits of ± 10 per cent have no appreciable effect on the performance characteristics. It is important to note that the meter circuit (having ground potential) is completely insulated from the d-c high-voltage circuit, even if the ground connection may by accident be interrupted.

Figure 3.4*b* shows a typical waveshape of the total alternating current $I_{2\sim}$ and the smoothed direct current I_{2-}, which is proportional to I_{1-} and is substantially independent of the a-c supply voltage E_P. Comparison between the waveshapes obtained with *series*-connected a-c windings (Fig. 3.4*a*) and those obtained with *parallel*-connected a-c windings (Fig. 3.4*b*) makes it evident that there is a fundamental difference in the operating conditions of these two basic circuits, both of which have special fields of application in saturable-reactor technique.

3.4. Auto-connected Saturable-reactor Circuits. Figure 3.6 shows two examples for bridge-type saturable-reactor circuits with a-c or d-c load R_L. In these "auto-connected"[10] nonfeedback circuits both alternating and direct currents flow in the same winding of the twin-core elements. The direct control current I_C is led to terminals in the a-c windings, N_1, N_2, N_3, N_4, which are equipotential with respect to the a-c supply voltage E_P. The actual a-c and d-c magnetizations are properly superimposed in both cores, so that each element behaves just as it would with separate a-c and d-c windings.

Such arrangements make it possible to save a considerable amount of material and to avoid leakage reactance between the a-c and the d-c cir-

* Preferably using a circuit arrangement as shown in Fig. 3.5.

cuits. A. U. Lamm[11] and B. Nordfeldt[12] have shown that the basic principle of the auto-connected single-phase saturable-reactor circuit (Fig. 3.6) can be extended to three- and six-phase networks. According to Lamm's theoretical investigations, these arrangements "are fundamentally simpler in their mode of action, and their current and voltage relations are easier to compute. In those cases where several parallel paths are

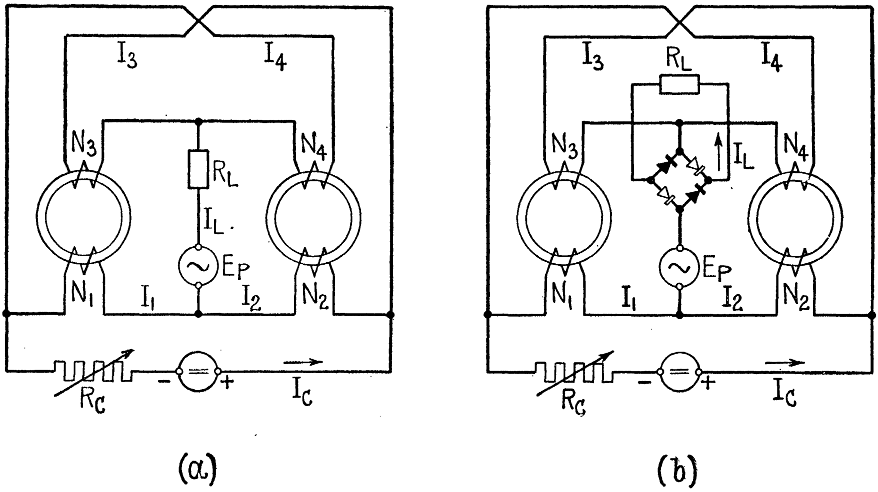

FIG. 3.6. Auto-connected nonfeedback saturable-reactor circuits (*a*) with an a-c load; (*b*) with a d-c load.

available for the direct current, it may be necessary to take special measures in order to compensate incidental resistance asymmetry due to irregularities of manufacture."

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6. See, for example, E. Both, "Some Factors Governing the Achievement of a Rectangular Hysteresis Loop," Magnetic Materials Symposium, U.S. Naval Ordnance Laboratory, Washington, D.C., pp. 39–46, June 15, 1948.
7. W. Krämer, "Fremderregter Stromwandler als Universalwandler zum Oszillographieren von Wechselströmen mit Gleichstromgliedern" (Separately excited current transformer operating as a universal-type transformer for oscillographic investigations on alternating currents with d-c components), *Elektrotechnische Zeitschrift* (*ETZ*), Vol. 60, pp. 393–395, 1939.
8. See, for example, Figs. 8 and 9 of Krämer's paper in *Elektrotechnische Zeitschrift* (*ETZ*), Vol. 60, p. 395, 1939. These figures show oscillograms of unidirectional currents containing very large d-c components.
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CHAPTER 4

Operating Conditions of Nonfeedback Magnetic-amplifier Circuits

4.1. Introduction. The basic nonfeedback types of magnetic-amplifier circuits dealt with in the preceding chapter and used for controlling theater lighting loads and for special d-c measurements are characterized by a moderate power amplification (about 10 to 100) and inherent current-transformer properties, if a high-permeability nickel-iron-alloy core material is used.

It has already been stressed that there is a fundamental difference in the operating conditions of circuits with *series*-connected a-c windings (Fig. 3.4*a*) and those having *parallel*-connected a-c windings (Fig. 3.4*b*). This difference is caused essentially by the effect of double-frequency currents, which either are prevented from flowing through the control circuit (Fig. 3.1*c*: "series circuit" with choke coil L_C) or are circulating in the closed-loop circuit of the parallel-connected a-c windings N'_L, N''_L (Fig. 3.5).

This chapter describes briefly the typical operating conditions of nonfeedback circuits with high-permeability cores and gives actual performance characteristics of such circuits.

4.2. Effects of Double-frequency Currents on the Performance of Magnetic-amplifier Circuits. In 1940 to 1943, the author[1] described experimental investigations on high-performance magnetic amplifiers and also some measurements concerning the effects of double-frequency currents on the performance of various circuits with series-connected a-c windings. The results of these measurements can be summarized as follows:

1. Transfer characteristics representing the output (load) current as a function of the input (control) current are considerably influenced by the actual value of total resistance of the d-c control-circuit loop.
2. These transfer characteristics can be considerably improved in many cases by providing a properly rated condenser across the series-connected d-c control windings (or across some equivalent additional windings).

3. Voltage-current characteristics representing the "quiescent current" (the load current with no-input-current conditions) as a function of the a-c power-supply voltage can be improved in such a way that constant-current characteristics are obtained, if such a condenser is used.
4. If the control current is produced by means of a barrier-layer photocell, a condenser across the control windings (or some equivalent windings) must be used to prevent the double-frequency currents from being rectified and converted into an additional direct current, which produces, if present, an undesired additional d-c magnetization of the cores.
5. The effect of such a condenser on the speed of response of the magnetic amplifier has to be watched, because the condenser providing a short-circuit path for the induced double-frequency currents has also a certain effect on the actual time constant of the complete arrangement.

4.3. "Natural" and "Forced" Magnetization Conditions. T. Buchhold[2] in 1942 gave a qualitative explanation of the flux and current waveforms in saturable-reactor circuits with applied sinusoidal voltage and no-load conditions. He showed that the harmonics of the ampere-turns of the twin-core elements add together in such a way that only the fundamental and *odd* harmonics appear in the alternating load current, while *even* harmonics either circulate between interconnected windings or enter into the d-c control circuit.

Buchhold distinguishes between "natural" ("natürliche") and "forced" ("erzwungene") magnetization, depending on whether even-harmonic currents can flow freely or are suppressed. He describes these two typical magnetization conditions in the following way:

1. In the saturable-reactor circuit with *parallel*-connected a-c windings, and in the circuit with *series*-connected a-c windings and a d-c control circuit representing a very *low* impedance with regard to even-harmonic currents (or with a properly rated condenser across the series-connected d-c control windings), the even-harmonic currents can flow freely. This represents "natural" magnetization conditions, as characterized by a typical sharp-pulse waveshape of the load current (Fig. 3.4*b*).
2. In the saturable-reactor circuit with *series*-connected a-c windings and a d-c control circuit representing a very *high* impedance (*e.g.*, choke coil or high series resistance) with regard to even-harmonic currents, these currents are suppressed. This represents "forced" magnetization conditions, as characterized by a rectangular waveshape of the load current (Fig. 3.4*a*).

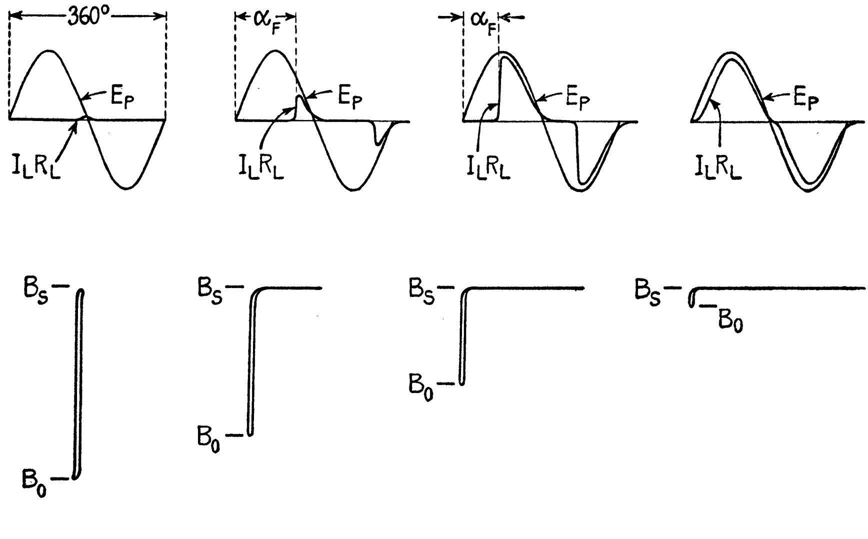
The essential difference between these two magnetization conditions in respect to the waveshape of the alternating load current consists in the fact that the third harmonic and other odd harmonics, which are present in both, occupy in the case of forced magnetization such a phase position that the waveshape is flat-topped (nearly rectangular: Fig. 3.4*a*), while the phase position in the case of natural magnetization is such that the peak of the current wave is made more acute (sharp-pulse waveshape: Fig. 3.4*b*). The fact that the third-harmonic components and other odd-harmonic components are in antiphase in these two boundary cases makes it possible to get rid of the third harmonic and other odd harmonics of the alternating load current. Indeed, by applying series-connected a-c windings and moderate suppression of the even-harmonic currents, an almost purely sinusoidal waveshape of the load current can be obtained.[3] Of course, this will be possible only if the load resistance is so small that the a-c voltage across the series-connected a-c windings is sinusoidal.

Referring to the practical applications of the two magnetization conditions, it is to be noted that forced magnetization is generally used in connection with nonfeedback circuits for measuring large direct currents and oscillographic investigations, as illustrated in Fig. 3.3. Natural magnetization is mostly used in connection with nonfeedback circuits for measuring high direct voltages (Fig. 3.5) and those types of magnetic-amplifier circuits which utilize feedback.

4.4. Definition of Elementary Operation of the Saturable Reactor with High-permeability Core Material and Natural Magnetization Conditions. The basic nonfeedback circuit with parallel-connected a-c windings N'_L, N''_L (Fig. 3.5) will serve as a demonstration of the elementary operation of the saturable reactor with high-permeability core material and natural magnetization conditions. It is assumed that a variable, pure direct current I_C is applied to the control windings N'_C, N''_C of this circuit to adjust the magnitude of the alternating load current I_L flowing through an ohmic load resistor R_L.

The upper part of Fig. 4.1, showing the sinusoidal power-supply voltage E_P and the actual load voltage I_LR_L as functions of time, indicates the operating conditions with varying values of control current I_C. This example indicates the manner in which the a-c load current I_L and the voltage drop I_LR_L can be smoothly adjusted by means of changes of direct control current I_C.

Corresponding magnetic conditions of one core element are shown in the lower part of Fig. 4.1, which gives the so-called dynamic minor hysteresis loops (see also Fig. 3.2*b*) associated with the four cases Fig. 4.1*a* to *d*. The effect of control current I_C on the minimum or initial flux density B_0 is indicated. It is important to note that the maximum flux density at the minimum current or cutoff condition is so adjusted that the hysteresis loop is symmetrical and shows a total excursion of flux density of just less than twice the saturation flux density B_S of the magnetic material. F. G. Logan[4] has stressed that this is one of the more important design features of such circuits, especially where it is desired



(a) (b) (c) (d)

FIG. 4.1. Diagrams illustrating the elementary operation of the saturable reactor with high-permeability core material and natural magnetization conditions (*a*) with minimum value of control current I_C; (*b*) with low value of I_C; (*c*) with medium value of I_C; (*d*) with high value of I_C. The upper part of the illustration shows the sinusoidal power-supply voltage E_P and the actual load voltage I_LR_L as functions of time; the lower part shows the corresponding dynamic minor hysteresis loops of one core element. (*F. G. Logan.*[4])

to secure the optimum range of output for a given reactor structure and control signal.

Referring to the upper part of Fig. 4.1, it is evident that the actual output-voltage waves (I_LR_L), resembling the output of a gas-filled, grid-controlled electronic tube, contain a portion with very great slope, which is associated with "firing" of the saturable-reactor element. The time at which firing occurs depends on the actual magnetomotive force of the control windings N'_C, N''_C carrying the pure direct current I_C. Thus, actual magnitude of the "firing angle" α_F depends on the initial magnetic-flux density B_0 associated with this magnetomotive force: For *increased* values of B_0, firing occurs at *earlier* instants, *i.e.*, at *smaller* firing angles

α_F.* Before firing (during the cutoff period), there is substantially no voltage drop $I_L R_L$ across the load resistance R_L, the whole of the applied voltage E_P being sustained by the a-c windings N'_L, N''_L. After firing, a large part of E_P appears across the load R_L, the maximum value of load voltage $I_L R_L$ being dependent on the actual values of the ohmic resistances in the a-c load circuit.

Saturable-reactor circuits, as described thus far in Chap. 3, are susceptible to control by various means. The fundamental principle of operation consists in the adjustment of the value of initial flux density B_0 during the cutoff period. For example, the following means may be employed to adjust this value of B_0:

1. Pure direct current, as derived from batteries, thermocouples, photoelectric cells, etc.
2. Unidirectional current, derived from full-wave-rectifier circuits
3. Unidirectional current, derived from half-wave-rectifier circuits (synchronous half-wave pulses)
4. Synchronous alternating current, adjustable either in amplitude or in phase (*e.g.*, by means of a phase-shifting network)
5. Synchronous a-c component of a unidirectional current, adjustable either in amplitude or in phase, derived from half-wave-rectifier circuits

Application of "modulation," *i.e.*, control by means of amplitude variation of asynchronous alternating current, is also feasible in many cases. In addition, control may also be achieved by means of adjustable impedance elements (*e.g.*, variable resistors, condensers, or reactors) across the already available windings or across additional windings of the saturable-reactor elements.

The elementary operation of the saturable reactor with high-permeability core material and natural magnetization conditions has been recognized to be that of a switch (A. U. Lamm[5]), making at a controllable instant of the supply-voltage wave and breaking at or near the zero passage of the current wave. Such a performance is achieved by utilizing the saturable-reactor units as circuit elements of alternately very high and very low impedance, an effect facilitated by the existence of a point of rapid change in slope of the magnetizing characteristic.[6]

It is important to stress the fact that application of a suitable form of core construction (preferably spirally wound tape cores or "sideways-air-gap" cores) in combination with a rectangular-hysteresis-loop core

* In European literature, the angle α_F where the "transductor element" becomes "completely discharged" is sometimes termed the "control angle," in analogy with the ignition angle for grid-controlled mercury-arc rectifiers.

material (*e.g.*, Orthonol, Deltamax, etc.) results in a very close approximation of actual experimental conditions (Fig. 2.1*a*) to the usual theoretical assumptions represented by an idealized magnetization curve which consists of three straight lines,[7] as shown in Fig. 2.1*b*. This analytical representation involves the assumptions that the core losses are negligible and that below saturation the permeability is infinite while in saturation the differential permeability is zero. As a result of the very close approximation between these theoretical assumptions and actual experimental conditions, performance characteristics of such magnetic amplifiers can be represented by simple linear equations with sufficient accuracy to be used in the design of these amplifiers.[8]

Indeed, the very close approximation between the performance characteristics of an "ideal" saturable-reactor circuit and those of an actual circuit makes it evident that the conception of regarding a saturable-reactor element not as a variable reactor but as a "magnetic switch" is justified. Here, it is interesting to note that, according to recently proposed British definitions,[9]

1. "A transductor is basically a twin-core saturable reactor in which, at the minimum current or cutoff condition (when the alternating voltage supplied to the transductor is a maximum), the total a-c flux swing approaches the available maximum as limited by core saturation. Thus a transductor does not work by virtue of changes of incremental permeability of the core with change of d-c excitation, but usually as a consequence of the distortion of the flux waveform from the sinusoidal form as the d-c excitation tends to push the flux wave into saturation."
2. "A magnetic amplifier is an electrical power amplifier embodying one or more transductors together with the rectifiers, resistors, capacitors, transformers, etc., which are necessary to utilize the power-amplifying properties of the transductor or transductors."

4.5. Actual Performance Characteristics of Nonfeedback Magnetic-amplifier Circuits. In 1940, the author[10] pointed out that the following characteristic curves with nonfeedback conditions represent very important fundamentals of magnetic-amplifier design:

1. Load current I_L as a function of power-supply voltage E_P with d-c control current I_C as a parameter
2. Actual impedance E_P/I_L as a function of E_P with I_C as a parameter, particularly with $I_C = 0$
3. Load current I_L as a function of d-c control current I_C with load resistance R_L as a parameter

Figures 4.3 to 4.8 give results of various measurements representing typical examples of actual performance characteristics of nonfeedback magnetic-amplifier circuits with Mumetal or Orthonol tape cores.* These examples prove that there is a very close approximation between idealized and actual performance, if such high-permeability core materials are used.

The test circuit (Fig. 4.2) contains two equally rated Mumetal or Orthonol tape-core elements having series-opposing-connected a-c load windings N'_L, N''_L, and series-aiding-connected d-c control windings N'_C, N''_C, so that no induced voltage of the fundamental wave and the odd harmonics of the a-c supply voltage E_P (frequency $f_P = 60$ cps) can appear at the input terminals. The a-c load resistor R_L and a full-wave-type selenium dry-disk rectifier are series-connected with N'_L and N''_L. Thus, the direct load current I_L flowing through the moving-coil precision-type milliammeter A_L represents the average value of the rectified current which flows through N'_L and N''_L. Actual average value of the a-c supply voltage E_P, indicated by a rectifier-type voltmeter V_P, can be varied by means of a regulating transformer T_P.

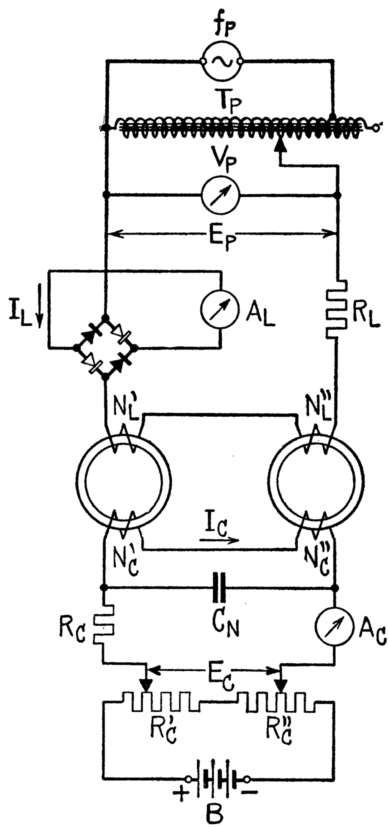


FIG. 4.2. Test circuit used for the determination of actual performance characteristics of nonfeedback magnetic-amplifier circuits.

The control current I_C flowing through a moving-coil precision-type milliammeter A_C, a fixed resistor R_C, and the control windings N'_C, N''_C is supplied from a storage battery B and can be controlled by means of two potentiometer resistors R'_C (rough adjustment) and R''_C (fine adjustment). A condenser C_N (about 4 μf) is connected in parallel to the control windings, in order to provide a very small impedance for even-harmonic currents being induced in these windings, so that variations in total resistance of the control circuit have no influence upon the operating conditions of the two saturable-reactor elements. Thus, the test circuit is always working with natural magnetization conditions.

* The author made these measurements in the U.S. Naval Ordnance Laboratory, Silver Spring, Md.

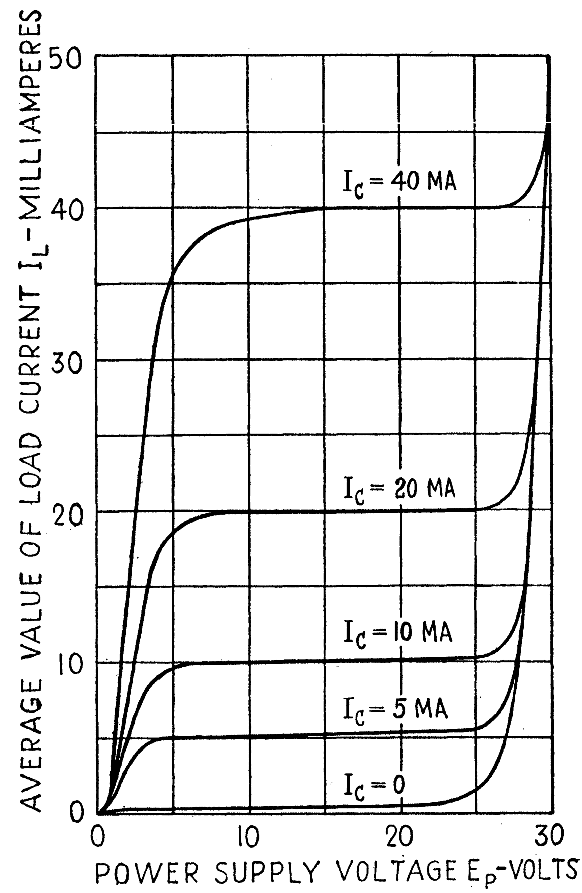


FIG. 4.3. Load current I_L as a function of power-supply voltage E_P with control current I_C as a parameter (results of measurements with Mumetal tape cores).

As mentioned previously, the measurements on the test circuit of Fig. 4.2 with Mumetal and Orthonol 2-mil tape cores are divided into determination of the three characteristic curves:

1. $I_L = f(E_P)$ with I_C as a parameter
2. $E_P/I_L = f(E_P)$ with $I_C = 0$
3. $I_L = f(I_C)$ with R_L as a parameter

In both cases, the two saturable-reactor elements consisted of spirally wound 2-mil tape cores having the following dimensions:

Outer diameter, 2.0 in.
Inner diameter, 1⅝ in.
Tape width, 0.5 in.

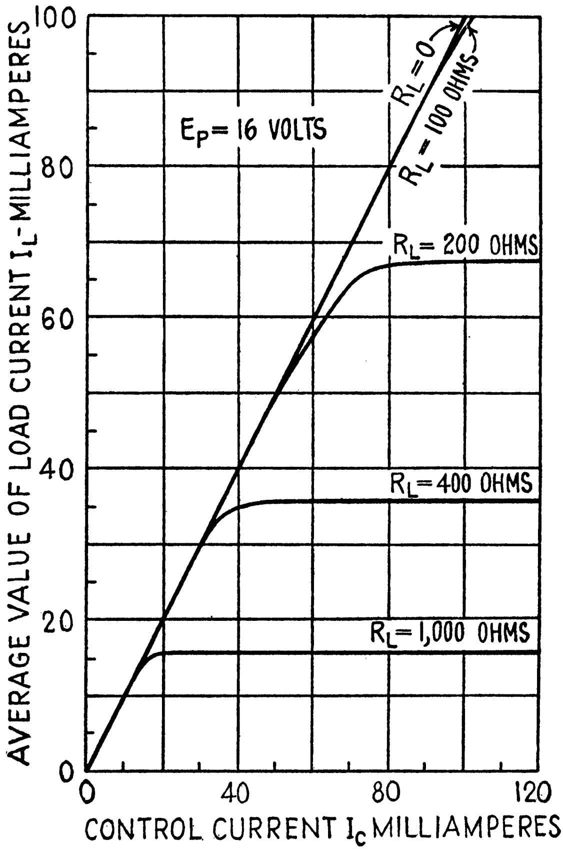


FIG. 4.4. Load current I_L as a function of control current I_C with load resistance R_L as a parameter (results of measurements with Mumetal tape cores).

The tape cores had the following multilayer toroidal windings of enamel-insulated copper wire:

1. Inner a-c load windings N'_L, N''_L of 1,000 turns each of No. 28 wire (Brown and Sharpe gauge)
2. Outer d-c control windings N'_C, N''_C of 1,000 turns each of No. 28 wire (Brown and Sharpe gauge)

The magnetic circuit of each core has an effective area of

$$0.084 \text{ in.}^2 = 0.54 \text{ cm}^2$$

and a mean length of 5.7 in. = 14.5 cm.

Thus, the measurements with Mumetal and Orthonol elements have been made under almost identical conditions with regard to the dimensions of the cores and of the windings; only the core material has been changed to get the net differences between the actual performance characteristics, with nonfeedback conditions and natural magnetization, as defined in Sec. 4.3.

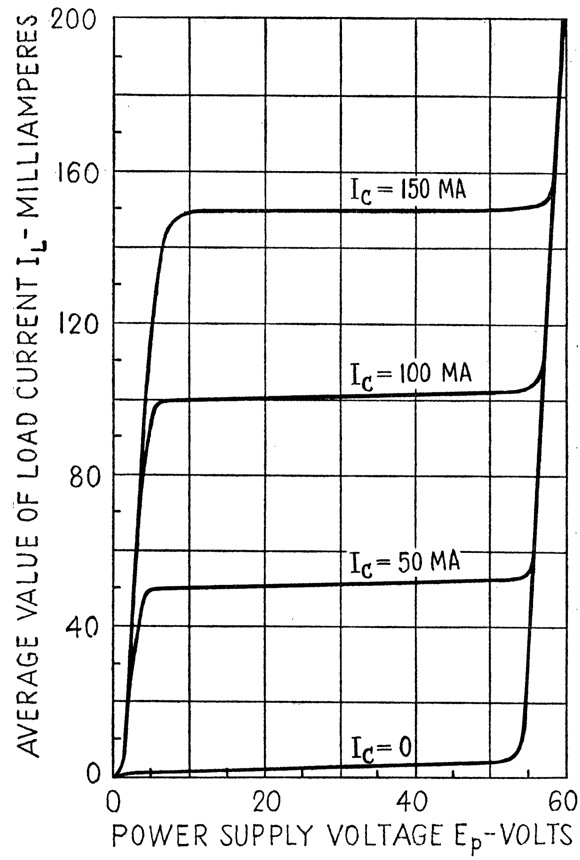


FIG. 4.5. Load current I_L as a function of power-supply voltage E_P with control current I_C as a parameter (results of measurements with Orthonol tape cores).

Figures 4.3 and 4.4 (Mumetal cores) and 4.5 and 4.6 (Orthonol cores) show the average value of load current I_L as a function of average value of sinusoidal power-supply voltage E_P, with direct control current I_C as a parameter, and average value of I_L as a function of I_C, with load resistance R_L as a parameter. The actual or measurable value of load current I_L is the resultant of the "exciting-current component" I_{L0}, which is practically independent of the control current I_C and of the "load component" I_{LC} flowing in response to changes in control current I_C. Equal increments in control ampere-turns I_CN_C produce approximately equal increments in load-component ampere-turns $I_{LC}N_L$ in the linear region of the characteristics. Thus, since I_{L0} can normally

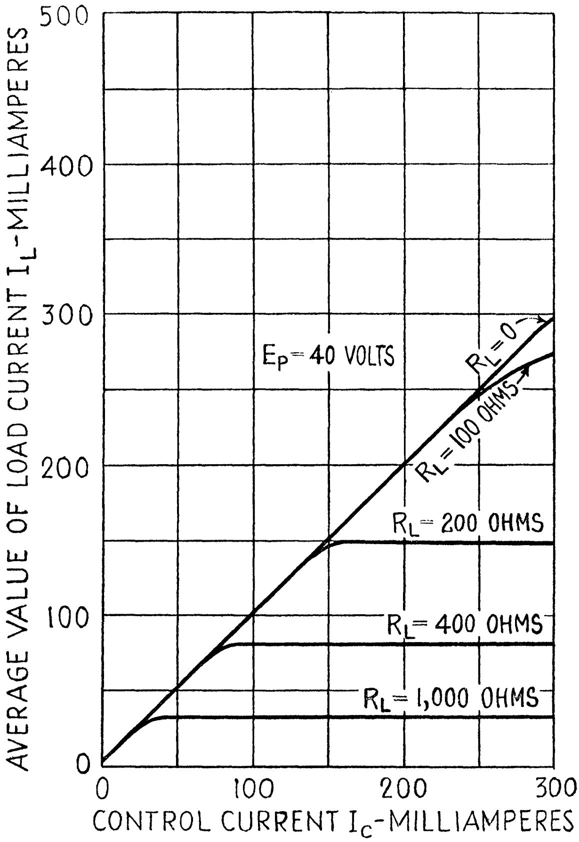


FIG. 4.6. Load current I_L as a function of control current I_C with load resistance R_L as a parameter (results of measurements with Orthonol tape cores).

be made very small compared with I_{LC}, the following equation holds to a good degree of accuracy:

$$I_{LC}N_L = I_LN_L = I_CN_C \tag{4.1a}$$

or

$$\frac{I_L}{I_C} = \frac{N_C}{N_L} \tag{4.1b}$$

There is a linear relationship between I_L and I_C, because of the inherent "current-transformer properties," as already described in Chap. 3. Over certain ranges, changes in applied a-c voltage E_P and changes in load resistance R_L have no appreciable effect upon the current ratio

I_L/I_C ("current gain"), which depends only on the ratio of d-c to a-c turns. However, this is true only when the measured load current I_L is expressed in average values over one half cycle. Therefore, it is essential that a moving-coil instrument with a rectifier (Fig. 4.2) be

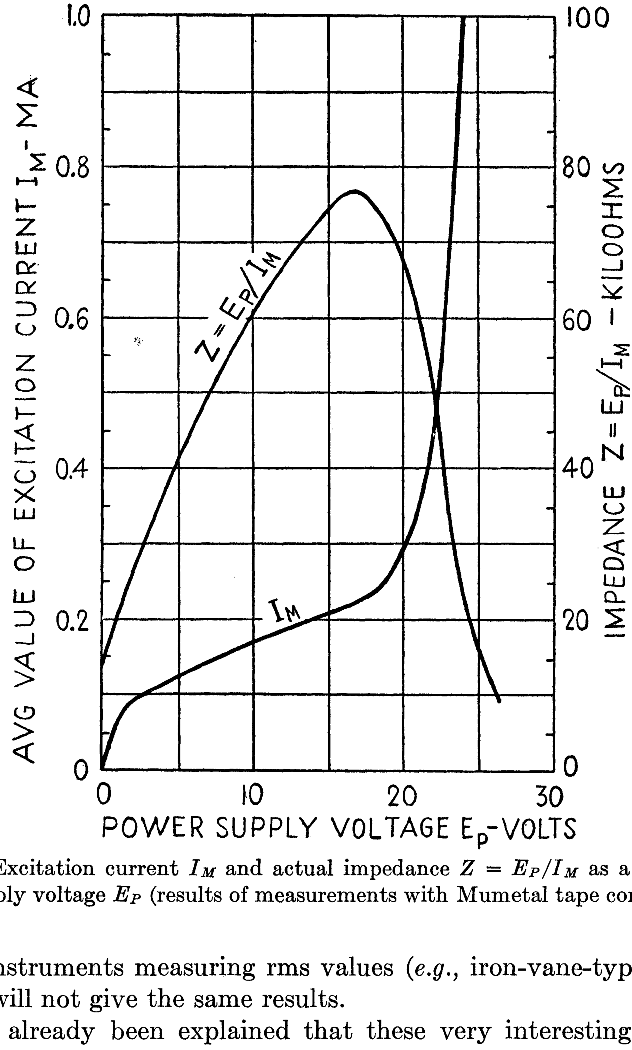


FIG. 4.7. Excitation current I_M and actual impedance $Z = E_P/I_M$ as a function of power-supply voltage E_P (results of measurements with Mumetal tape cores).

used. Instruments measuring rms values (*e.g.*, iron-vane-type milliammeters) will not give the same results.

It has already been explained that these very interesting facts can be used as a basic principle for the design of a so-called d-c instrument transductor to be especially applied for the measurement of very large d-c values in the range from 100 to 10,000 amp, preferably employing a bar-type transformer construction.

Alternatively, the a-c windings of the two saturable-reactor elements

(Fig. 4.2) may be connected in parallel, as shown in Fig. 3.5. The voltage across each element is then the supply voltage, and the number of turns $N'_L = N''_L$ has had to be doubled. Here the ratio between load

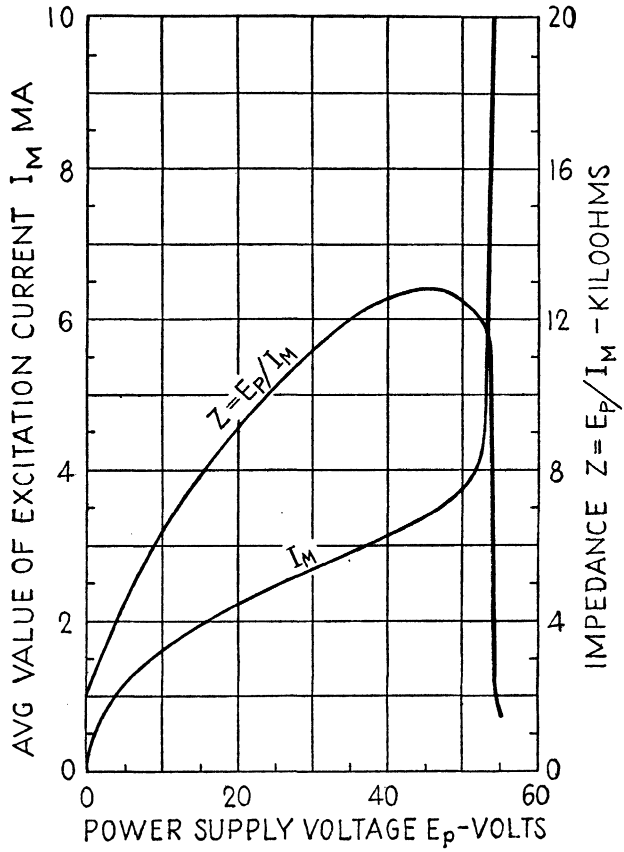


FIG. 4.8. Excitation current I_M and actual impedance $Z = E_P/I_M$ as a function of power-supply voltage E_P (results of measurements with Orthonol tape cores).

current I_L and control current I_C is given by the equation

$$\frac{I_L}{I_C} = 2\frac{N_C}{N_L} \tag{4.2}$$

In practice the arrangement with a-c windings in series (Fig. 4.2) is preferable for magnetic amplifiers because it has a lower time lag. But in special cases, *e.g.*, in the design of so-called direct-voltage transductors[11] for the measurement of high direct voltages, the parallel circuit (Fig. 3.5) has some advantages, as shown in Sec. 3.3.

Figures 4.7 (Mumetal cores) and 4.8 (Orthonol cores) show:

1. The average value of excitation current I_M (the load current I_L with control current $I_C = 0$) as a function of the sinusoidal power-supply voltage E_P
2. The actual impedance $Z = E_P/I_M$ as a function of E_P

Discussing the results of these measurements, as presented in Figs. 4.3 to 4.8, it was learned that:

1. Generally, the characteristic properties of the basic nonfeedback saturable-reactor circuit are far more pronounced when the rectangular-hysteresis-loop core material (Orthonol) is used.
2. With Orthonol cores the power-supply voltage E_P can be varied within considerably larger limits without an appreciable effect upon the current ratio I_L/I_C than in the case where Mumetal cores are used.
3. Maximum value of impedance $Z = E_P/I_M$ corresponding to the maximum value of a-c permeability occurs with $E_P = 16$ volts (Mumetal) or with $E_P = 45$ volts (Orthonol).
4. Therefore, the maximum value of E_P which can be applied to the Orthonol-core circuit (about 45 volts) is nearly three times as high as when the Mumetal-core circuit is used (about 16 volts), thus increasing substantially the power-handling capacity of a given weight of core material.

4.6. Generalized Performance Characteristics of Nonfeedback Magnetic-amplifier Circuits. Referring to Figs. 4.3 to 4.6, it is evident that the performance characteristics representing the results of various measurements have been plotted in actual voltage and current values. Such characteristics may be generalized by expressing the results of these measurements in terms of volts per turn, E_P/N_L, for the voltage axis, ampere-turns I_LN_L for the ordinates, and ampere-turns I_CN_C for the parameters of the d-c control windings.

Figure 4.9 shows, as a typical example, the conversion of the actual voltage-current characteristic of Fig. 4.5 into a generalized performance characteristic. It contains also a "load line" (corresponding to the load resistance $R_L = 440$ ohms), calculated in terms of R_L/N_L, the load being expressed in terms of R_L/N_L^2. By using this method of presentation, it is possible to make the characteristic general for particular constructions of core elements, and the information given by such a generalized characteristic can be adapted to special requirements by inserting actual values for the number of turns N_L and N_C.

4.7. Application of Load Lines for Determination of the Limiting Load. It has already been mentioned that the inherent current-transformer properties can be expected only in the *linear* region of the characteristics. If the actual value of load resistance R_L is too high, then the pulses of load current I_L cannot attain the necessary peak value, and the ampere-turn equivalence relation, as expressed by Eq. (4.1), cannot be satisfied.

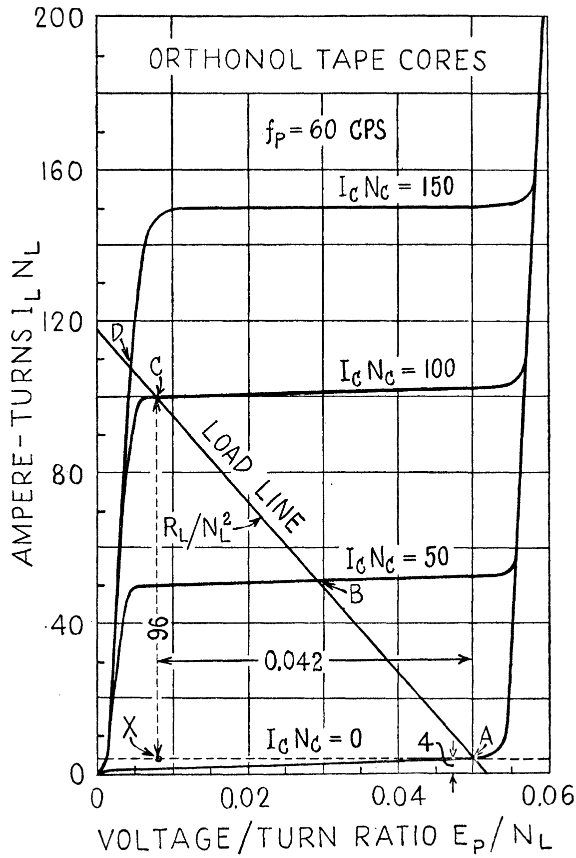


FIG. 4.9. Generalized performance characteristic of a nonfeedback magnetic-amplifier circuit with Orthonol tape cores (conversion of the actual voltage-current characteristic of Fig. 4.5). The "load line" *ABCD* corresponds to the load resistance R_L = 440 ohms.

Referring to Fig. 4.9, the operation of this limit can be illustrated in the following way:

The slope of the load line *ABCD* is such that the ratio

$$\frac{XA}{XC} = \frac{I_L R_L/N_L}{I_L N_L} = \frac{R_L}{N_L^2} \tag{4.3a}$$

represents the load, expressed in terms of R_L/N_L^2; actually, with

$$R_L = 440 \text{ ohms}$$

$$\frac{XA}{XC} = \frac{42 \text{ volts}/1{,}000 \text{ turns}}{96 \text{ amp-turns}} = 4.4 \times 10^{-4} \qquad (4.3b)$$

The position of the original operating point A (with $I_CN_C = 0$), near the bend of the zero-control-current characteristic, corresponds to the average value of the exciting current I_M (actually 4 ma) with the rated average value of the operating voltage E_A (actually 50 volts). It is very important that E_A be rated so that A is not too near the bend of the zero-control-current characteristic; otherwise, even very small variations of power-supply voltage E_P might produce large changes of exciting current I_M. For this reason, characteristics giving $I_M = f(E_P)$ and $E_P/I_M = f(E_P)$, as shown in Figs. 4.7 (Mumetal) and 4.8 (Orthonol), have proved to be very valuable for the design procedure of magnetic amplifiers. In the example of Fig. 4.6, the operating voltage E_A has been considerably reduced ($E_P = 40$ volts) so that variations of E_P within the limits of ± 25 per cent cannot jeopardize correct operation of the circuit.

When the control current I_C is increased, then the load current I_L will increase, but actual voltage across the a-c windings N'_L, N''_L will evidently decrease because of the increased voltage drop I_LR_L. In this case, the actual operating point will be given, at each value of control ampere-turns I_CN_C, by the intersection (points B, C, D) of the load line with the corresponding characteristic.

In the example of Fig. 4.9, the limiting value of control ampere-turns I_CN_C is a little more than 100 ($I_C = 100$ ma). If the control current I_C is now increased beyond this value, *e.g.*, up to 150 ma, so that $I_CN_C = 150$, then the voltage drop I_LR_L reduces the voltage across the a-c windings N'_L, N''_L to too small a value to permit the pulses of load current I_L to reach a high enough peak value. Thus, within the range of from 0 to 100 ma, there is a linear relationship between I_C, I_L, I_LR_L, and I_LN_L. However, this performance ceases shortly beyond $I_C = 100$ ma; a further increase in I_C will not produce any further increase in I_L. The lower the load resistance R_L, the later the limitation in load current I_L. This fact is also evident from the transfer characteristics giving $I_L = f(I_C)$ with R_L as a parameter (Fig. 4.6).

It is to be noted that such a load line as is drawn in Fig. 4.9 will be a straight line only (1) as long as average values of voltages and currents are used, and (2) if the load is a pure ohmic resistance. Otherwise, the load line becomes an ellipse. However, to a certain extent, with other loads it is usually sufficient to treat the load line as straight, because this graphical procedure is used in most cases for an approximate evaluation only.

Evidently, it is possible to shift the effective operating point from the original position (A) to another intermediate point, *e.g.*, to point B. This can be achieved by applying an additional magnetomotive force (bias effect) on the cores,

1. By means of a substantially constant bias current I_B on which the control current I_C is superimposed in the common control windings N'_C, N''_C
2. By means of a substantially constant bias current I_B, applied to separate "bias windings" N'_B, N''_B having a proper number of turns and being series-connected in such a way that no voltages of fundamental frequency can appear in the bias circuit

In both cases, the resultant magnetomotive force will be proportional to the actual difference between bias current I_B and control current I_C. In the first case, the resultant magnetomotive force will be represented by the actual ampere-turns $(I_B \pm I_C)N_C$, and in the second case, the resultant magnetomotive force will be proportional to the actual ampere-turns $(I_BN_B \pm I_CN_C)$. For example, if, with $R_L/N_L^2 = 4.4 \times 10^{-4}$ (Fig. 4.9), a constant bias current $I_B = 50$ ma is applied to separate d-c bias windings N'_B, N''_B, each having 1,000 turns, the effective operating point will move from its original position (A) to the point B. In this case, a *reversible* control current of ± 50 ma would cause the load current I_L to operate on either side of this point. This possibility of obtaining polarity-sensitive performance characteristics is of paramount importance with regard to those numerous applications in which the control current I_C will change its direction, as in magnetic servo amplifiers and other similar devices. Such arrangements having special bias circuits will be described in later chapters.

4.8. Literature concerning Theoretical Investigations on Nonfeedback Magnetic-amplifier Circuits. Besides the authors already referred to in Chaps. 3 and 4, contributions concerning theoretical investigations on the operating conditions of nonfeedback magnetic-amplifier circuits have been made by L. Dreyfus,[12] H. Schunk,[13] A. Boyajian,[14] F. Schröter,[15] O. E. Charlton and J. E. Jackson,[16] T. Wasserrab,[17] G. Hauffe,[18] G. F. Partridge,[19] W. Hartel,[20] K. Reuss,[21] A. U. Lamm,[5] W. Schilling,[22] H. S. Kirschbaum and E. L. Harder,[23] S. E. Tweedy,[24] E. H. Frost-Smith,[25] S. E. Hedström and L. F. Borg,[26] A. G. Milnes,[27] H. M. Gale and P. D. Atkinson,[28] D. W. VerPlanck and M. Fishman,[29] J. H. Reyner,[30] W. J. Dornhoefer and V. H. Krummenacher,[31] and F. E. Butcher and R. Willheim.[32]

Further literature concerning the development of various nonfeedback types of magnetic-amplifier circuits and special applications will be given in the next chapter.

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CHAPTER 5

Nonpolarized, Polarized, and Duodirectional Types of Nonfeedback Magnetic-amplifier Circuits

5.1. Classification of Nonfeedback Magnetic-amplifier Circuits. The basic saturable-reactor circuit, as described in Chaps. 3 and 4, represents a *nonpolarized* magnetic amplifier which is unable to discriminate between "positive" and "negative" control currents. Referring to Figs. 4.4 and 4.6, it is evident that exactly the same load current I_L would be obtained with "reversed" control current I_C. Thus, the complete transfer characteristic (Fig. 5.1*a*) is symmetrical.

However, with regard to various special applications, it is necessary to provide some discrimination between positive and negative control currents. It has already been mentioned in Sec. 4.7 that it is possible to obtain polarity-sensitive performance characteristics by applying an additional magnetomotive force (bias) on the cores. In this case, an asymmetrical transfer characteristic (Fig. 5.1*b*) will be obtained so that a positive control current $+I_C$ causes an increase of load current I_L while a negative control current $-I_C$ causes a decrease of I_L. Evidently, I_L will never change its direction. Such an arrangement having an asymmetrical transfer characteristic, from application of a bias magnetization, represents a *polarized* magnetic amplifier.

The statement that a magnetic-amplifier circuit is polarity-sensitive does not necessarily mean that this amplifier is polarized by applying an additional bias magnetization on the cores. Figure 5.15 illustrates the possibility of obtaining the desired discrimination between positive and negative control currents by means of two nonpolarized magnetic-amplifier units operating in connection with a switch, relay, a special rectifier circuit, or other equivalent device. In contrast to such arrangements, the term "polarized" shall define those types of circuits which are inherently polarity-sensitive as a result of the application of a bias magnetization.

According to the fundamental principles of servomechanisms, a magnetic servo amplifier must operate as a balance detector, in a similar way as a galvanometer in a Wheatstone-bridge or a potentiometer circuit does. In this case, the load current I_L must change its direction according to the direction of control current I_C, as illustrated in Fig. 5.1*c*. In order to obtain magnetic amplifiers having this type of transfer characteristic, it is necessary to use two saturable-reactor units in a balanced arrangement, preferably in a push-pull circuit. Such an arrangement having response characteristics of a real balance detector represents a *duodirectional* magnetic amplifier, or a "magnetic amplifier of the balance detector type."[1]

5.2. Nonpolarized Circuits (Neutral Type[2,3]). Referring to the actual performance characteristics of nonfeedback circuits, as shown in Figs. 4.3 to 4.9, it is evident that the load current $I_L = I_{L0} + I_{LC}$ is not zero when there is no control current I_C. Actually, with $I_C = 0$, the magnetizing current I_M (Figs. 4.7 and 4.8) flows through the load resistor R_L and produces the corresponding load voltage $I_M R_L$.

Application of a control current I_C causes the load current to increase from I_{L0} to I_L. The increment $I_{LC} = I_L - I_{L0}$ represents the "load component" flowing in response to changes in control current I_C. It is to be noted that, even when high-permeability core material is used, the actual value of the excitation-current component I_{L0} may be a substantial percentage of that of the load component I_{LC}.

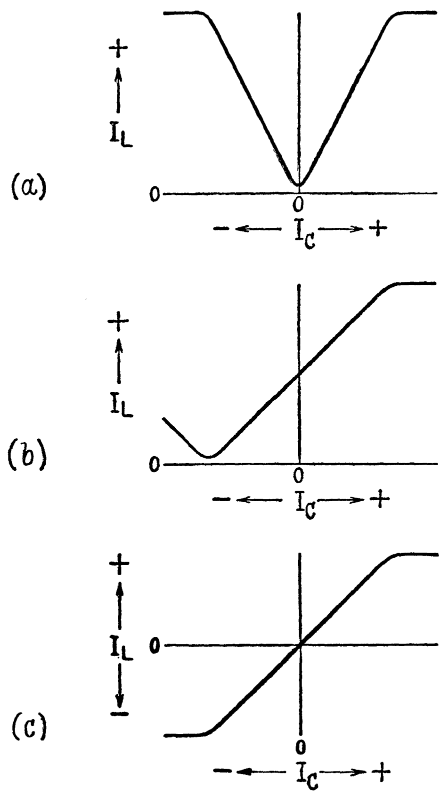

FIG. 5.1. Three typical transfer characteristics of magnetic-amplifier circuits: (*a*) complete characteristic $I_L = f(I_C)$ of a nonpolarized circuit; (*b*) characteristic $I_L = f(I_C)$ of a polarized circuit using an additional bias magnetization; (*c*) characteristic $I_L = f(I_C)$ of a duodirectional circuit of the balance-detector type.

Thus, with regard to many practical applications of nonfeedback circuits, it is necessary to provide some special circuits to compensate for undesired effects of the magnetization current I_M. For example, when two or more nonfeedback circuits are connected in cascade (multistage circuits), then it is necessary to compensate for the effect of the magnetizing current of one stage on the following stage. In such arrangements, the entire design revolves around the essential problem that $I_L = 0$ if $I_C = 0$, and that fulfillment of this condition must be mostly independent of changes in magnitude and frequency of the a-c power-supply voltage, changes in ambient temperature, etc.

There are several different means which may be used for the solution of this basic problem. However, in any case, the fundamental principle

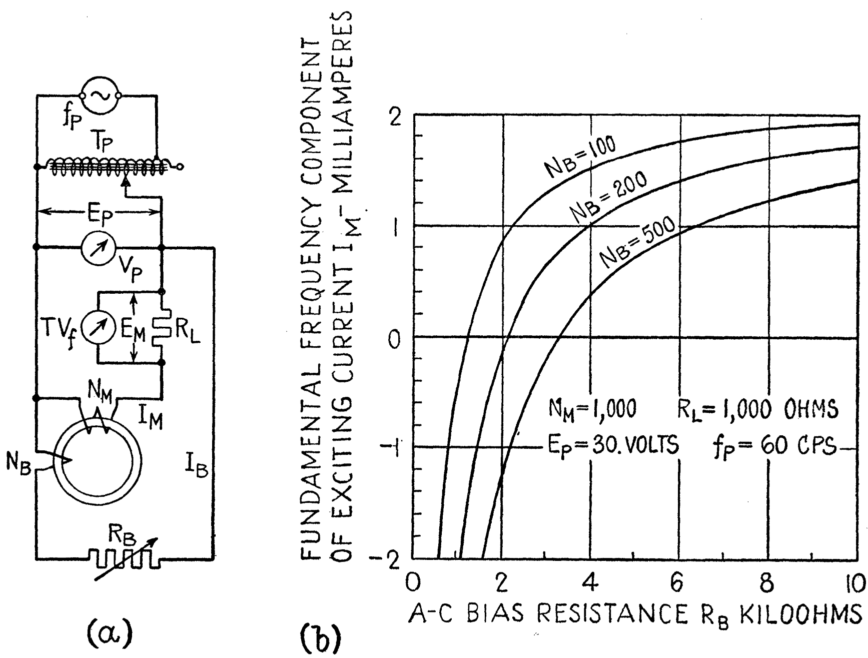

FIG. 5.2. Compensation of the magnetizing current of a saturable-reactor element by means of an a-c bias circuit. (*Corresponding to "Magnetischer Verstärker," Swiss patent* 228,534, 1942.) (*a*) Diagram of the test circuit; (*b*) results of measurements.

consists in applying an additional current which balances out the undesired effects of the magnetization current I_M.

In the literature, the excitation-current component I_{L0} (the load current I_L with no-signal conditions) is sometimes termed "quiescent-current value," "Q current," or "standing current." This value corresponds to the "quiescent point" or "operating point Q" of a three-element vacuum tube.

a. Compensation of the Magnetizing Current by Means of an A-C Bias Circuit.[4] In some cases, it will be sufficient to balance out the fundamental-frequency component of the magnetizing current I_M. This can be achieved in a simple way (Fig. 5.2) by applying an a-c bias current

I_B flowing through a separate bias winding N_B with a series resistor R_B. It is to be noted, however, that the number of turns of N_B must always be considerably smaller than that of the load winding N_M to avoid undesired effects of the voltage which will be induced in N_B by transformer action.

Figure 5.2*a* shows the test circuit which has been used for measuring the actual value of the fundamental wave of magnetizing current I_M as a function of the a-c bias resistor R_B with the number of turns of N_B as a parameter. This circuit contains:

1. An Orthonol 2-mil tape core having a load winding N_M (1,000 turns) and the a-c bias winding N_B (100, 200, or 500 turns)
2. An ohmic load resistor $R_L = 1{,}000$ ohms with a phase-sensitive vacuum-tube voltmeter TV_f which is inherently frequency-selective, so that only the fundamental-frequency component of the actual voltage drop $E_M = I_M R_L$ will be measured
3. A regulating transformer T_P with a-c voltmeter V_P for varying and indicating the actual value of the power-supply voltage E_P (frequency $f_P = 60$ cps)

Figure 5.2*b* showing results of measurements makes it evident that the fundamental-frequency component of magnetizing current I_M can be reduced, completely compensated, or even overcompensated when such an a-c bias circuit is used. Thus, for example, a separately excited wattmeter, an induction-type two-phase motor, or some other frequency-selective device responding only to the fundamental wave will not respond if the fundamental-frequency component of I_M is compensated in this way.

Because the fundamental wave of I_M is nearly in phase with the power-supply voltage E_P, when rectangular-hysteresis-loop core material is used, sufficiently good compensation can be achieved by means of a simple ohmic resistor R_B. In this case, the bias current I_B is also nearly in phase with E_P, and it is generally not necessary to provide additional means for adjusting the phase angle between I_B and E_P. However, with other core materials, it may be possible to obtain complete compensation of I_M only if the phase relationship between I_B and E_P can be varied, *e.g.*, by means of a properly rated reactor or condenser in the a-c bias circuit. In this case, additional adjustment must be made so that the fundamental waves of I_M and I_B are in phase, in order to bring I_M to zero.

Of course, an *asymmetrical* balanced circuit, as shown in Fig. 5.2*a*, will be sufficiently balanced only within a certain range of power-supply voltage. When E_P is varied within a limited range of about ± 5 per cent, then the balance conditions will still be satisfied. However, with larger

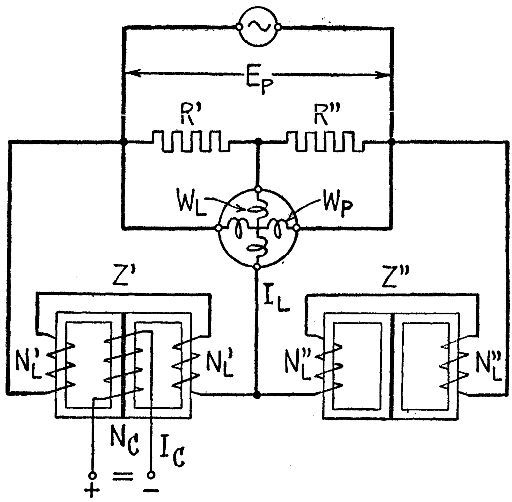

FIG. 5.3. Symmetrical Wheatstone-bridge circuit with two reactor units and a separately excited electrodynamic instrument, for telemetering of direct currents. (*G. Keinath.*[5])

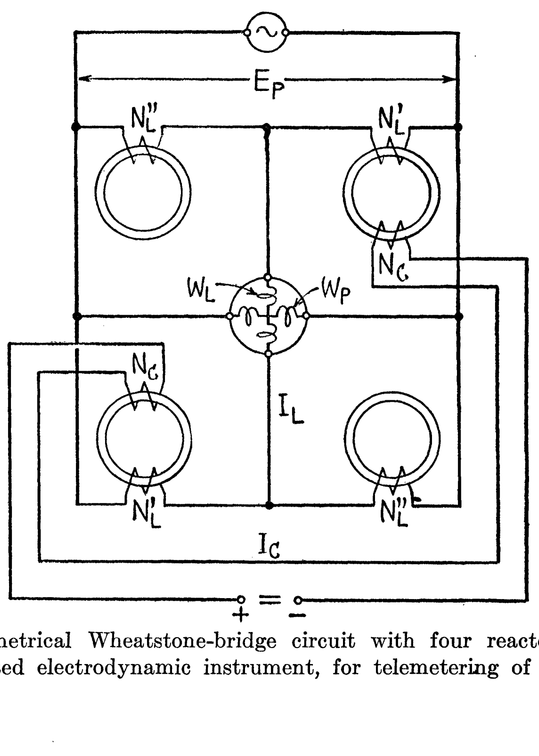

FIG. 5.4. Symmetrical Wheatstone-bridge circuit with four reactor units and a separately excited electrodynamic instrument, for telemetering of direct currents. (*G. Keinath.*[5])

variations of E_P, the actual changes of I_M may reach excessive values. In such cases, *symmetrical* arrangements of the bridge or differential-transformer type will give considerably improved performance characteristics.

b. Compensation of the Quiescent Current by Means of a Symmetrical Wheatstone-bridge Circuit. In 1922, G. Keinath[5] introduced several

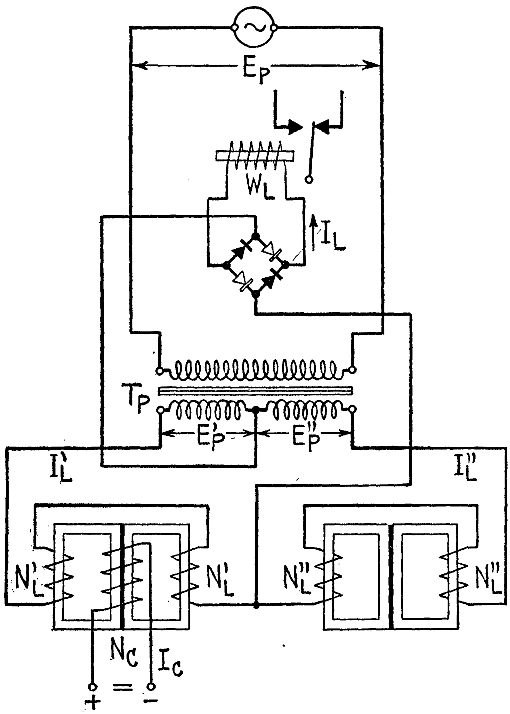

FIG. 5.5. Symmetrical Wheatstone-bridge circuit with two reactor units, a center-tapped power-supply transformer, and a rectifier-relay device. (*A. S. FitzGerald, United States patent* 2,027,311, 1932.)

saturable-reactor types of symmetrical bridge circuits, preferably operating in connection with a separately excited electrodynamic instrument, for telemetering of direct currents. The basic idea of these arrangements consists in adjusting the impedances of the four branches of the bridge network in such a way that the load current is zero when there is no control current.

One of these circuits (Fig. 5.3) contains (1) two equally rated reactor units Z', Z'' with a-c windings N'_L, N''_L, (2) two ohmic resistors R', R'', and (3) a separately excited electrodynamic indicating or recording instrument with windings W_P and W_L. The first reactor unit Z' has a control winding N_C carrying the direct current I_C to be measured; the second reactor unit Z'', being a dummy, or counterpoise, has an identical magnetic core and equally rated a-c windings ($N'_L = N''_L$), but no control winding. With $I_C = 0$, the bridge is completely balanced, and the load current I_L flowing through W_L is zero. However, when a control current I_C is applied, then the electrodynamic instrument measures the net increment of load current I_L.

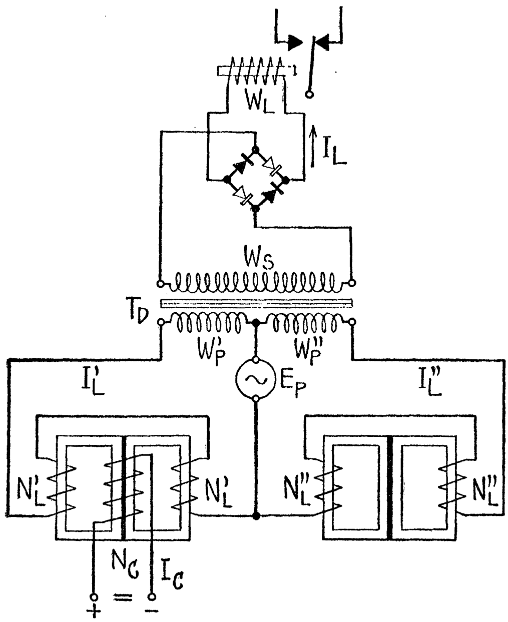

FIG. 5.6. Symmetrical differential-transformer circuit with two reactor units and a rectifier-relay device. (*F. W. Lee.*[8])

Another symmetrical bridge circuit (Fig. 5.4), as described by Keinath,[6] contains four equally rated reactor units with a-c windings $N'_L = N''_L$. The impedances of two opposite branches of the bridge are controlled by means of control windings N_C carrying the direct current I_C to be measured. Various modifications of the control circuit are possible. For example, three equally rated d-c control windings (N_C), properly insulated from each other, may be used to measure the sum of the actual values of three variable direct currents.

In 1932, A. S. FitzGerald[7] devised a symmetrical bridge circuit (Fig. 5.5) containing (1) two equally rated reactor units with a-c windings N'_L, N''_L, (2) a center-tapped power-supply transformer T_P, and (3) a relay with full-wave rectifier. With $E'_P = E''_P$, the impedances of the two reactor units can be adjusted so that the load current $I_L = I'_L - I''_L$ is zero, if control current $I_C = 0$. However, when I_C is applied, then the relay receives the net increment of load current I'_L.

c. Compensation of the Quiescent Current by Means of a Symmetrical Differential-transformer Circuit. This circuit (Fig. 5.6), devised in 1926 by F. W. Lee,[8] is based upon the principle that with $I_C = 0$ the load current $I_L = \text{const} \times (I'_L - I''_L)$ will also be zero, if the actual impedances of the two reactor units are properly adjusted so that the opposing ampere-turns of the primary windings W'_P and W''_P of the differential

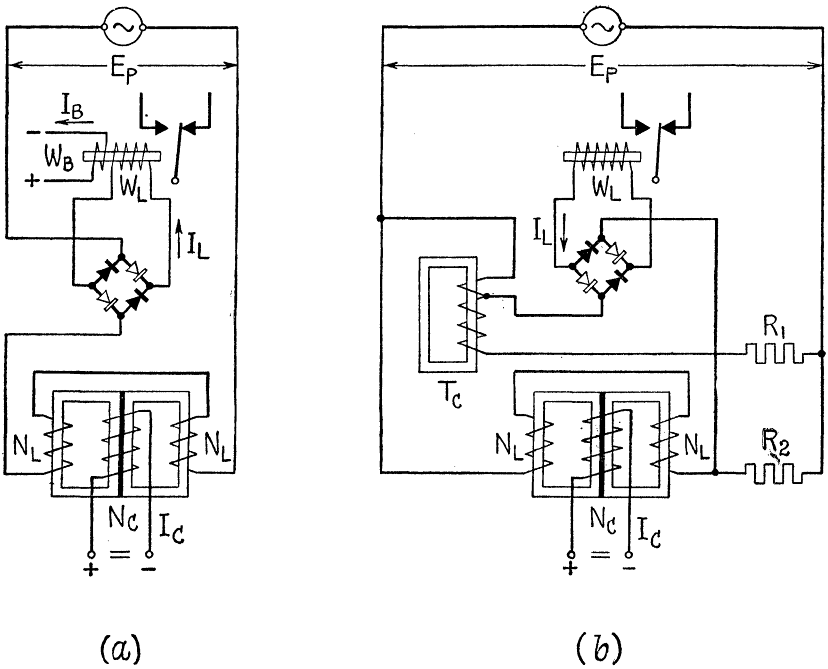

FIG. 5.7. Compensating for the quiescent current in a simple arrangement containing a single saturable-reactor unit by (*a*) providing a bucking winding (W_B) on the relay (*A. S. FitzGerald, United States patent* 2,027,311, 1932); (*b*) using an asymmetrical bridge circuit (*A. S. FitzGerald, United States patent* 2,461,046, 1946).

transformer T_D are equal. In this circuit too, the first reactor unit has a control winding N_C, while the second reactor, being a counterpoise, has an identical magnetic core and equally rated a-c windings ($N'_L = N''_L$), but no control winding. When a control current I_C is applied, then the circuit will become unbalanced, and a corresponding load current I_L, a function of I_C, will flow through the secondary winding W_S and the full-wave rectifier feeding the relay winding W_L.

d. Compensating for the Quiescent Current by Means of a Bucking Winding on the Load. Figure 5.7*a* illustrates the possibility of providing a bucking winding W_B on the relay (A. S. FitzGerald,[9] 1932). This winding carries a constant direct current I_B adjusted so as to produce ampere-turns equal and opposite to those which are due to the current I_L in the main winding W_L. If $I_C = 0$, the actual ampere-turns $I_L W_L - I_B W_B$ are zero. When a control current I_C is applied, then the relay will receive the ampere-turns $I_L W_L - I_B W_B$, the load current I_L being a function of I_C.

e. Compensating for the Quiescent Current by Means of an Asymmetrical Bridge Circuit. In 1946, A. S. FitzGerald[10] showed that it is possible

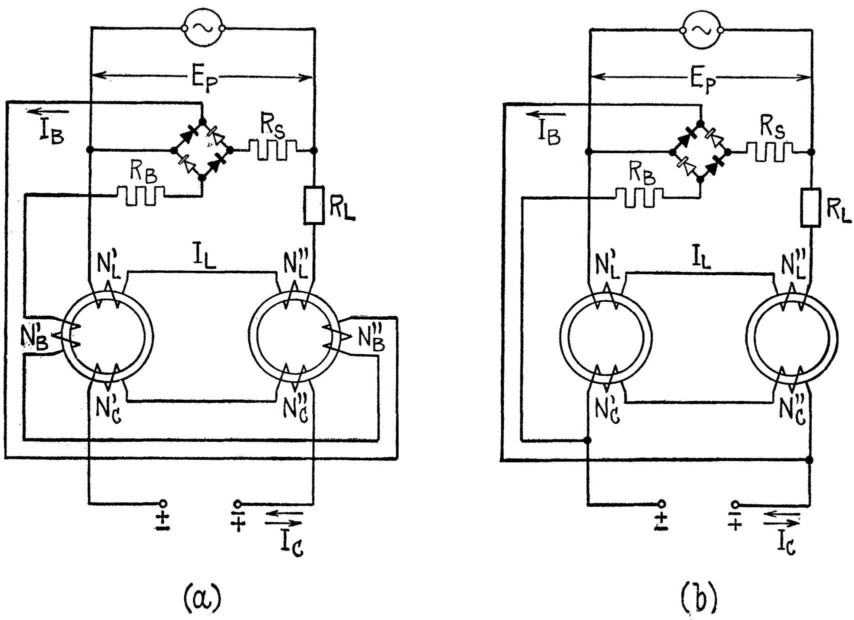

FIG. 5.8. Two examples of polarized magnetic-amplifier circuits (bias type), applying an additional magnetomotive force on the core elements (*a*) by means of a direct bias current I_B, applied to separate windings N'_B, N''_B; (*b*) by means of a direct bias current I_B on which the control current $+I_C$ is "galvanically" superimposed in the common control windings N'_C, N''_C acting simultaneously as bias windings.

to obtain excellent results by using an asymmetrical bridge network (Fig. 5.7*b*), which comprises, in addition to the single saturable-reactor unit with the usual a-c and d-c windings N_L and N_C, a transformer T_C and two fixed resistors R_1 and R_2. Thus, the second reactor unit, acting as a counterpoise (Figs. 5.3, 5.5, and 5.6) and having also a nickel-iron-alloy core, has been replaced by a less expensive transformer which is made of regular-grade transformer-type core material. It was found that the use of nickel-iron alloy of the same type as is used in the d-c-controlled reactor does not bring any significant improvement.

5.3. Polarized Circuits (Bias Type). Referring to the performance characteristic of Fig. 4.9, it is evident that the effective operating point can be shifted from the original point (A) to another intermediate point of the load line, *e.g.*, to point B. This may be achieved, as illustrated in Fig. 5.8, by applying an additional magnetomotive force (bias) on the core elements in two ways. In both cases, bias current I_B is supplied through a full-wave-rectifier circuit with fixed resistors R_S, R_B from the power-supply voltage E_P. Thus, I_B is proportional to E_P. The resultant magnetomotive force, applied to the core elements, will

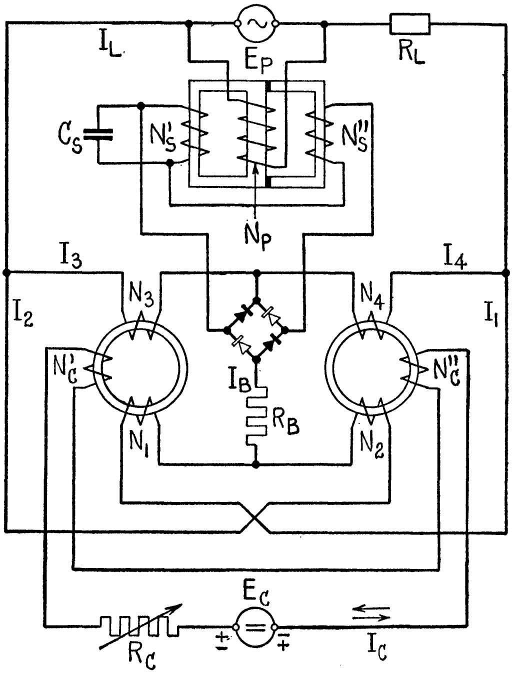

FIG. 5.9. Supplying the direct bias current I_B to the a-c windings of an auto-connected bridge-type circuit by means of a simple constant-voltage transformer.

be proportional to the actual difference between I_B and I_C. In the first case (Fig. 5.8*a*) the resultant magnetomotive force will be proportional to the actual ampere-turns ($I_BN_B \pm I_CN_C$), and in the second case (Fig. 5.8*b*) the resultant magnetomotive force will be represented by the actual ampere-turns $(I_B \pm I_C)N_C$. For example, if with $\frac{R_L}{N_L^2} = 4.4 \times 10^{-4}$ (Fig. 4.9), a bias current $I_B = 50$ ma is applied to separate d-c bias windings N_B', N_B'' (Fig. 5.8*a*), each having 1,000 turns, the effective operating point would move from A to B. In this case, a reversible control current $I_C = \pm 50$ ma would cause the load current I_L to operate on either side of the new working point B. Because an asymmetrical transfer characteristic, as shown in Fig. 5.1*b*, will be obtained in this way, a positive control current $+I_C$ causes an increase of I_L, while a negative control current $-I_C$ causes a decrease of I_L. However, I_L will never change its direction.

Of course, it is also possible to supply the direct bias current I_B to the a-c windings of an auto-connected bridge-type circuit (Fig. 5.9). In

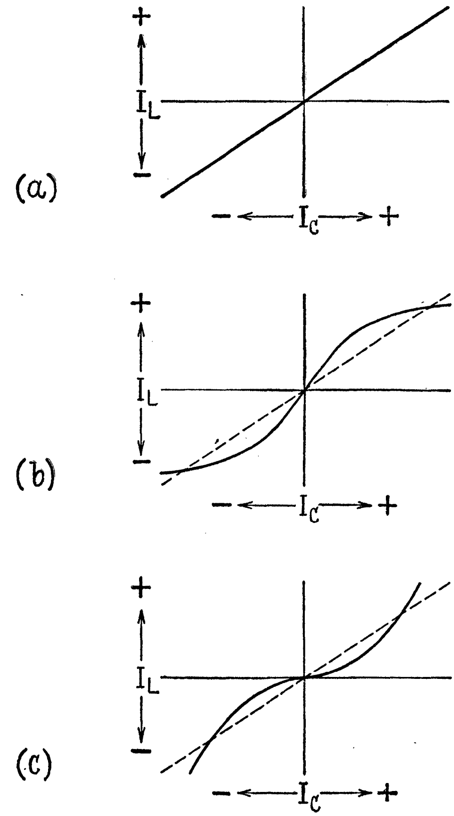

FIG. 5.10. Three typical transfer characteristics of special interest in the field of duodirectional magnetic-amplifier circuits: (*a*) and (*b*) are very practicable, but (*c*) is generally unsuitable.

this example, a simple constant-voltage transformer N_P, N'_S, N''_S with shunt condenser C_S is provided, to make I_B independent of fluctuations in power-supply voltage E_P.

Another method for producing constant bias magnetization in polarized types of magnetic amplifiers consists in applying an arrangement of permanent magnets which creates a constant magnetization in the core elements (P. H. Dowling,[11] 1929). In special cases, the use of permanent magnets for biasing purposes may have some advantages (E. H. Frost-Smith,[12] 1948).

5.4. Duodirectional Circuits (Push-Pull Type). In magnetic amplifiers of the balance-detector type, the load current I_L must change its direction according to the reversible direction of control current I_C, as illustrated in Fig. 5.1*c*. In order to obtain this type of transfer characteristic, it is necessary to use two equally rated saturable-reactor units in a balanced arrangement, preferably in a push-pull circuit.

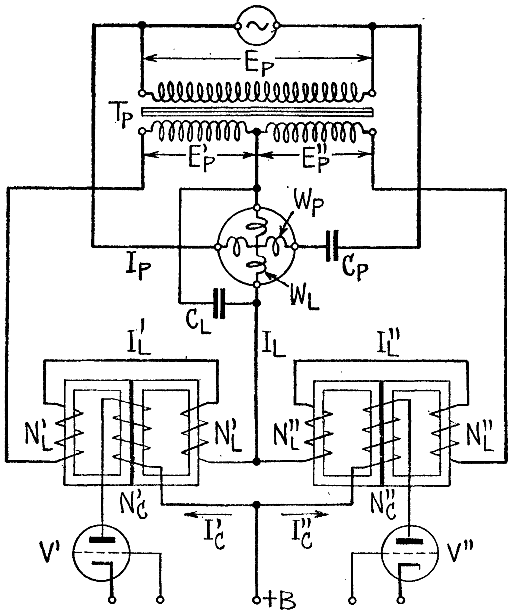

FIG. 5.11. Replacing the output-stage tubes of an electronic servo amplifier by two saturable-reactor units operating in a simple nonfeedback push-pull circuit without additional bias magnetization.

Figure 5.10 illustrates three typical transfer characteristics of special interest in the field of duodirectional magnetic-amplifier circuits.

Figure 5.10*a* shows a linear characteristic curve, the output current $\pm I_L$ being exactly proportional to the input current $\pm I_C$.

Figure 5.10*b* shows a nonlinear characteristic curve, following a nearly linear law with low values of input current $\pm I_C$, but having considerable deviations with higher input-current values.

Figure 5.10*c* shows another nonlinear characteristic curve, having reduced steepness with low values of input current $\pm I_C$, but following a nearly linear law with higher input-current values.

It is to be noted that transfer characteristics, as illustrated in Fig. 5.10*a* and *b*, are very practicable, but that characteristics corresponding to Fig. 5.10*c* are generally unsuitable. In principle, sensitivity of a balance detector must be especially high with low values of input current and may be reduced with higher input-current values. Therefore, a characteristic curve like that shown in Fig. 5.10*c* must be considered as not practicable.

a. Saturable-reactor-type Push-Pull Circuit without Additional Bias Magnetization.[13] Figure 5.11 illustrates the possibility of replacing the d-c-controlled output-stage tubes of an electronic servo amplifier by two saturable-reactor units operating in a simple nonfeedback push-pull circuit. This arrangement contains:

1. A center-tapped power-supply transformer T_P providing two equal voltages E'_P, E''_P, which are proportional to the a-c supply voltage E_P (frequency $f_P = 60$ or 400 cps)
2. Two equally rated saturable-reactor units with a-c load windings N'_L, N''_L, and d-c control windings N'_C, N''_C
3. The amplifier field winding W_L of a two-phase induction-type reversible motor with shunt condenser C_L
4. The line field winding W_P of this motor with series condenser C_P
5. The plate circuit ($+B$) of two triodes V', V'' operating in push-pull so that I''_C decreases when I'_C increases, and vice versa

The mode of operation of this circuit may be illustrated by Table 5.1.

TABLE 5.1

I'_C (ma)	I''_C (ma)	$I'_C - I''_C$ (ma)	$I_L = I'_L - I''_L$ (ma)
10.0	0	+10.0	+100
7.5	2.5	+5.0	+50
5.0	5.0	0	0
2.5	7.5	−5.0	−50
0	10.0	−10.0	−100

Evidently, the circuit has the desired duodirectional transfer characteristic of Fig. 5.1*c*. However, an additional bias magnetization is not necessary in this case, because both saturable-reactor units are already properly biased by the quiescent currents $I'_C = I''_C = 5.0$ ma. It is to be noted that such an arrangement responds to the actual difference $I'_C - I''_C$ of the two plate-current values I'_C and I''_C, which are functions of the actual error-signal value.

Proper phase relationship between the currents I_P and I_L and maximum torque conditions of the two-phase motor will be obtained by means of the condensers C_P and C_L. Application of the power-supply transformer T_P makes it possible to match the impedance values E'_P/I'_L, E''_P/I''_L, and the actual impedance of the motor circuit, so that optimum operating conditions will be secured.

b. Saturable-transformer-type Push-Pull Circuit without Additional Bias Magnetization. Proper impedance matching can be secured without

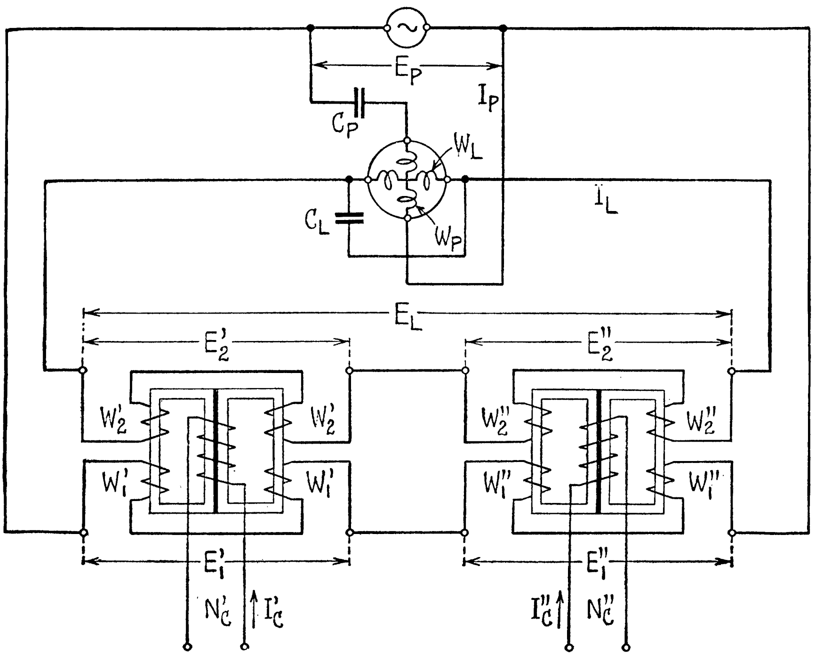

FIG. 5.12. Saturable-transformer-type nonfeedback push-pull circuit without additional bias magnetization.

application of a center-tapped power-supply transformer by providing two saturable-transformer units, as shown in Fig. 5.12. Here, the primary windings W'_1, W''_1 and the secondary windings W'_2, W''_2 are equally rated and series-connected in such a way that push-pull action is obtained (Table 5.2).

TABLE 5.2

I'_C (ma)	I''_C (ma)	$I'_C - I''_C$ (ma)	$E_L = E'_2 - E''_2$ (volts)
10.0	0	+10.0	+100
7.5	2.5	+5.0	+50
5.0	5.0	0	0
2.5	7.5	−5.0	−50
0	10.0	−10.0	−100

In this case too, an additional bias magnetization is not necessary, because both saturable-transformer units are already properly biased by the quiescent currents $I'_C = I''_C = 5.0$ ma. The circuit responds to the actual difference $I'_C - I''_C$, so that a duodirectional type of transfer characteristic (Fig. 5.1c), will be obtained.

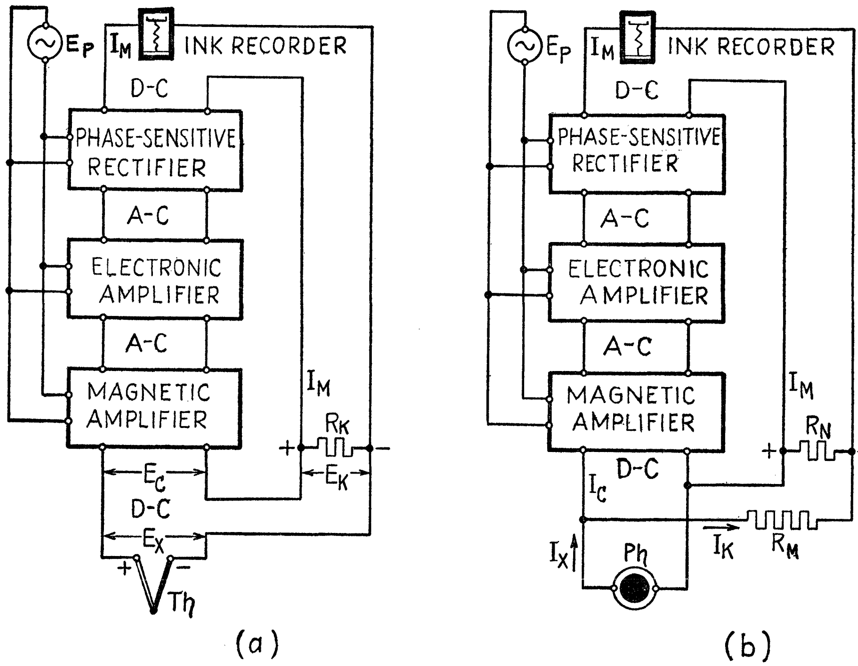

FIG. 5.13. Carrier-frequency type of d-c amplifier, which uses a duodirectional nonfeedback magnetic-amplifier circuit as a d-c/a-c converter or "magnetic modulator." (*W. A. Geyger.*[14]) (*a*) Measurement of small direct voltages (E_X) produced by thermocouple *Th*; (*b*) measurement of small direct currents (I_X) produced by a barrier-layer photoelectric cell *Ph*.

c. Saturable-reactor-type Push-Pull Circuit with Additional Bias Magnetization. In 1944, the author[14] described a carrier-frequency type of d-c amplifier (Fig. 5.13) which uses a duodirectional nonfeedback magnetic-amplifier circuit as a d-c/a-c converter or modulator. Combination of this arrangement with a moving-coil ink recorder offers the possibility of making a permanent record of very small direct voltages E_X (Fig. 5.13*a*) or currents I_X (Fig. 5.13*b*) created, for instance, by thermocouples, photoelectric cells, or radiation pyrometers.

Referring to Fig. 5.13*a*, it is to be noted that the voltage E_X to be measured (0.50 mv) and the opposing compensating voltage drop $E_K = I_M R_K$ (0.49 mv) give the resultant input voltage $E_C \doteq E_X - E_K$

(0.01 mv). This very small d-c voltage E_C is converted by means of the magnetic-amplifier circuit of Fig. 5.14 into a considerably larger a-c voltage $I_L R_L$ (about 2 mv), which is applied to a conventional type of three-stage electronic amplifier. The a-c output of this amplifier (voltage gain about 2,000) is converted into a direct current I_M by means of a phase-sensitive rectifier or demodulator. Because voltage E_X is nearly

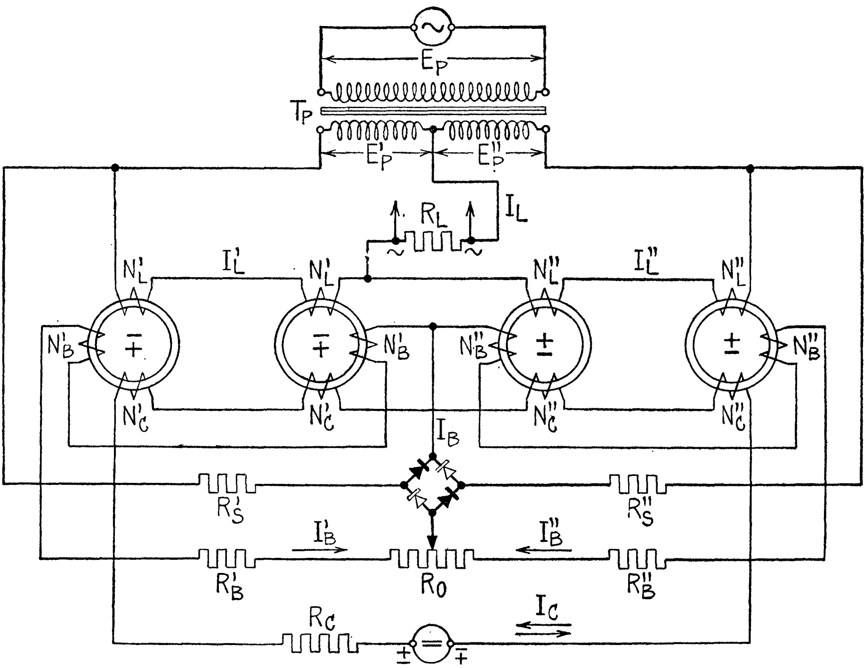

FIG. 5.14. Duodirectional nonfeedback magnetic-amplifier circuit operating as a "magnetic modulator" in the d-c amplifier arrangement of Fig. 5.13. (*W. A. Geyger.*[14])

balanced by the opposing voltage drop $E_K = 0.98E_X$, the effective input impedance is considerably increased (about fifty times), and there is a linear relationship between I_M and E_X, as a result of the very large amount (about 98 per cent) of degenerative feedback in the d-c input circuit.

Figure 5.13*b* illustrates the possibility of measuring very small direct currents in a similar way. In this case, the current I_X to be measured and the superimposed compensating current $I_K = I_M R_N/(R_N + R_M)$ give the resultant input current $I_C = I_X - I_K$; and the magnetic-amplifier circuit of Fig. 5.14 converts this very small direct current I_C into a corresponding a-c voltage, which is applied to the electronic amplifier. Because current I_X is nearly balanced by the superimposed current $I_K = 0.98I_X$, the effective input impedance is considerably reduced (about fifty times), and there is a linear relationship between I_M and I_X, as a result of the very large amount (about 98 per cent) of degenerative feedback in the d-c input circuit.

The duodirectional nonfeedback magnetic-amplifier circuit of Fig. 5.14 has proved to be very suitable for this purpose. This circuit uses

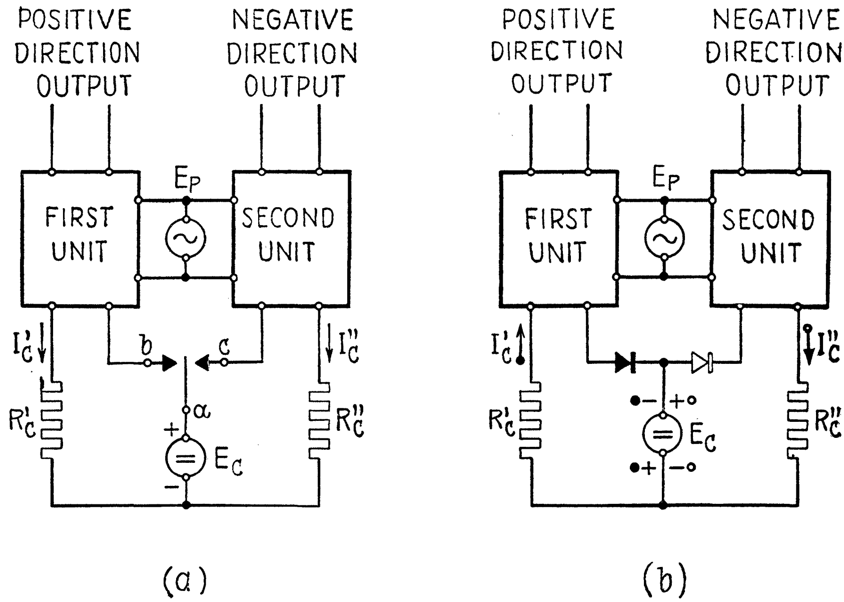

FIG. 5.15. Polarity-sensitive circuits with two nonpolarized magnetic-amplifier units operating in connection with (a) a switch or relay or (b) a special rectifier circuit. (*A. S. FitzGerald.*[15])

an additional bias magnetization, because the d-c quantity to be measured (E_X or I_X in Fig. 5.13) varies between zero and a certain maximum value corresponding to full-scale deflection of the ink recorder. The d-c bias circuit contains (1) bias windings N'_B, N''_B, (2) a full-wave dry-disk rectifier, (3) fixed series resistors R'_S, R''_S, R'_B, R''_B, and (4) a potentiometer-type resistor R_0 for zero adjustment: $I_M = 0$ with $E_X = 0$ or $I_X = 0$. Application of a constant-voltage transformer (T_P) is advisable to make the drift rate of the arrangement as small as possible. D-c amplifiers, as shown in Fig. 5.13, change the d-c quantity to be measured (E_X or I_X) into an a-c voltage of proportional magnitude in the duodirectional magnetic-amplifier circuit acting as a modulator. This a-c voltage may readily be amplified by an ordinary a-c amplifier. As an a-c amplifier is inherently more stable than a d-c amplifier, it is comparatively easy to maintain a very constant adjustment of the system if such a "magnetic modulator" is used.

5.5. Polarity-sensitive Circuits without Bias Magnetization.[15] It has already been mentioned that polarity-sensitive response characteristics can also be obtained by means of two nonpolarized magnetic-amplifier units operating in connection with a switch, a special rectifier circuit, or other equivalent devices.

Figure 5.15*a* illustrates the possibility of devising such an arrangement which gives two separated outputs, one for the first switch position $b - a$ and one for the second switch position $c - a$, although the control voltage E_C does not change its polarity. The two nonpolarized magnetic-amplifier units (without bias magnetization) are operated alternately, either by control current I'_C (positive-direction output), or by I''_C (negative-direction output). Proper gain adjustment can easily be achieved by using two separate series resistors R'_C and R''_C.

The arrangement of Fig. 5.15*b* gives two separated outputs, one for positive direction and one for negative direction of reversible control voltage E_C. Discrimination between positive and negative polarity of E_C is obtained by means of two half-wave rectifiers, which act alternately, corresponding to the actual polarity of E_C. The two nonpolarized magnetic-amplifier units (without bias magnetization) are operated and adjusted in the same way as the arrangement of Fig. 5.15*a*.

A. S. FitzGerald[15] has stressed that such polarity-sensitive magnetic-amplifier circuits with two separated outputs, one for a positive signal and one for a negative signal, have the following advantages:

1. "Over wide ranges of signal level, output is delivered by one side only, the current in the other side being of negligible magnitude. When the signal polarity is reversed, the outputs of the two units likewise change over."
2. "When it is desired further to amplify the output" of such an arrangement "by means of additional stages, as, for example, up to a power level suitable for driving a sizable motor, this may be done by means of simple or neutral types of amplifier" such as those shown in Figs. 5.5 to 5.7.
3. "These arrangements are more effective in handling substantial amounts of power than usual polarized types of magnetic amplifier in which substantial losses usually occur in the differential or balancing circuit structure."
4. "Furthermore, most types of motors used in reversing control systems do not operate with reverse polarity input, but more usually have two different wires, one or other of which is energized for forward or reverse rotation, respectively."

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4. See "Magnetischer Verstärker" (Magnetic amplifier), Swiss patent 228,534, issued August 31, 1943 (application filed May 13, 1942), Figures 1 and 2. See also R. A. Ramey, "Separate Magnetization of Magnetic Amplifiers," United States patent 2,644,129, issued June 30, 1953 (application filed December 27, 1950).
5. G. Keinath, "Schaltung zur Messung, besonders Fernmessung, von Gleichstrom durch Beeinflussung der Permeabilität und Stromstärke eines magnetisch mit dem Gleichstromkreis gekuppelten, das Anzeigeinstrument enthaltenden Wechselstromkreises" (Circuit arrangement for the measurement, especially telemetering, of direct current by influencing the permeability and the current intensity of an a-c circuit, which is magnetically coupled with the d-c circuit and contains the measuring instrument), German patent 386,332, issued December 13, 1923 (application filed September 1, 1922), Figures 1 to 3.
6. G. Keinath, German patent 386,332, Figures 2 and 3.
7. A. S. FitzGerald, "Magnetic Amplifier," United States patent 2,027,311, issued January 7, 1936 (application filed June 21, 1933; in Great Britain, July 1, 1932), Figure 2.
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10. A. S. FitzGerald, "Magnetic Amplifier System," United States patent 2,461,046, issued February 8, 1949 (application filed May 3, 1946).
11. P. H. Dowling, "Electrical Translating Apparatus," United States patent 1,739,-579, issued December 17, 1929 (application filed June 20, 1928), Figure 4.
12. E. H. Frost-Smith, "The Theory of Magnetic Amplifiers and Some Recent Developments," *Journal of Scientific Instruments*, Vol. 25, pp. 268–272, August, 1948.
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CHAPTER 6

Basic External-feedback Types of Magnetic-amplifier Circuits

6.1. Introduction. It has already been stressed (Sec. 3.2) that actual value of the power gain

$$K_{P0} = \frac{\text{output power } P_L}{\text{input power } P_C} = \frac{I_L{}^2}{I_C{}^2}\frac{R_L}{R_C} = \frac{N_C{}^2}{R_C}\frac{R_L}{N_L{}^2} \tag{6.1}$$

of magnetic-amplifier circuits *without* feedback (about 10 to 100) depends on the design and size of the saturable-reactor elements and on proper matching of the load resistance R_L.

Reference to the transfer characteristic of Fig. 4.6 shows that it is possible to increase R_L within certain limits without changing current ratio I_L/I_C and control-circuit resistance R_C. Evidently, within these limits, the power gain $K_{P0} = P_L/P_C$ is directly proportional to the load resistance expressed in terms of $R_L/N_L{}^2$. It is to be noted, however, that the procedure of increasing the power gain by increasing the load resistance can be applied only when the maximum output power demanded is considerably lower than the possible value obtainable from the actual size of the saturable-reactor elements.

It is possible to obtain much higher power gains in magnetic amplifiers, without such a limitation, by providing regenerative-feedback circuits. The fundamental principle of these circuits consists in combining a portion of the output power with the input power in such a way that the actual power gain K_{PF} *with* feedback is A^2 times larger than the original power gain K_{P0} *without* feedback:

$$K_{PF} = K_{P0}A^2 = \frac{I_L{}^2}{I_C{}^2}\frac{R_L}{R_C}A^2 = \frac{N_C{}^2}{R_C}\frac{R_L}{N_L{}^2}A^2 \tag{6.2}$$

In this case, actual value of power gain, as defined by Eq. (6.2), can be considerably increased without changing the optimum value of load resistance which gives the maximum output power obtainable from the given size of the saturable-reactor elements.

6.2. General Comments on Feedback in Magnetic Amplifiers. In 1915, E. F. W. Alexanderson[1] gave an early illustration of the application of regenerative feedback in magnetic amplifiers: Additional d-c windings carry the alternating output current after rectification. These windings are therefore sometimes referred to as external-feedback windings.

For preliminary considerations, the rectified load current I_L (the unidirectional feedback current I_F) may be considered as an additional direct current, whose d-c control action is identical to that of the direct current I_C flowing through the control windings N_C. Corresponding to the use of feedback in electronic amplifiers, this additional d-c control action of feedback current I_F may be either aiding or opposing to the d-c control action of the input current I_C. If aiding control conditions are used, the feedback is "positive," or "regenerative." However, if opposing control conditions are used, the feedback is "negative," or "degenerative."

It is important to note that there are several essential differences between the fundamental mode of operation of the basic external-feedback types of magnetic-amplifier circuits and that of the conventional feedback arrangements, as used in electronic amplifiers. For this reason, numerous authors[2] prefer the use of the term "self-excitation" for the afore-mentioned additional windings carrying the rectified output current, leaving the term "feedback" to cover (1) the additional windings used for "true positive feedback" (*e.g.*, by means of "differential-feedback windings"[3,4]), and (2) special windings for stabilizing purposes (derivative-feedback windings), over-all and servo-type feedback effects, slope compensation, etc.

Many interesting discussions on this matter have proved that it is nearly as hard for engineers in the magnetic-amplifier art to give a good answer to the question, "Feedback or self-excitation?" as it is for them to agree on definitions of the terms "saturable reactor," "transductor," and "magnetic amplifier." The author, when outlining this book, had to come to a clear decision concerning this question of terminology. Without attempting to discuss the subject in detail, he wishes to offer his own comments, as follows:

1. The basic regenerative types of magnetic-amplifier circuits are characterized by the fact that the winding components producing the regeneration are *inductively* coupled with the control winding components. These circuits do not have the inherent property of influencing the actual magnitude of the input-circuit impedance, as, for example, the electronic amplifier of the cathode-follower type does.

2. Those regenerative types of magnetic-amplifier circuits producing regeneration by means of "self-saturation," *without additional windings*, have no "feedback channel" at all, and no current is "fed back" in any way.

These two arguments make it evident that the term "feedback," when interpreted *literally*, does *not* cover the afore-mentioned basic magnetic-amplifier circuits.

3. It has already been stressed (Sec. 6.1) that the fundamental principle of regenerative-feedback circuits consists in combining a portion of the output power with the input power in such a way that the actual power gain *with* regeneration is substantially larger than the original power gain *without* regeneration. This definition of "feedback" covers practically all types of *inherently regenerative* circuits, in the field of magnetic amplifiers as well as in that of electronic amplifiers.
4. The term "feedback" ("Rückkopplung"), as defined by T. Buchhold[5] in his classic work, has been in use in this country for a number of years in an established art. This two-syllable word is much shorter and easier to pronounce than the longer five-syllable word "self-excitation" and its various derivatives, which are even longer.

With regard to these two arguments, the term "feedback" will generally be used in this book; the term "self-excitation" will be quoted exceptionally, to show the equivalent expressions, as used in some parts of European literature.

6.3. The Basic External-feedback Circuits. Referring to Eq. (4.1) expressing the inherent current-transformer properties of the basic nonfeedback circuit with high-permeability core material, it is evident that the number of ampere-turns $I_C N_C$ required for the control windings is approximately equal to the ampere-turns $I_L N_L$ of the load windings. Thus, the number of control ampere-turns $I_C N_C$ must be comparatively large, because the control windings must supply sufficient magnetomotive force to equilibrate the peak a-c magnetomotive force of the load windings.

It is possible to rectify the alternating current I_L flowing through the load windings N_L and to use this rectified current to supply part of the necessary d-c ampere-turns, thereby decreasing the d-c ampere-turns which must be supplied from the control circuit. In this case, the "external-feedback windings" N_F carrying the rectified (pulsating) load current (feedback current I_F) will supply sufficient magnetomotive force to equilibrate that of the a-c load windings N_L, while the d-c control windings N_C have to supply only the magnetomotive force required to saturate the magnetic circuits of the saturable-reactor elements. Application of such external-feedback circuits gives the possibility of increasing the actual power gain to very high values (up to about 10^6 to 10^7).

a. The Series Type of External-feedback Circuit. This circuit (Fig. 6.1*a*) (P. H. Dowling, 1928)[6] contains two separate and equally rated saturable-reactor elements having series-opposing-connected a-c load

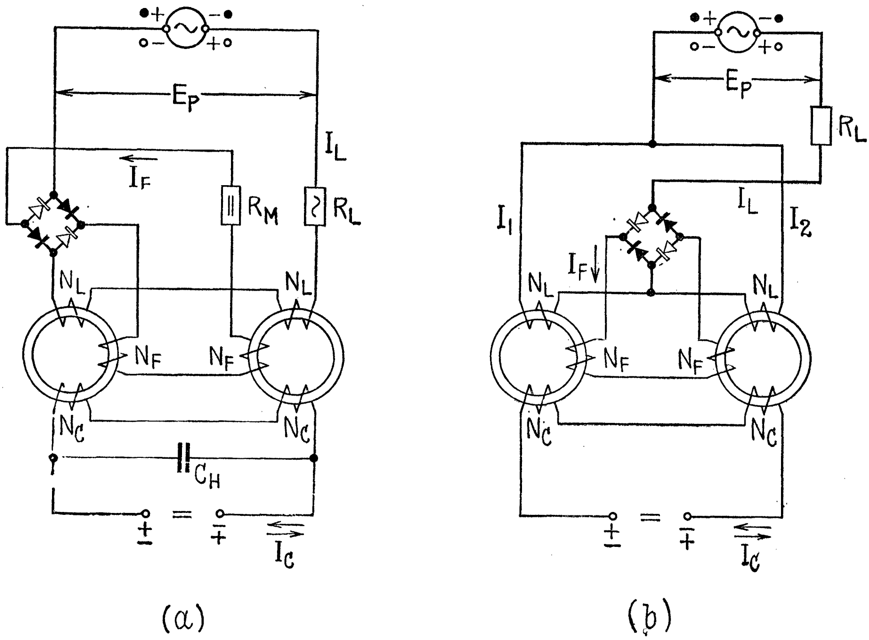

FIG. 6.1. The basic external-feedback circuits: (*a*) the circuit with series-connected a-c load windings N_L; (*b*) the circuit with parallel-connected a-c load windings N_L.

windings N_L, series-aiding-connected d-c control windings N_C, and series-aiding-connected d-c feedback windings N_F, so that the voltages induced in each of the d-c twin windings N_C and N_F by the fundamental wave and by the odd harmonics of the power-supply voltage E_P are opposed. The d-c load resistor R_M (*e.g.*, a moving-coil type of milliammeter) and the external-feedback windings N_F are supplied by a full-wave bridge-type dry-disk rectifier with direct current I_F representing the average value of the rectified load current I_L flowing through the a-c windings N_L and the a-c load resistor R_L.

A properly rated condenser C_H is connected across the control windings N_C in order to provide a very small impedance for even-harmonic-frequency currents being induced in these windings. Thus, the circuit, as shown in Fig. 6.1*a*, is working under "natural" magnetization conditions (Sec. 4.3), because the even-harmonic-frequency currents can flow freely through the condenser C_H. Consequently,

1. These even-harmonic-frequency currents are prevented from flowing through the rectifier bridge circuit representing a half-wave-rectifier device with regard to any alternating currents being induced in the feedback windings N_F.
2. Variations in total impedance of the control circuit carrying the reversible direct current I_C have no influence upon the working conditions of the two saturable-reactor elements.

The feedback windings N_F produce an additional d-c magnetization, which is proportional to the feedback current I_F representing the converted load current I_L. It is to be noted that the direction of direct current flow I_F and of the corresponding additional d-c magnetization is always the same and particularly is not dependent upon the direction of direct current flow $\pm I_C$. This is true during both half cycles.

However, the resultant d-c control action will be greatly influenced by changes in direction of the control current I_C, the additional control action of the feedback current I_F being either aiding or opposing to the control action of I_C and consequently producing either an increase or a decrease of load current I_L. Therefore, such an arrangement represents a polarized magnetic-amplifier circuit, since its operation will be influenced by changes in direction of the control current.

Considering positive (regenerative) feedback and assuming an ideal full-wave rectifier, the d-c ampere-turns assisting those provided by the control circuit will be

$$I_F N_F = I_L N_F \tag{6.3}$$

and the total d-c ampere-turns AT_{dc}, due to a control current I_C in the control windings N_C, will be

$$AT_{dc} = I_C N_C + I_L N_F \tag{6.4}$$

Substituting this for $I_C N_C$ in Eq. (4.1*a*) and neglecting the very small exciting-current component I_{L0} as before, we get

$$I_L N_L = I_C N_C + I_L N_F \tag{6.5}$$

Therefore

$$\frac{I_L}{I_C} = \frac{N_C}{N_L} \frac{1}{1 - N_F/N_L} \tag{6.6}$$

The current gain I_L/I_C has been increased in the ratio $1/(1 - N_F/N_L)$.

In the case of negative (degenerative) feedback, however, the d-c feedback ampere-turns $I_L N_F$ oppose the d-c control ampere-turns $I_C N_C$, and we get

$$\frac{I_L}{I_C} = \frac{N_C}{N_L} \frac{1}{1 + N_F/N_L} \tag{6.7}$$

In general, the effects of exciting current I_{L0} on the performance characteristics are not negligible. When $I_C = 0$, the residual current in the a-c load circuit produces direct current in the feedback windings N_F; this results in the load current being considerably greater than in the nonfeedback amplifier circuit ($N_F = 0$) when $I_C = 0$.

b. The Parallel Type of External-feedback Circuit. Alternatively, the a-c windings of the two saturable-reactor elements may be connected in parallel, as shown in Fig. 6.1*b*. The voltage across each element is then the power-supply voltage E_P, and the number of turns N_L has to be doubled. Here, the ratio between load current I_L and control current I_C is given by the following equations:

No-feedback conditions ($N_F = 0$):

$$\frac{I_L}{I_C} = 2\frac{N_C}{N_L} \tag{6.8}$$

Positive (regenerative) feedback conditions:

$$\frac{I_L}{I_C} = 2\frac{N_C}{N_L} \frac{1}{1 - 2N_F/N_L} \tag{6.9}$$

Negative (degenerative) feedback conditions:

$$\frac{I_L}{I_C} = 2\frac{N_C}{N_L} \frac{1}{1 + 2N_F/N_L} \tag{6.10}$$

Therefore, the current ratio I_L/I_C will be either increased in the ratio $1/(1 - 2N_F/N_L)$ or decreased in the ratio $1/(1 + 2N_F/N_L)$. In this case, too, the effects of exciting current I_{L0} on the performance characteristics are not negligible. With $I_C = 0$, the residual current in the parallel-connected a-c load windings produces a corresponding direct current in the feedback windings N_F; this results in the load current being considerably greater than in the nonfeedback circuit ($N_F = 0$) when $I_C = 0$.

Referring to the effects of even-harmonic-frequency currents, it is to be noted that, in the external-feedback circuit with parallel-connected a-c windings N_L, these currents will circulate in the closed-loop circuit of these windings, and not more than a very small fraction will appear in the control circuit. For this reason, providing a condenser across the control windings is not necessary.

For the transient condition there is an important difference in the operation of the two external-feedback circuits with *series*-connected (Fig. 6.1*a*) or with *parallel*-connected (Fig. 6.1*b*) a-c windings: When a d-c voltage is suddenly applied to the control windings N_C, the current in N_C and also the d-c flux increase slowly. The actual delay in building up of the steady flux is greater for the parallel-type circuit (Fig. 6.1*b*), because a transient circulating current is induced to flow in the closed-loop circuit of N_L; this current provides the opposite magnetization to the current in N_C. This prevents the rapid change of flux. For this reason, the series-type circuit (Fig. 6.1*a*) is preferable for magnetic amplifiers because it has a lower time lag.

6.4. Definition of Elementary Operation of the Basic External-feedback Circuit with Series-connected A-C Windings and High-permeability Core Material. The simple arrangement of Fig. 6.1*a* will serve as a demonstration of the elementary operation of the basic external-feedback circuit with series-connected a-c windings and high-permeability core material, working under natural magnetization conditions. It is assumed that:

1. A variable pure direct current $\pm I_C$ is applied to the control windings N_C of this circuit to adjust the magnitude of the alternating load current I_L flowing through an ohmic load R_L.
2. The full-wave, bridge-type dry-disk rectifier supplying the feedback windings N_F has ideal performance characteristics so that actual feedback current I_F represents the average value of the rectified load current I_L flowing through the series-connected a-c windings N_L.

Evidently, actual magnitude of the positive or negative feedback effect is a function of the turns ratio N_F/N_L, and $F_T = 100N_F/N_L$ represents the theoretical "feedback factor" expressed in per cent.[7] The ampere-turn ratio

$$A = \frac{I_L N_L}{I_C N_C} = \frac{1}{1 \pm N_F/N_L} = \frac{1}{1 \pm 0.01F_T} \tag{6.11}$$

and the ratio $(1 + N_F/N_L)/(1 - N_F/N_L)$ are quoted in Table 6.1 for different representative values of feedback factor F_T.

With nonfeedback amplifiers ($N_F = 0$) and with amplifiers utilizing a comparatively small feedback factor ($F_T = 50$ to 90 per cent), the magnetizing current I_{L0} is very small compared with I_{LC}—indeed, when using Orthonol tape cores, I_{L0} is practically 0.3 to 2 per cent of the maximum value of I_{LC}—and the equations expressing the relationship between current ratio I_L/I_C and turns ratio N_F/N_L hold to a good degree of accuracy (about ± 1 to 2 per cent). However, with higher feedback factors, the effects of I_{L0} on the characteristics of the amplifier are not negligible. When $I_C = 0$, the residual current in the a-c circuit produces direct current in the feedback windings N_F; this results in load current I_L considerably greater than in the case that $F_T = 0$ to 90 per cent when $I_C = 0$.

TABLE 6.1

F_T (per cent)	$\dfrac{1}{1 - N_F/N_L}$	$\dfrac{1}{1 + N_F/N_L}$	$\dfrac{1 + N_F/N_L}{1 - N_F/N_L}$
0	1	1.0	1
50	2	0.667	3
80	5	0.556	9
90	10	0.526	19
95	20	0.513	39
98	50	0.5050	99
99	100	0.5025	199
99.5	200	0.50125	399

Figure 6.2 shows schematically the resultant $I_LN_L = f(I_CN_C)$ characteristic illustrating working conditions with larger feedback factors. If $I_C = 0$, then the load current I_L has a certain value I_Q, which is sometimes called the "quiescent-current value,"[8] or "Q current," because it

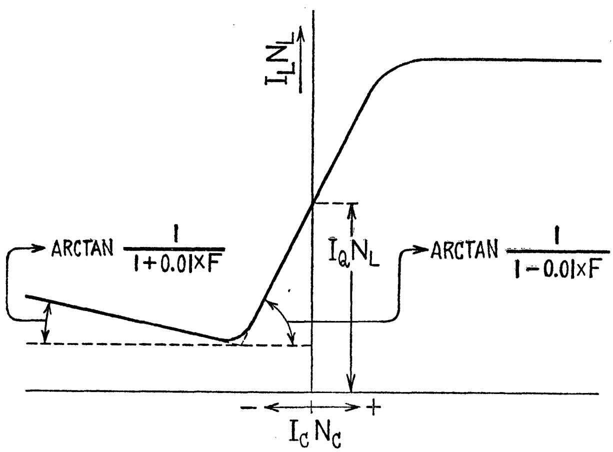

FIG. 6.2. Input-output characteristic of the basic external-feedback circuit with series-connected a-c load windings.

corresponds to the "quiescent point Q" of a three-element vacuum tube. In practice, the actual value of I_Q is dependent upon the size of the cores, magnetic properties of the core material, magnitude and frequency of the applied power-supply voltage E_P, turns ratio N_F/N_L, and departure of the feedback rectifier from the ideal characteristic.

Moreover, it is possible to reduce I_Q by means of a bias circuit producing an additional d-c magnetization which is opposed to the d-c magnetization produced by the feedback windings N_F and independent of the d-c control magnetization.

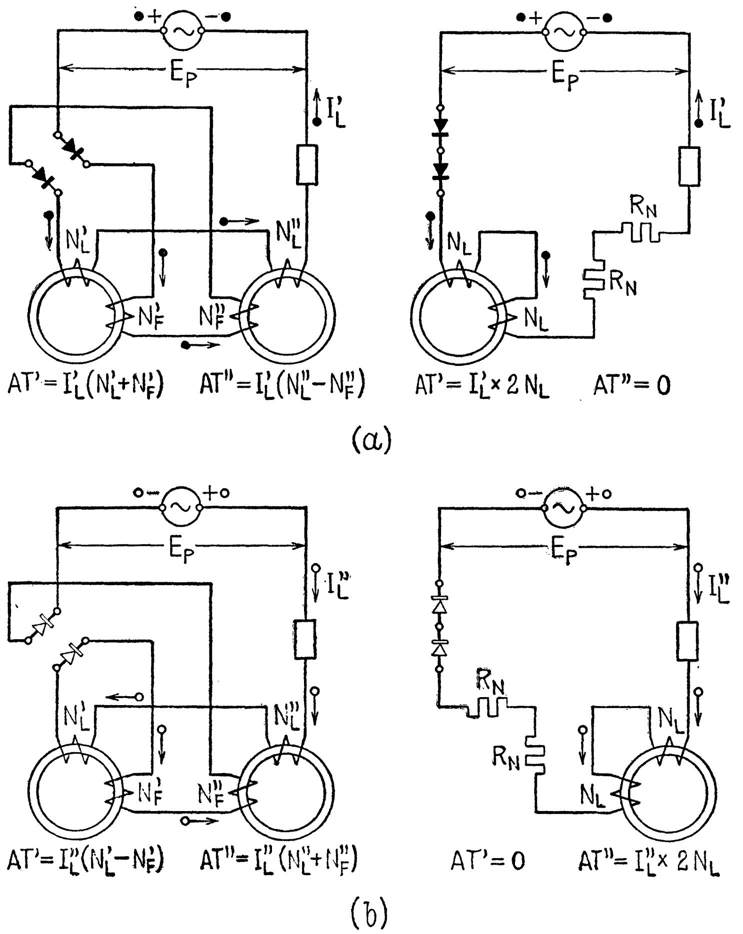

FIG. 6.3. The fundamental principle of the basic external-feedback circuit with series-connected a-c load windings N'_L and N''_L. (*a*) Operating conditions during the first half-cycle period. (*b*) Operating conditions during the second half-cycle period.

Figure 6.3 illustrating the fundamental principle of the basic external-feedback circuit with series-connected a-c windings shows that:

a. During the *first* half-cycle period the ampere-turns of the windings N'_L and N'_F are *additive*, while those of the windings N''_L and N''_F are *subtractive*.

b. During the *second* half-cycle period the ampere-turns of the windings N'_L and N'_F are *subtractive*, while those of the windings N''_L and N''_F are *additive*.

Thus, if the number of turns of the feedback windings is *exactly equal* to that of the a-c windings; *i.e.*, if

$$N'_F = N'_L \quad \text{and} \quad N''_F = N''_L \tag{6.12}$$

then

During the *first* half-cycle period:

$$AT' = I'_L(N'_L + N'_F) = I'_L \times 2N_L \tag{6.13a}$$
$$AT'' = I'_L(N''_L - N''_F) = 0 \tag{6.13b}$$

During the *second* half-cycle period:

$$AT' = I''_L(N'_L - N'_F) = 0 \tag{6.14a}$$
$$AT'' = I''_L(N''_L + N''_F) = I''_L \times 2N_L \tag{6.14b}$$

Since the ampere-turns are exactly equal in magnitude, they cancel and thus "compensate" each other exactly. This condition defines "compensated feedback" or "full compensation" (T. Buchhold[9]).

If the circuit of Fig. 6.1*a*, instead of being fully compensated, is designed so that the number of turns of the feedback windings is *smaller* than that of the a-c windings, *i.e.*, if

$$N'_F < N'_L \quad \text{and} \quad N''_F < N''_L \tag{6.15}$$

then $N_F/N_L < 1$, and the magnitude of the actual feedback effect will be correspondingly reduced. This condition defines "undercompensated feedback."[10]

However, if the circuit of Fig. 6.1*a* is designed so that the number of turns of the feedback windings substantially *exceeds* that of the a-c windings, *i.e.*, if

$$N'_F > N'_L \quad \text{and} \quad N''_F > N''_L \tag{6.16}$$

then $N_F/N_L > 1$, and the circuit is not stable ("trigger action").[11] This condition defines "overcompensated feedback."[12]

Referring to compensated-feedback conditions, it is interesting to note that always one core element operates with $2I_LN_L$ ampere-turns, while the other represents an ohmic resistance $2R_N$; and during the first and second half-cycle periods the two core elements act alternately in this way.

6.5. Actual Performance Characteristics of the Basic External-feedback Circuit with Series-connected A-C Windings and Rectangular-hysteresis-loop Core Material. In 1950, the author[7] presented results of various measurements on magnetic amplifiers having Orthonol tape cores with multilayer toroidal windings. He showed typical performance characteristics of the basic external-feedback circuit with series-connected a-c windings, operating with a power-supply frequency of 400 cps under natural magnetization conditions. These measurements were made using either indicating-pointer instruments or a

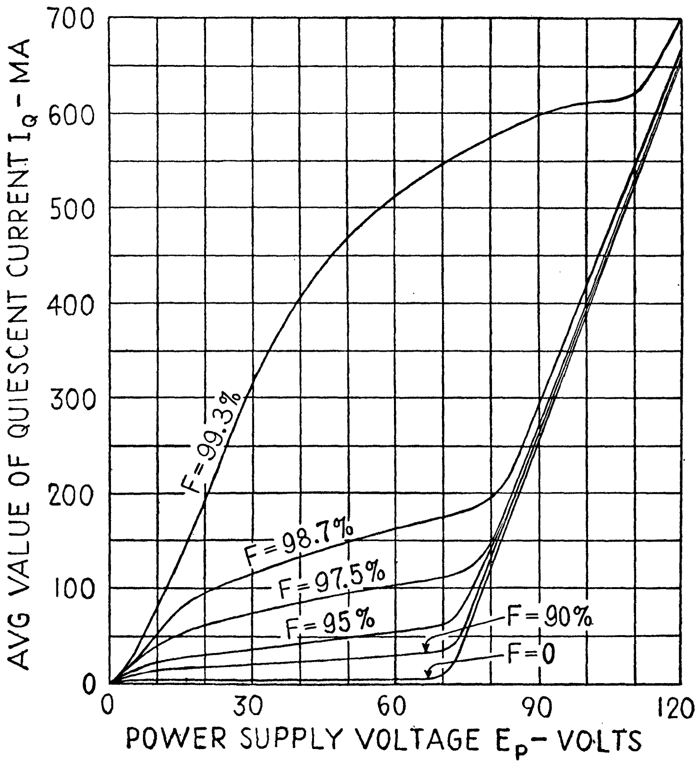

FIG. 6.4. Quiescent current I_Q as a function of power-supply voltage E_P with actual feedback factor F as a parameter ($f_P = 400$ cps).

special d-c potentiometer device making it possible to measure the actual ratio I_L/I_C with an absolute accuracy of ± 0.02 to 0.03 per cent. The resulting empirical relationships based on experimental evidence have proved to be of considerable importance for successful design of high-performance magnetic amplifiers, especially those of the push-pull balance-detector type.

a. Characteristics for Zero-control-current Conditions. Figure 6.4 shows quiescent current I_Q as a function of power-supply voltage E_P with feedback factor F as a parameter. The higher the feedback factor, the larger is the percentage change of quiescent current corresponding to a certain percentage change of E_P.

b. Characteristics for Various Values of Control Current. Figure 6.5 presents results of measurements on a small Orthonol-tape-core amplifier ($N_L = 400$, $N_C = 1{,}000$) with and without feedback. Evidently, with nonfeedback conditions, there is a linear relationship between

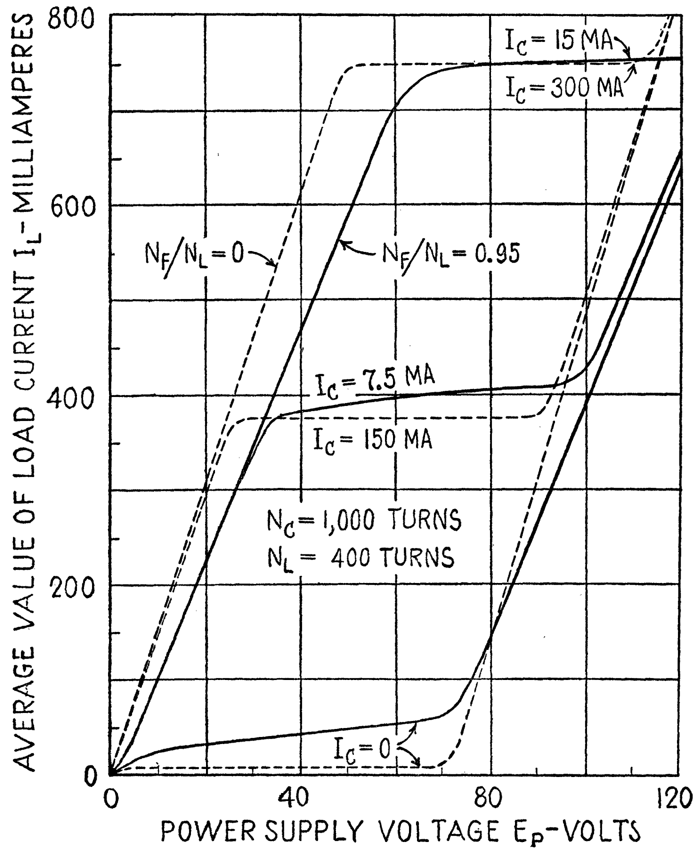

FIG. 6.5. Load current I_L as a function of power-supply voltage E_P with control current I_C as a parameter ($f_P = 400$ cps). Dashed curves: without feedback ($N_F/N_L = 0$). Full curves: with positive feedback, using the theoretical feedback factor $F_T = 95$ per cent ($N_F = 380$, $N_F/N_L = 0.95$).

output ampere-turns I_LN_L and input ampere-turns I_CN_C. Over certain ranges large changes in power-supply voltage E_P have no appreciable effect upon the current ratio I_L/I_C, which depends only on the ratio of d-c to a-c turns. These very interesting facts have already been stressed in Sec. 4.5, Figs. 4.3 to 4.8, presenting actual performance characteristics of nonfeedback circuits with Mumetal or Orthonol 2-mil tape-core elements corresponding to the power-supply frequency $f_P = 60$ cps.

The results of similar measurements with $f_P = 400$ cps and turns ratio $N_F/N_L = 0.95$ (Fig. 6.5) show that the slope of the curves $I_L = f(E_P)$ is considerably increased. Furthermore, with $I_L = 750$ ma, the actual value of control current $I_C = 15$ ma with $N_F/N_L = 0.95$ (full curves) is one-twentieth of the control current $I_C = 300$ ma with $N_F/N_L = 0$ (dashed curves). Thus, the current gain I_L/I_C has been increased twenty times.

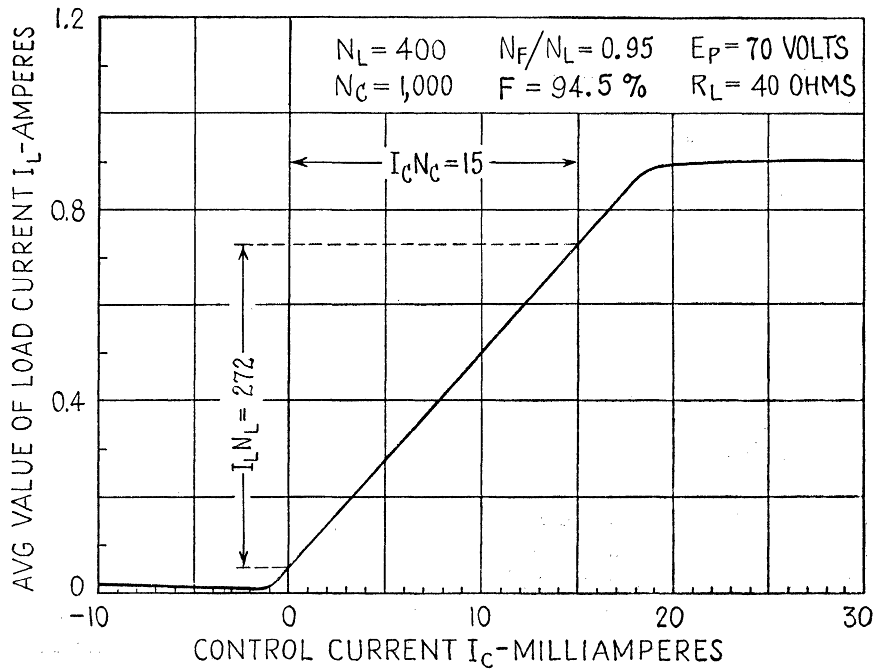

FIG. 6.6. Load current I_L as a function of control current I_C with $N_F/N_L = 0.95$ ($f_P = 400$ cps).

Figure 6.6, showing the corresponding transfer characteristic $I_L = f(I_C)$ with $N_F/N_L = 0.95$, $R_L = 40$ ohms, and $E_P = 70$ volts ($f_P = 400$ cps), makes it evident that the relationship between $I_L N_L$ and $I_C N_C$ is fairly linear. However, feedback-rectifier imperfections become evident, and the *actual* feedback factor F_A derived from Eq. (6.11):

$$F_A = \pm 100\left(1 - \frac{I_C N_C}{I_L N_L}\right) = \pm 100\left(1 - \frac{1}{A}\right) \qquad (6.17a)$$

is

$$F_A = \pm 100(1 - \tfrac{15}{272}) = \pm 100(1 - 0.055) = 94.5 \qquad (6.17b)$$

while the *theoretical* value of the applied feedback factor is

$$F_T = \frac{100N_F}{N_L} = 95.0 \text{ per cent}$$

The higher the applied feedback factor, the larger is the difference between theoretical and actual values of F, because of departure of the feedback rectifiers from the ideal characteristic.

c. Measurement of Current Ratio I_L/I_C Using the D-C Potentiometer Method.[7] Referring to Figs. 6.5 and 6.6 it will be noted that measurement of current ratio I_L/I_C by means of directly indicating ammeters of the moving-coil type, even when the best-grade instruments are used, involves practical difficulties because the actual changes in pointer deflection of such instruments, only about 0.1 to 2 per cent, are very small. In order to avoid these difficulties, a special d-c potentiometer circuit has been devised which is based upon the principle of well-known potentiometer methods generally applied for testing of instrument transformers. Using this potentiometer method with high-precision standard resistors (accuracy about ± 0.01 per cent), it is possible to measure current ratio I_L/I_C and ampere-turn ratio I_LN_L/I_CN_C with an absolute accuracy of ± 0.02 to 0.03 per cent.

Figure 6.7 illustrates the fundamental principle of the d-c potentiometer device combined with the basic external-feedback circuit, as shown in Fig. 6.1*a*. In this example, two separate full-wave bridge-type rectifiers are provided to make the actual resistance across the d-c terminals of the feedback rectifier as small as possible. This rectifier feeds only the feedback windings N_F, whereas the "load rectifier" is provided for supplying the load resistor R_L with the direct load current I_L. It has been found that improved performance characteristics may be obtained in this way. The voltage drop I_LR_1 across a fixed standard resistor R_1 will be balanced by a variable voltage drop produced by the control current I_C on a special potentiometer-resistor network containing standard resistors (resistance boxes) R_2, R_3, R_4, and R_5 and a slide-wire potentiometer R_6 having linear characteristic. If these voltage drops are not equal, a current I_G flows through the moving-coil-type galvanometer G operating as a balance detector. However, with full balance conditions, $I_G = 0$, and the actual resistance $\pm r$ on the slide-wire potentiometer, which is exactly proportional to the percentage change of current ratio I_L/I_C, gives directly the values to be measured as functions of control current I_C, power-supply voltage E_P, and frequency f_P.

The potentiometer method, as illustrated in Fig. 6.7, has proved to be very practical for the development of magnetic amplifiers utilizing feedback factors in the range of $F = 0$ to 95 per cent. Furthermore, the method will be generally valuable for systematic investigations concerning testing and comparison of different magnetic core materials. Such measurements can preferably be made using a self-balanced d-c potentiometer, in order to obtain direct automatic recording of the

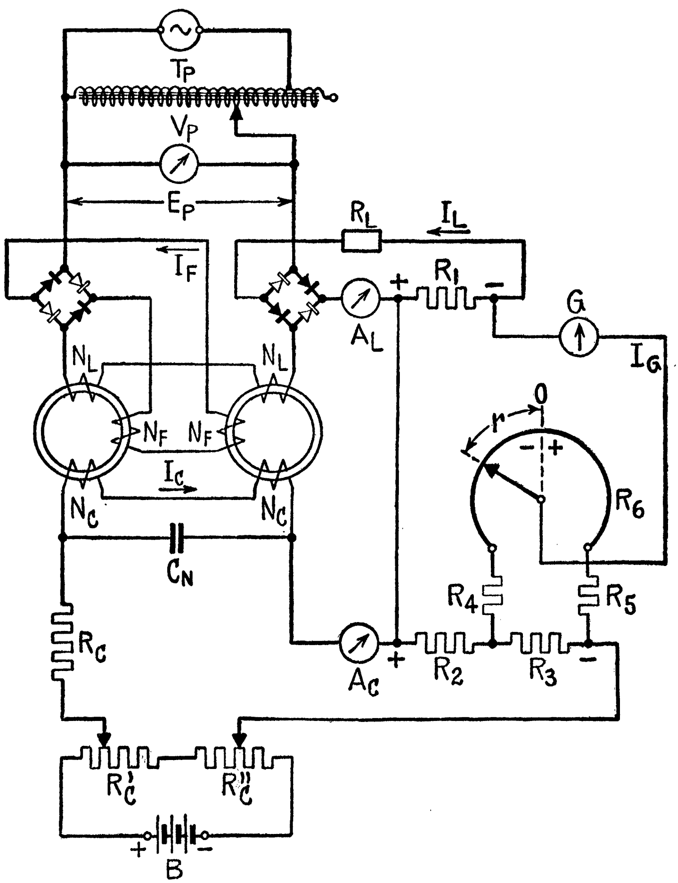

FIG. 6.7. Potentiometer device combined with a magnetic-amplifier circuit of the external-feedback type.

magnetic-amplifier characteristics. This will be especially valuable in the manufacturing process of magnetic materials of the nickel-iron-alloy type.

d. Results of Potentiometer Measurements on a Magnetic Amplifier Having Orthonol Tape Cores. The percentage change of current ratio I_L/I_C has been measured as a function of control current I_C, power-supply voltage E_P, and frequency f_P on a special magnetic amplifier that operates with 50 per cent positive (regenerative) feedback and has an output power $I_L^2R_L = 60$ watts, power gain 140 (ratio between output and input power), and rated current ratio $I_L/I_C = 10.0$ ($f_P = 200$ to 800 cps). Figures 6.8 to 6.10 show results of these measurements.

It is interesting to note that the total change of current ratio I_L/I_C does not exceed about 4 per cent, or ± 2 per cent, although the actual

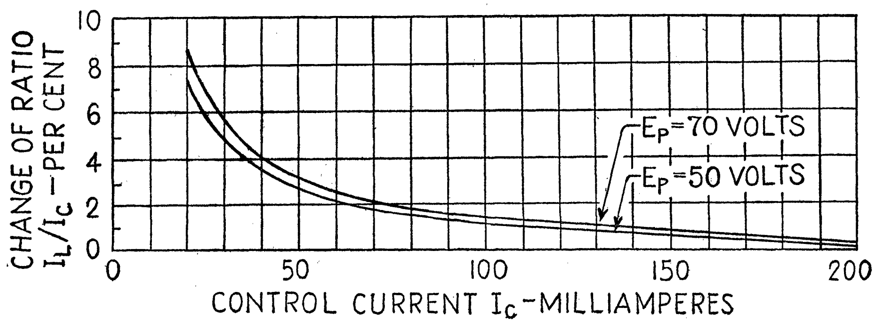

FIG. 6.8. Percentage change of current ratio I_L/I_C as a function of control current I_C with power-supply voltage E_P as a parameter ($f_P = 400$ cps).

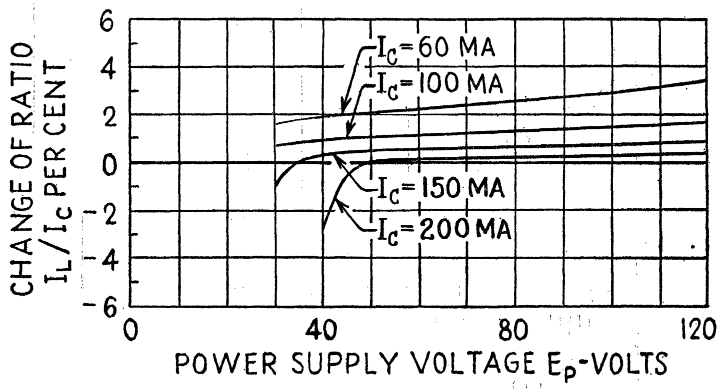

FIG. 6.9. Percentage change of current ratio I_L/I_C as a function of power-supply voltage E_P with control current I_C as a parameter ($f_P = 400$ cps).

variations in control current, supply voltage, and frequency are very large. The results of these measurements lead to the conclusion that the actual behavior of the magnetic-amplifier circuit investigated having Orthonol tape cores approximates very closely to the corresponding characteristics of an ideal amplifier based upon the assumption of an idealized magnetization curve which consists of three straight lines, as shown in Fig. 2.1*b*. As a result of this very close approximation between actual experimental conditions and usual theoretical assumptions, performance characteristics of such high-performance magnetic amplifiers can be represented by simple linear equations with sufficient accuracy to be used in the design of these amplifiers, especially those of the push-pull balance-detector type.

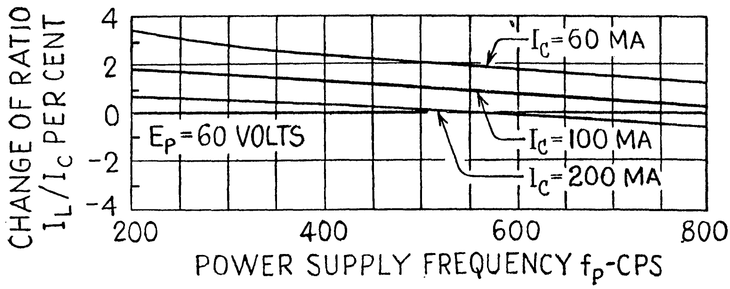

FIG. 6.10. Percentage change of current ratio I_L/I_C as a function of power-supply frequency f_P with control current I_C as a parameter (E_P = 60 volts).

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CHAPTER 7

Development of External-feedback Types of Magnetic-amplifier Circuits

7.1. Introduction. The basic external-feedback circuits, as described in the preceding chapter, represent d-c-controlled "polarized" magnetic amplifiers which are able to discriminate between "positive" and "negative" control currents. Referring to the asymmetrical transfer characteristic of Fig. 6.2, it is evident that a positive control current $+I_C$ causes an increase of load current I_L, while a negative control current $-I_C$ causes a decrease of I_L. However, I_L will never change its direction.

In order to obtain external-feedback types of magnetic amplifiers having duodirectional transfer characteristics, as shown in Fig. 5.1*c*, it is necessary to apply certain modifications of the basic circuit with series-connected a-c load windings (Fig. 6.1*a*).

Furthermore, when using duodirectional magnetic amplifiers in connection with high-performance servomechanisms, it will be advisable in many cases to apply the a-c error voltage to the control circuit of the amplifier without incorporation of a phase-sensitive rectifier device.

Another requirement, which is particularly important for the design of magnetic servo amplifiers, consists in the possibility of controlling the actual magnitude of the quiescent current (I_Q in Fig. 6.2) to obtain the most favorable operating conditions.

Finally, a multistage design of a magnetic servo amplifier will be preferred in many cases to obtain substantially improved performance characteristics with regard to power gain and speed of response.

This chapter describes various possibilities for successful development of external-feedback types of magnetic-amplifier circuits.

7.2. Auto-connected External-feedback Circuits. Figure 7.1 shows an "auto-connected, self-excited parallel transductor," as described by A. U. Lamm[1] and U. H. Krabbe.[2] This bridge-type external-feedback circuit, particularly suitable for installations with large power outputs, makes it possible to save a considerable amount of material and to avoid leakage reactance between the load and feedback windings. However, a current transformer T_F must be provided in this circuit to achieve the necessary insulation between a-c circuit and feedback rectifier.

In this arrangement, feedback current I_F proportional to load current I_L is led to terminals in the windings N_1, N_2, N_3, N_4, which are equipotential with respect to the terminals in the a-c circuit. The actual magnetizations produced by I_L and I_F are properly superimposed so that each saturable-reactor element behaves in the same way as it would with separate windings N_L and N_F (Fig. 6.1*b*). Figure 7.1 illustrates that an a-c load R'_L or a d-c load R''_L may be used in this arrangement.

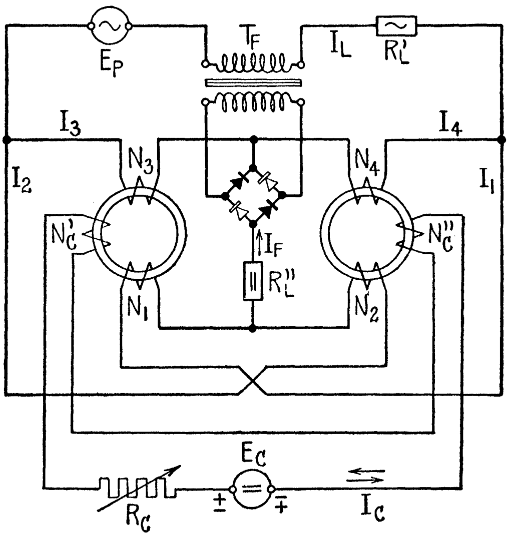

FIG. 7.1. Auto-connected external-feedback circuit. (*A. U. Lamm.*[1])

The basic principle of the auto-connected single-phase external-feedback circuit can be extended to three-phase arrangements. The "Nordfeldt connection"[3] with intermediate three-phase transformer makes it possible to lead the output direct current directly to three separate saturable-reactor units in such a way that the desired feedback effect ("self-excitation") is obtained.

7.3. A-C Controlled External-feedback Circuits. In 1940, G. Barth[4] showed that external-feedback-type magnetic amplifiers may be controlled by means of a variable alternating current having the same frequency as the power-supply generator. This can be achieved in various ways.

Figure 7.2*a* illustrates the possibility of a-c control without providing separate control windings on the saturable-reactor elements. In this case, the alternating control voltage E_C, synchronous with power-supply voltage E_P, is applied to the a-c terminals of the feedback rectifier through a sufficiently high impedance, *e.g.*, a variable series resistor R_C. Thus, the alternating load current I_L and the synchronous variable control current I_C are superimposed, so that the actual value of d-c feedback ampere-turns $I_F N'_F = I_F N''_F$ is a function of E_C.

It is to be noted that this circuit acts as a phase discriminator, like a phase-sensitive rectifier. Actual magnitudes of current I_L (a-c load R'_L) and I_F (d-c load R''_L) will correspond to the actual phase displacement between control voltage E_C and power-supply voltage E_P. In practice, the phase relationship between E_P and E_C will be adjusted so that the

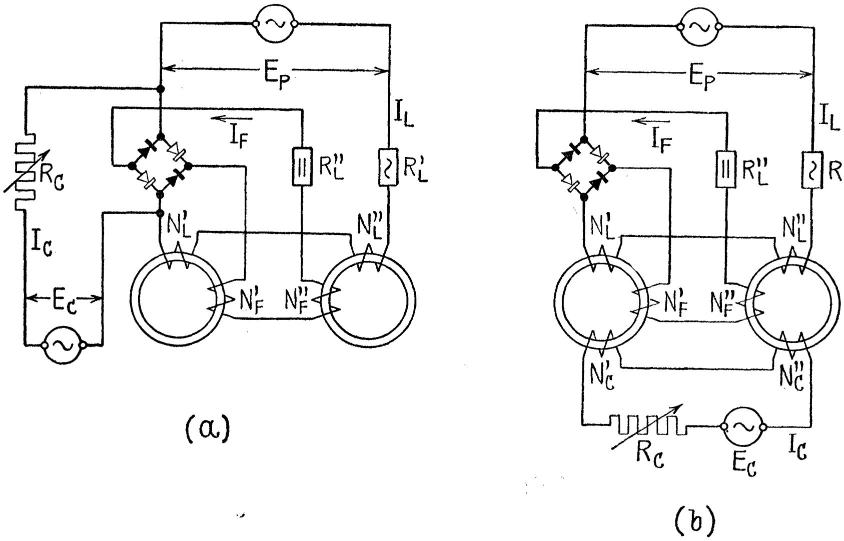

FIG. 7.2. A-c-controlled external-feedback circuits (*G. Barth*[4]): (*a*) a-c control without providing separate control windings on the saturable-reactor elements; (*b*) a-c control by means of series-opposing-connected control winding elements N'_C and N''_C.

actual displacement is nearly 0 or 180° to obtain most favorable operating conditions. However, in some cases, the circuit may be controlled, preferably with a substantially constant control current I_C, by changing the actual phase displacement between E_P and E_C.

Figure 7.2*b* illustrates the mode of operation of a modified external-feedback circuit which is a-c controlled by means of series-opposing-connected control winding elements N'_C and N''_C. Comparison of Figs. 6.1*a* and 7.2*b* shows that the terminals of one control winding element (N''_C) have been reversed to obtain a-c control instead of d-c control. It is to be noted, however, that the voltages induced in each of the elements N'_C and N''_C by the fundamental wave and by the odd harmonics of power-supply voltage E_P are *not* opposed in this case, and that corresponding alternating currents will circulate in the control circuit. Therefore, it will be necessary to connect a sufficiently high resistance R_C in series with N'_C and N''_C so that the voltages induced from the load windings N'_L, N''_L into the control windings N'_C, N''_C are unable to produce excessively large currents in the control-circuit loop.

The external-feedback circuit of Fig. 7.2*b* may be controlled in different ways; *e.g.*:

1. Providing a variable series resistor R_C and using a constant synchronous control voltage E_C, the constant phase relationship between E_C and E_P being nearly 0 or 180°.
2. Providing another type of variable series impedance (*e.g.*, a variable reactor or a variable condenser) in the a-c control circuit and using a constant synchronous voltage E_C, the constant phase displacement between E_C and E_P being adjusted so that most favorable operating conditions will be obtained.
3. Providing a constant series resistor R_C and using a variable synchronous control voltage E_C, the constant phase relationship between E_C and E_P being nearly 0 or 180°.
4. Providing a constant series resistor R_C and using a synchronous control voltage E_C of constant magnitude and variable phase displacement with respect to E_P.

7.4. External-feedback Circuits with D-C Bias. Referring to the typical transfer characteristic of Fig. 6.2, it is evident that with $I_C = 0$ the load current I_L has a certain value I_Q, which is called "quiescent current." It has already been stressed that actual magnitude of I_Q is dependent upon the size of the cores, magnetic properties of the core material, magnitude and frequency of the power-supply voltage E_P, turns ratio N_F/N_L, and departure of the feedback rectifier from the ideal characteristic. It has also been mentioned that it is possible to reduce I_Q by means of a bias circuit, producing an additional d-c magnetization which is opposed to the d-c magnetization produced by the feedback windings N_F and independent of the d-c control magnetization.

Figure 7.3*a* shows a d-c-controlled external-feedback circuit having special d-c bias windings N'_B, N''_B supplied by a full-wave rectifier with a-c series resistor R_S and d-c series resistor R_B. The bias current

$$I_B = \text{const} \times E_P$$

produces an additional d-c magnetization, which is opposing to the d-c magnetization produced by the feedback windings N_F and independent of the variable control current I_C.[5]

Moreover, it is possible to apply the bias current I_B directly to common d-c windings N'_C, N''_C (Fig. 7.3*b*) acting simultaneously as d-c bias windings and d-c control windings. In this case, however, the d-c series resistor R_B shunting the common windings must be so rated as to have a value high enough to avoid a considerable decrease of the control action of I_C.[6]

7.5. External-feedback Circuits with A-C Bias. Application of a-c bias on external-feedback circuits (G. Barth, 1942[7]) makes it possible to eliminate additional bias rectifiers, as used in the circuits of Fig. 7.3 with d-c bias.

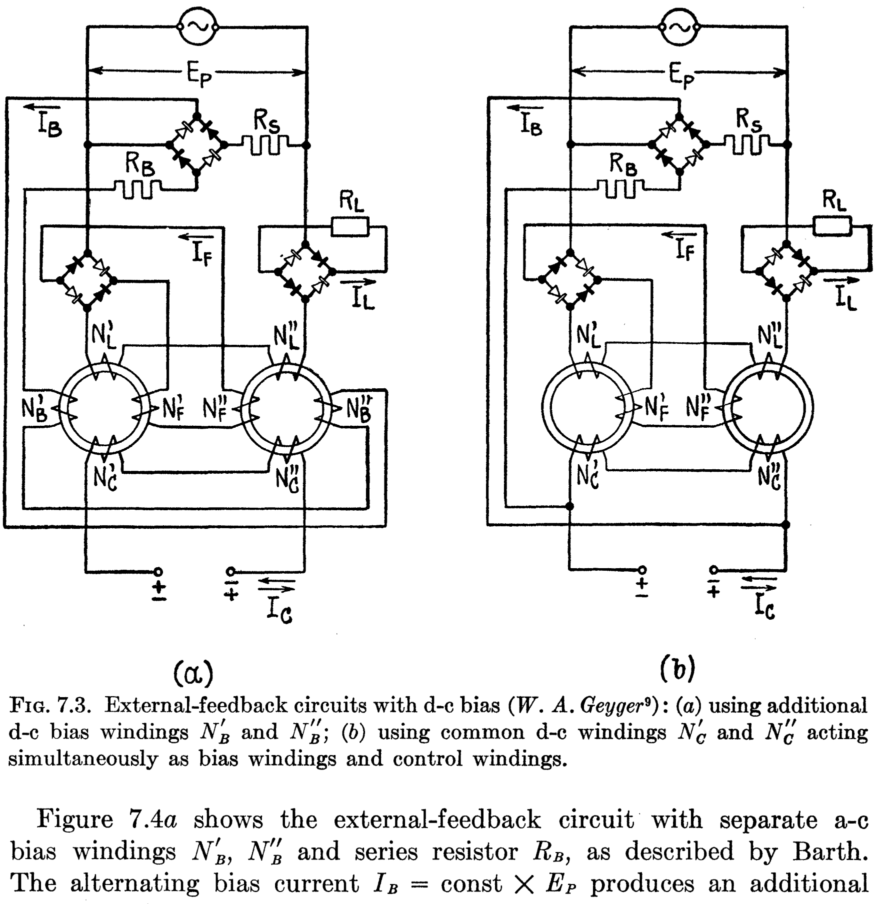

FIG. 7.3. External-feedback circuits with d-c bias (*W. A. Geyger*[9]): (*a*) using additional d-c bias windings N_B' and N_B''; (*b*) using common d-c windings N_C' and N_C'' acting simultaneously as bias windings and control windings.

Figure 7.4*a* shows the external-feedback circuit with separate a-c bias windings N_B', N_B'' and series resistor R_B, as described by Barth. The alternating bias current $I_B = \text{const} \times E_P$ produces an additional a-c magnetization, which is superimposed on the a-c magnetization of the load windings N_L', N_L'' in such a way that the magnitude of quiescent current I_Q can be reduced to the desired value. Since, when rectangular-hysteresis-loop core material is used, the fundamental wave of I_Q is nearly in phase with the power-supply voltage E_P, sufficient reduction of I_Q can generally be achieved by means of a series-connected ohmic bias resistor R_B. However, with other core materials and with higher values of turns ratio N_F/N_L, it will be necessary to provide additional means for adjusting the phase displacement between I_B and E_P, preferably a properly rated reactor or condenser in the bias circuit.

Figure 7.4*b* illustrates the possibility of applying the alternating bias current I_B directly to the winding elements N'_C, N''_C of an a-c-controlled external-feedback circuit. A properly rated shunt condenser C_B is provided to obtain optimum value of phase displacement between I_B and E_P.

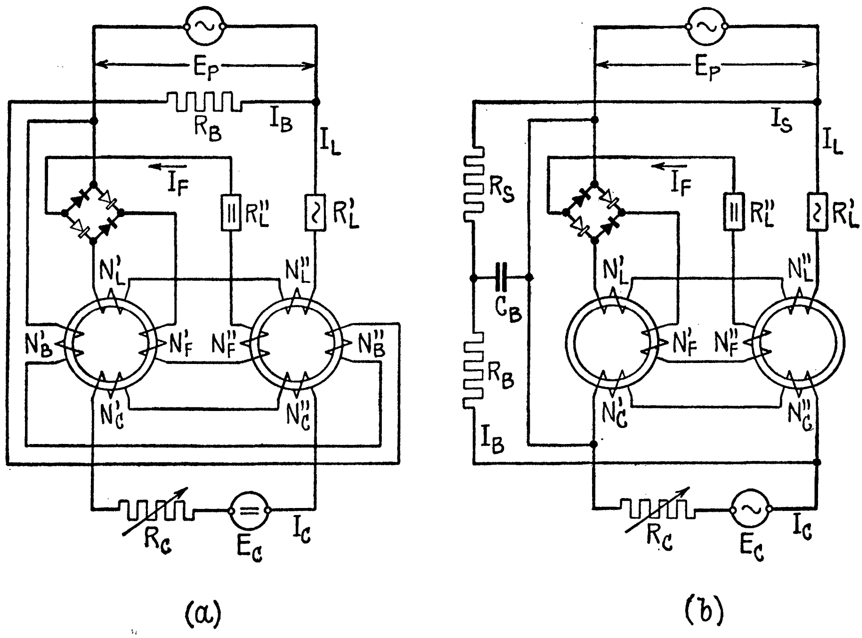

FIG. 7.4. External-feedback circuits with a-c bias (*G. Barth*[7]): (*a*) using additional a-c bias windings N'_B and N''_B; (*b*) using common a-c windings N'_C and N''_C acting simultaneously as bias windings and control windings.

Figure 7.5 shows typical results of measurements on a small magnetic amplifier with Mumetal tape cores using the circuit of Fig. 7.4*b* with $N'_F/N'_L = N''_F/N''_L = 1.0$ (compensated feedback). Average values of quiescent current I_Q have been measured as a function of the actual values of total bias resistance $R_B = R'_B + R''_B$, with $C_B = 0$ and with $C_B = 0.3\ \mu$f. These results make it evident that application of shunt condenser C_B gives improved operating conditions in this case. The quiescent current I_Q (about 100 ma without bias) can be reduced to about 1.5 ma when $C_B = 0.3\ \mu$f and $R_B = R'_B + R''_B = 2{,}200$ ohms.

7.6. Inherent Constant-current Characteristics of Voltage-sensitive Bias Circuits. In 1942, the author[8] suggested designing magnetic servo amplifiers utilizing external feedback in such a way that the magnitude of the quiescent current is substantially independent of changes in power-supply voltage. This was accomplished by applying properly rated voltage-sensitive bias circuits, which introduce inherent constant-current characteristics in magnetic amplifiers operating under compensated-feedback conditions.

Figure 7.6 shows the quiescent current I_Q as a function of power-supply voltage E_P if *constant* and *variable* biasing ampere-turns are

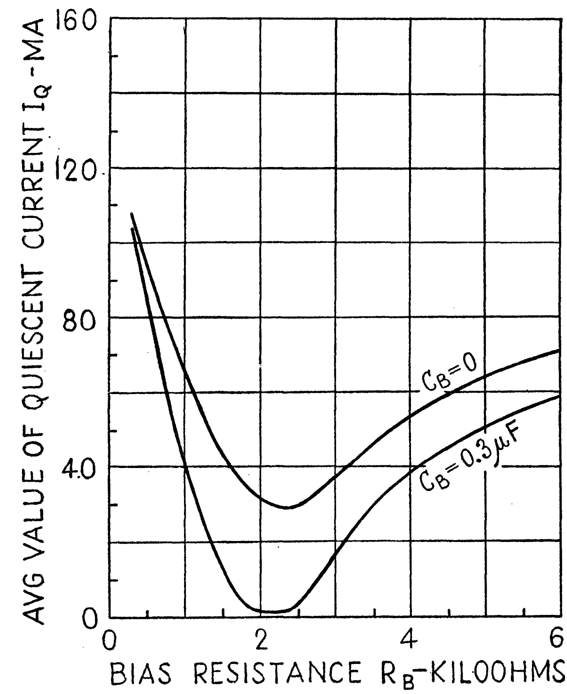

FIG. 7.5. Quiescent current I_Q as a function of a-c bias resistance $R_B = R_B' + R_B''$, without or with bias condenser C_B.

supplied to the cores. In both cases, the parameter is the actual quiescent-current value $I_Q = 50, 70$, or 100 ma corresponding to $E_P = 16$ volts.

The results of these measurements, using either constant or voltage-sensitive bias circuitry, show in a dramatic way the progress which is of paramount importance for successful development of magnetic servo amplifiers, which must have an extremely small drift rate so that the performance of the servomechanism is not appreciably influenced by changes of from ± 5 to 10 per cent in power-supply voltage.

It is interesting to note that special voltage-current characteristics, as shown in Fig. 7.6, are due to the inherent negative-feedback properties of voltage-sensitive bias circuits: When the power-supply voltage E_P

increases, then the bias current I_B increases correspondingly, so that the quiescent current I_Q will be automatically controlled and have a substantially constant magnitude. Thus, the bias windings N'_B, N''_B carrying $I_B = \text{const} \times E_P$ act as additional control windings to make I_Q practically independent of changes in voltage E_P within the limits from ± 5 to 10 per cent.

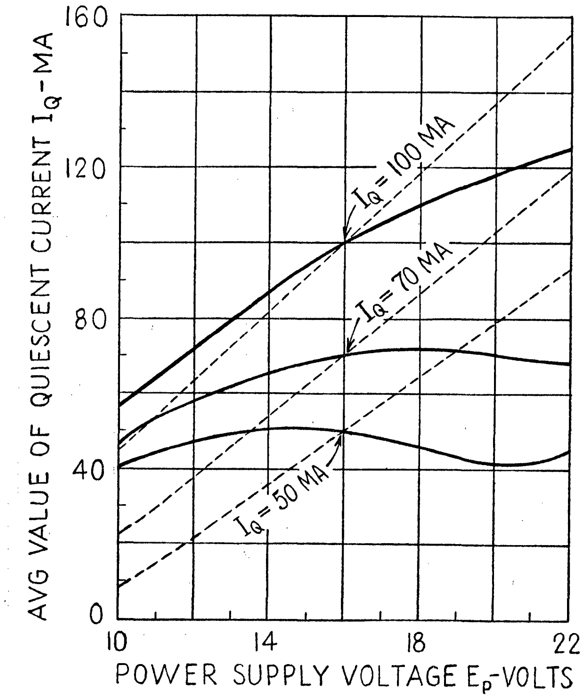

FIG. 7.6. Quiescent current I_Q as a function of power-supply voltage E_P (a) with constant biasing ampere-turns (dashed lines); (b) with variable biasing ampere-turns (full lines).

7.7. External-feedback Circuit with Mechanical Rectifier and Constant-voltage Transformer. It is important to note that the actual current gain I_L/I_C of very sensitive magnetic amplifiers with compensated feedback ($N_F = N_L$) will be substantially influenced by fluctuations of power-supply voltage and also by major changes in ambient temperature, because of the temperature-sensitive performance of the dry-disk rectifiers in the feedback circuit.

In 1940, the author[9,10] devised an external-feedback circuit (Fig. 7.7) having a mechanical type of feedback rectifier MR and a simple constant-voltage transformer T_E.* Replacing the conventional selenium

* See Sec. 18.4a, Magnetic Voltage Stabilizers.

or copper-oxide dry-disk type of feedback rectifier by a mechanical rectifier using either an oscillating or a rotating contact device has proved to be practicable in special cases, particularly for very sensitive input-stage circuits having a comparatively small output power of a few milliwatts only. The arrangement of Fig. 7.7 used a vibrating-reed

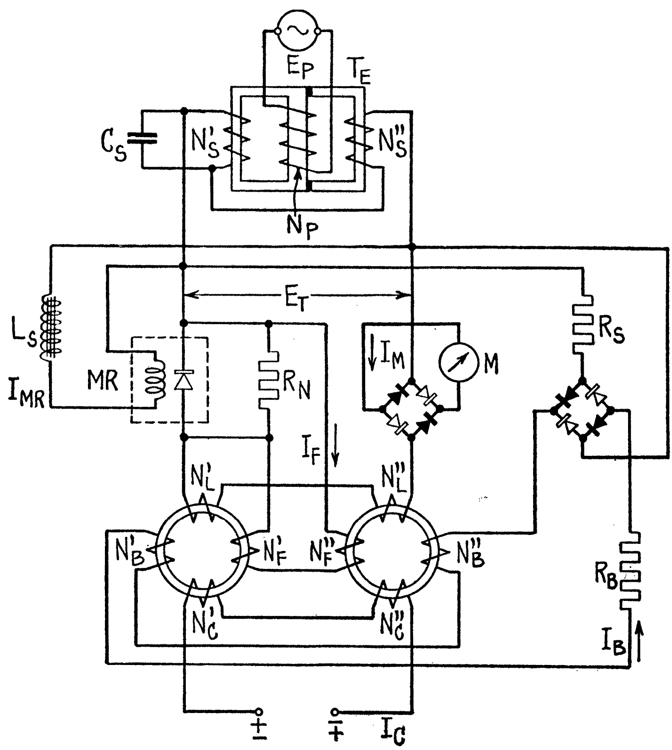

FIG. 7.7. External-feedback circuit with mechanical feedback rectifier *MR* and constant-voltage transformer T_E. (*W. A. Geyger.*[9])

type of rectifier (H. Pfannenmüller, 1932[11]) with a series reactor L_S for proper phase adjustment of the exciting current I_{MR} of the operating coil of *MR*. Of course, other types of mechanical rectifiers, *e.g.*, the "contact rectifier," as developed by F. Koppelmann,[12] may be useful for this purpose.

The arrangement of Fig. 7.7, especially designed for increasing the sensitivity of a moving-coil ink recorder, has a d-c bias circuit N'_B, N''_B, R_B, R_S with bias rectifier and an additional load rectifier supplying the ink recorder M with a direct current I_M (0 to 10 ma), which is a function of the very small direct control current I_C. The bias current I_B may be adjusted in such a way that with $I_C = 0$ the ink pen will be at the center of the scale of the recorder. Thus, reversible control currents $\pm I_C$ may be recorded with this arrangement.

7.8. The Feedback Diagram. T. Buchhold[13] used the so-called "feedback diagram" for defining the elementary operation of the basic external-feedback circuits working under compensated-, undercompensated-, or

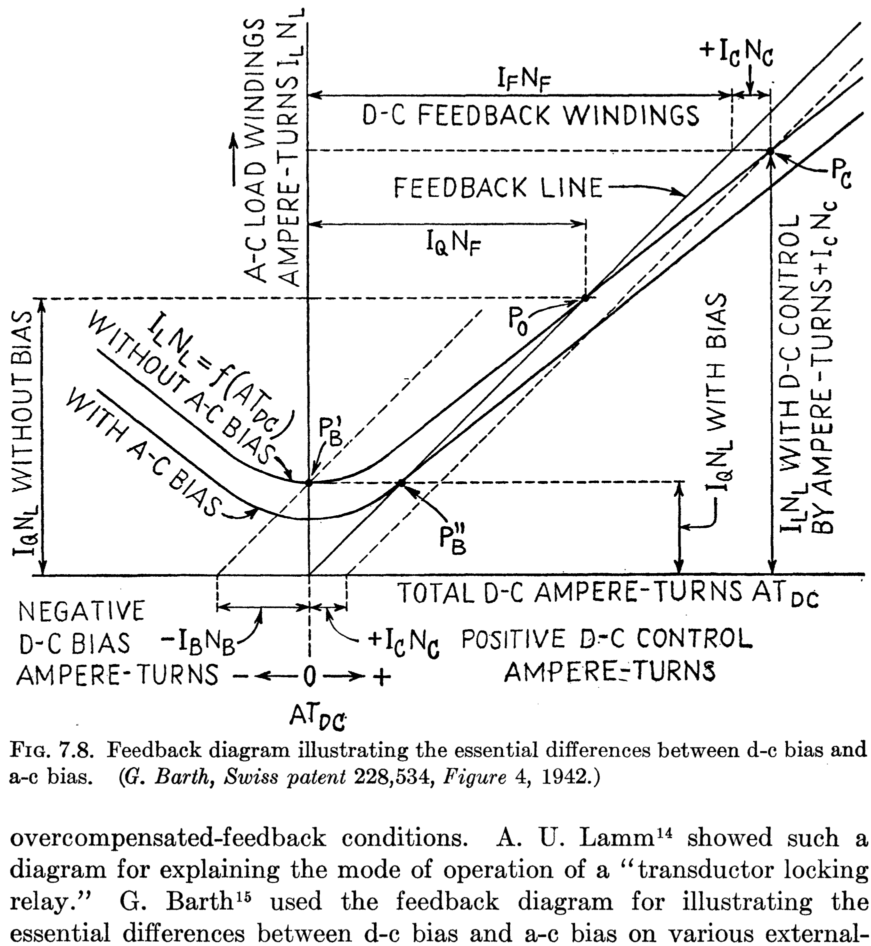

FIG. 7.8. Feedback diagram illustrating the essential differences between d-c bias and a-c bias. (*G. Barth, Swiss patent* 228,534, *Figure* 4, 1942.)

overcompensated-feedback conditions. A. U. Lamm[14] showed such a diagram for explaining the mode of operation of a "transductor locking relay." G. Barth[15] used the feedback diagram for illustrating the essential differences between d-c bias and a-c bias on various external-feedback circuits and similar arrangements.

Figure 7.8 shows the feedback diagram of an external-feedback circuit with series-connected a-c load windings having either d-c bias (Fig. 7.3) or a-c bias (Fig. 7.4). This diagram contains:

1. Two characteristics representing the ampere-turns $I_L N_L$ of the a-c load windings as a function of the total d-c ampere-turns AT_{dc}, *without* and *with* a-c bias, respectively

2. The "feedback line" representing the ampere-turns I_FN_F of the d-c feedback windings as a function of the total d-c ampere-turns AT_{dc}
3. Two dotted straight lines obtained by shifting the feedback line with an amount of positive d-c control ampere-turns $+I_CN_C$, or with an amount of negative d-c bias ampere-turns $-I_BN_B$

The characteristic "$I_LN_L = f(AT_{dc})$ without a-c bias" represents the relation between the average value of a-c load-current ampere-turns I_LN_L and the *total* d-c ampere-turns AT_{dc}, which latter are composed of the feedback ampere-turns I_FN_F, the positive control ampere-turns $+I_CN_C$, and the negative ampere-turns $-I_BN_B$ of the d-c bias windings.

Evidently, assuming ideal performance of the feedback rectifier, the feedback ampere-turns I_FN_F are at all times proportional to the a-c load-current ampere-turns I_LN_L. With compensated feedback

$$\left(\frac{N_F}{N_L} = 1\right)$$

$I_FN_F = I_LN_L$, and the slope of the feedback line is 45°, as shown in the schematic diagram of Fig. 7.8.

Supposing that no additional d-c ampere-turns are applied ($+I_CN_C = 0$ and $-I_BN_B = 0$), the actual operating point must be on the feedback line and must also be on the characteristic representing $I_LN_L = f(AT_{dc})$ without a-c bias. The circuit will consequently operate at P_0, the point of intersection of these two lines, and a quiescent current I_Q will flow corresponding to the indicated quiescent-current ampere-turns "I_QN_L without bias."

Application of the positive control ampere-turns $+I_CN_C$ will cause the operating point to rise from P_0 to P_C. It is to be noted that only a comparatively small amount $+I_CN_C$ is required, since the very large component I_FN_F is provided by the d-c feedback windings.

Application of the negative bias ampere-turns $-I_BN_B$ will cause the operating point to fall from P_0 to P'_B, the new point of intersection. Now a much smaller quiescent current I_Q will flow corresponding to the indicated quiescent-current ampere-turns "I_QN_L with bias." Evidently, in the case of d-c bias, the slope of the characteristic "$I_LN_L = f(AT_{dc})$ without a-c bias" is very low at point P'_B. This means that the magnetic amplifier has a very low sensitivity at this point and that a considerable amount of d-c control ampere-turns $+I_CN_C$ will be necessary to cause substantial change of load current I_L.

However, when a-c bias is applied, then the new operating point P''_B will be given by the intersection between the characteristic "$I_LN_L = f(AT_{dc})$ with a-c bias" and the feedback line, which, of course, remains unchanged.

In the example illustrated in Fig. 7.8, the quiescent current I_Q with a-c bias has been reduced to exactly the same value as before (when using the negative d-c bias ampere-turns $-I_B N_B$).

It is to be noted that, in the case of a-c bias, the slope of the characteristic "$I_L N_L = f(AT_{dc})$ with a-c bias," at the new operating point P''_B, is almost in the linear region of this characteristic, although I_Q has

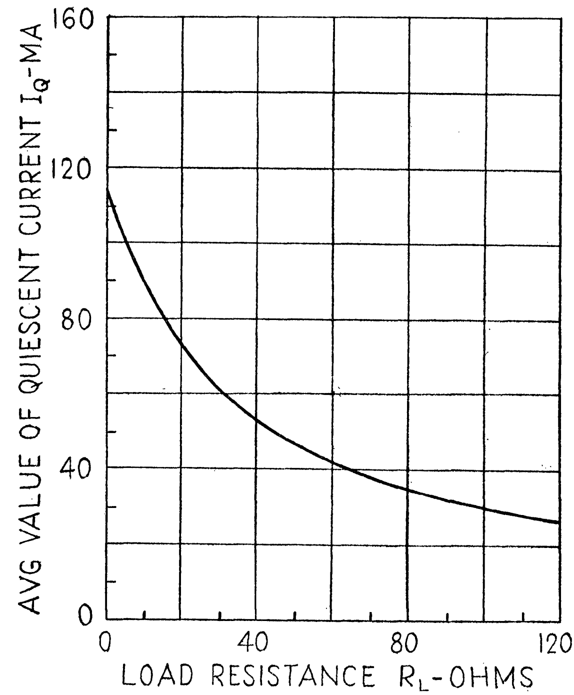

FIG. 7.9. Quiescent current I_Q as a function of load resistance R_L, without bias. $E_P = 16$ volts; $I_C = 0$.

been reduced in the same amount in both cases. Thus, the magnetic amplifier has nearly its maximum sensitivity at this point. This fact makes it evident that application of a-c bias circuitry may offer important advantages in certain cases, particularly when the quiescent current is to be reduced to extremely small values, corresponding to the actual magnitude of the excitation current I_M with nonfeedback conditions (Figs. 4.7 and 4.8).

7.9. Actual Performance Characteristics of External-feedback Magnetic-amplifier Circuits. Figures 7.9 to 7.12 give results of various measurements representing typical examples of actual performance characteristics of external-feedback circuits with Mumetal tape cores operating under compensated-feedback conditions ($N_F = N_L$).

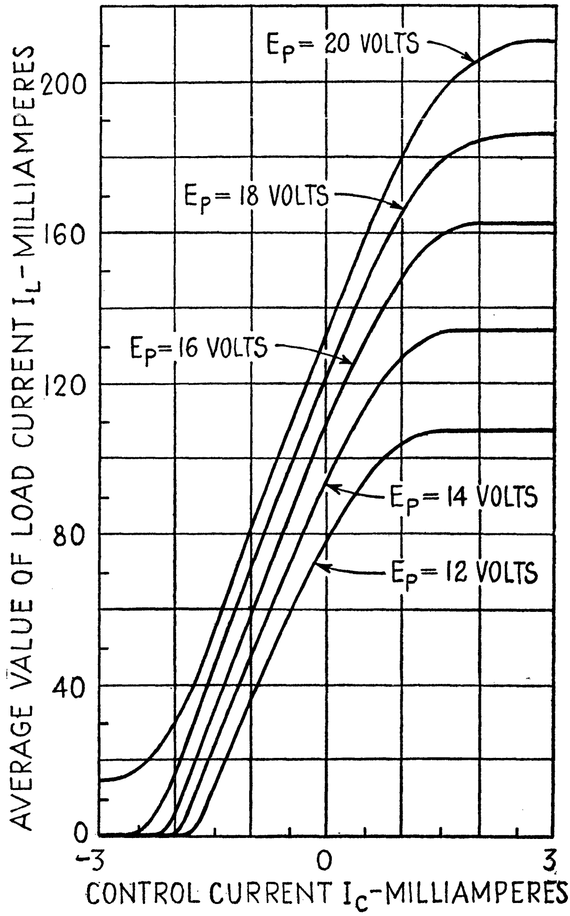

FIG. 7.10. Load current I_L as a function of control current I_C with power-supply voltage E_P as a parameter, without bias. $R_L = 3$ ohms.

The two saturable-reactor elements of the test circuit of Fig. 7.4*b* consisted of spirally wound 2-mil tape cores having the following dimensions:

Outer diameter, 2.0 in.
Inner diameter, 1⅝ in.
Tape width, 0.5 in.

The tape cores had the following multilayer toroidal windings of enamel-insulated copper wire:

1. Inner a-c load windings N'_L, N''_L of 1,000 turns each of No. 28 wire (Brown and Sharpe gauge)

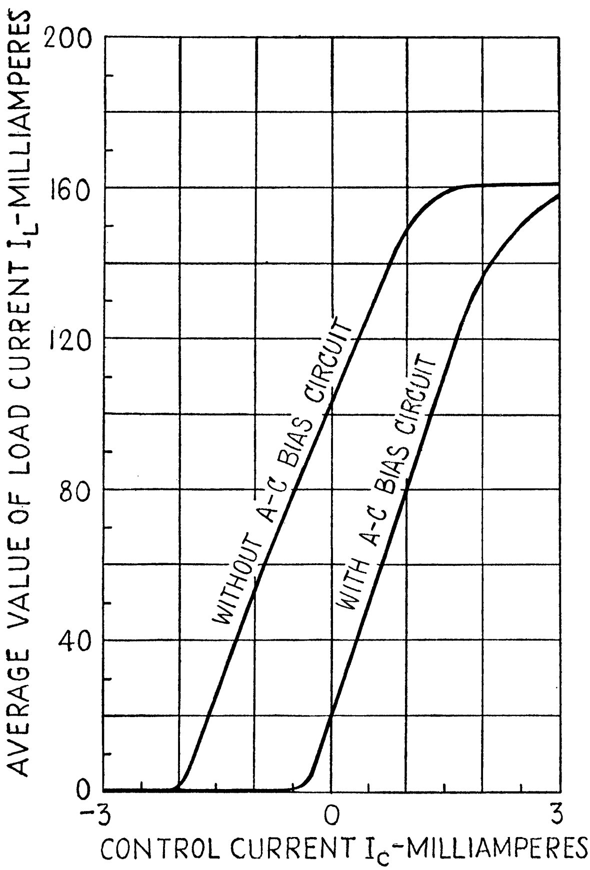

FIG. 7.11. Load current I_L as a function of control current I_C, without a-c bias (I_Q = 103 ma) and with a-c bias (I_Q = 20 ma). E_P = 16 volts; R_L = 3 ohms.

2. Middle d-c feedback windings N'_F, N''_F of 1,000 turns each of No. 28 wire (Brown and Sharpe gauge)
3. Outer d-c control windings N'_C, N''_C of 200 turns each of No. 28 wire (Brown and Sharpe gauge)

The magnetic circuit of each core had an effective area of

$$0.084 \text{ in.}^2 = 0.54 \text{ cm}^2$$

and a mean length of 5.7 in. = 14.5 cm.

The measurements, with power-supply frequency f_P = 60 cps, on the test circuit of Fig. 7.4*b*, using these Mumetal tape cores, determined various characteristic curves, as indicated in the legends of Figs. 7.9

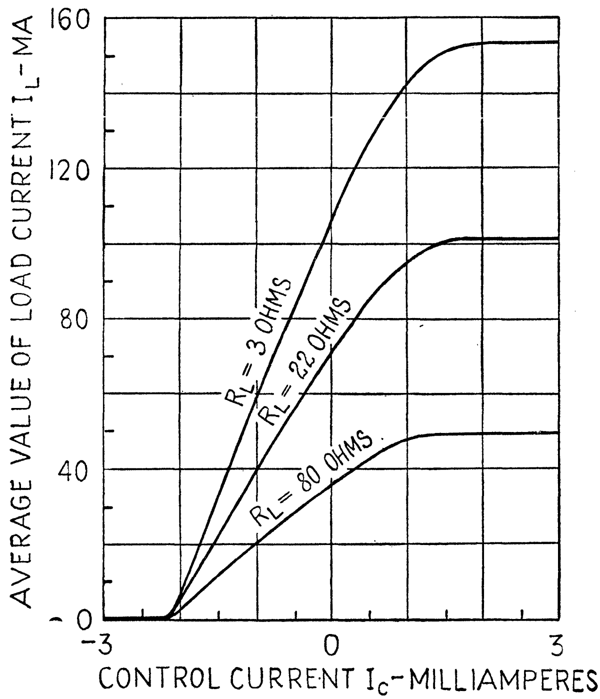

FIG. 7.12. Load current I_L as a function of control current I_C with load resistance R_L as a parameter, without bias. $E_P = 16$ volts.

to 7.12. Discussing the results of these measurements, as presented in the figures, it was learned that:

1. Actual magnitude of quiescent current I_Q will decrease when the load resistance R_L is increased.
2. The magnitude of power-supply voltage E_P may vary within the limits of from 16 to 18 volts without appreciably changing the slope of the characteristic $I_L = f(I_C)$.
3. This slope will decrease when E_P is reduced to values of about 12 to 14 volts.
4. This slope will decrease when the load resistance R_L is increased.
5. Application of an excessively high power-supply voltage, $E_P = 20$ volts or more, will prevent "cutoff" conditions.
6. The quiescent current I_Q may be reduced to one-fifth of its original value (Fig. 7.11), or to a much smaller value of about 2 to 6 per cent of its original value (Fig. 7.12), when the a-c bias circuit of Fig. 7.4*b* is used.

7.10. Literature Concerning Theoretical Investigations on External-feedback Magnetic-amplifier Circuits. Besides the authors already referred to in Chaps. 6 and 7, contributions describing theoretical investigations on the operating conditions of external-feedback magnetic-amplifier circuits have been made by S. E. Tweedy,[16] E. H. Frost-Smith,[17] S. E. Hedström and L. F. Borg,[18] A. G. Milnes,[19,20] H. M. Gale and P. D. Atkinson,[21] D. W. VerPlanck, M. Fishman, and D. C. Beaumariage,[22] J. H. Reyner,[23] W. J. Dornhoefer and V. H. Krummenacher,[24] and F. E. Butcher and R. Willheim.[25]

Further literature concerning the development of various external-feedback types of magnetic-amplifier circuits and special applications will be cited in the next chapter.

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3. B. Nordfeldt, "Polyphase Direct Current Saturated Inductance," United States patent 2,372,112, issued March 20, 1945 (application filed July 31, 1940; in Sweden, August 1, 1939).
4. See "Einrichtung mit mindestens einer regelbaren Drossel" (Arrangement having at least one controlled saturable reactor), Swiss patent 226,104, issued March 15, 1943 (application filed October 17, 1941; in Germany, August 21, 1940).
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7. See "Magnetischer Verstärker" (Magnetic amplifier), Swiss patent 228,534, issued August 31, 1943 (application filed May 13, 1942), Figures 3 and 6.
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CHAPTER 8

Duodirectional External-feedback Circuits with Two Saturable-reactor Elements

8.1. Classification of Duodirectional External-feedback Circuits. The conventional method for obtaining duodirectional transfer characteristics (Fig. 5.1*c*) on external-feedback-type magnetic amplifiers consists in providing two equally rated polarized units, as described in Chaps. 6 and 7, in a balanced arrangement, preferably a push-pull circuit. Thus, such an arrangement contains four saturable-reactor elements in a symmetrical a-c bridge or differential-type circuit, similar to the duodirectional nonfeedback circuits described in Sec. 5.4.

In 1952, the author[1] suggested another method for designing duodirectional types of magnetic amplifiers providing only *two* saturable-reactor elements. Two-stage magnetic servo amplifiers, based upon this method, have proved to give very satisfactory results, particularly when used in connection with those types of two-phase induction motors having such a low ratio of effective viscous friction to inertia that it is difficult to stabilize at high gain in a closed-loop servomechanism.

Therefore, duodirectional external-feedback magnetic-amplifier circuits may be classified in the following way:

Group 1: Push-pull circuits with *two* saturable-reactor elements
- *a.* With a-c load
- *b.* With d-c load

Group 2: Push-pull circuits with *four* saturable-reactor elements
- *a.* With a-c load
- *b.* With d-c load

This chapter describes typical examples of the first group of these circuits.

8.2. Push-Pull Circuits with Two Saturable-reactor Elements. The basic idea consists in feeding the two control winding elements N'_C and N''_C of the original external-feedback circuit with a-c or d-c load components (Fig. 8.1) in such a way that push-pull action is obtained. These winding elements may be incorporated into two separate control circuits (d-c or a-c), or they may be series-connected or parallel-connected in a common control circuit (d-c or a-c).

The fundamental principle of the duodirectional external-feedback circuit with only two saturable-reactor elements can be explained in a simple way by comparing its mode of operation with that of the original external-feedback circuit, as described in Chap. 6.

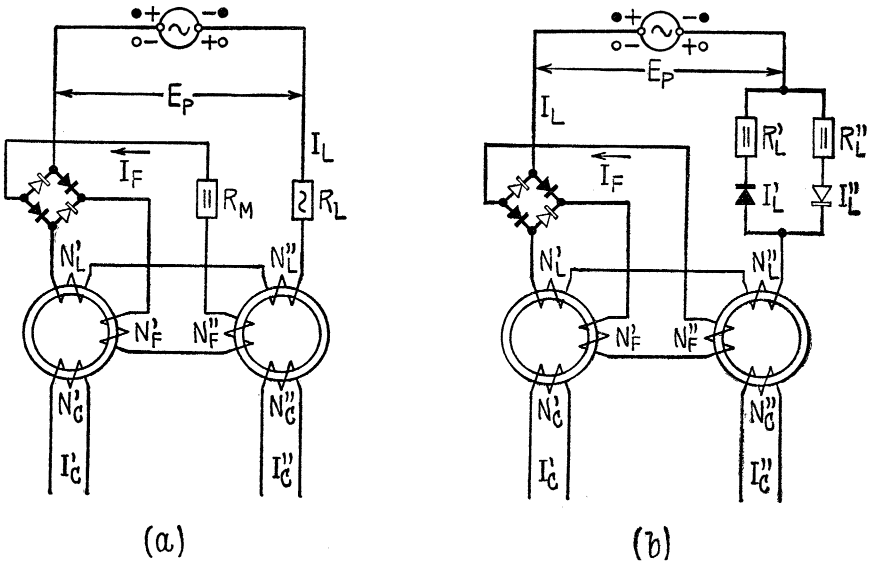

FIG. 8.1. Basic types of push-pull external-feedback circuits with two saturable-reactor elements: (*a*) with a-c load component R_L and d-c load component R_M; (*b*) with a special "splitting circuit" containing two half-wave-rectifier elements with the associated d-c load components R'_L (first-half-cycle pulses I'_L) and R''_L (second-half-cycle pulses I''_L).

a. Mode of Operation of the Original External-feedback Circuit. Figure 8.2 illustrates four different possibilities for controlling the external-feedback circuit by means of a reversible control voltage E_C. Figure 8.3 shows that, in this case, the half-cycle pulses I'_L, I''_L and I'_M, I''_M will both increase or decrease when the d-c (or a-c) control voltage E_C is changed from zero to positive or negative values (or from zero to values having 0 or 180° phase displacement between E_C and E_P, respectively). Thus, the actual impedance values of the two saturable-reactor elements will at all times decrease or increase simultaneously, as indicated in Fig. 8.2 by the same symbol ($\pm$) in both core elements.

b. Mode of Operation of the Duodirectional External-feedback Circuit. Figure 8.4 illustrates four different possibilities for controlling the external-feedback circuit in such a way that push-pull action is obtained, using a reversible control voltage E_C, which may be either d-c or a-c.

Figure 8.5 shows that, in this case, the half-cycle pulses I'_L and I'_M will increase and the half-cycle pulses I''_L and I''_M will decrease (and vice versa), when the d-c (or a-c) control voltage E_C is changed from zero to positive or negative values (or from zero to values having 0 or 180°

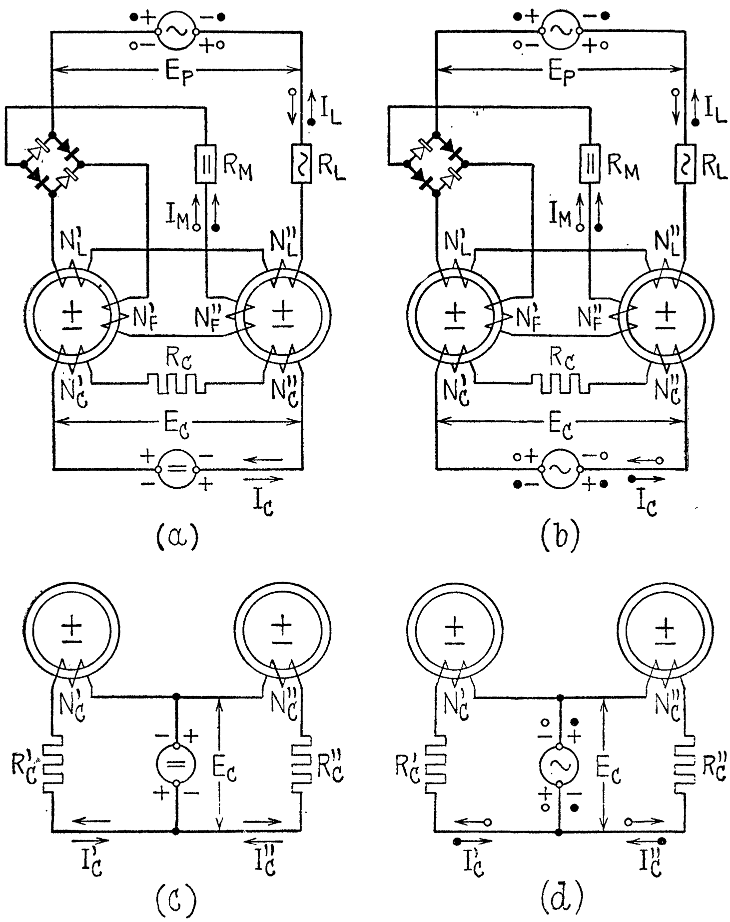

FIG. 8.2. Four diagrams illustrating the mode of operation of the original external-feedback circuit: (*a*) d-c control with series-connected control windings N'_C and N''_C; (*b*) a-c control with series-connected control windings N'_C and N''_C; (*c*) d-c control with parallel-connected control windings N'_C and N''_C; (*d*) a-c control with parallel-connected control windings N'_C and N''_C.

phase displacement between E_C and E_P, respectively). Thus, the actual impedance values of the two saturable-reactor elements will always change in opposite directions, as indicated in Fig. 8.4 by opposite symbols ($\pm$ and $\mp$) in the two core elements.

Referring to special problems which are involved in push-pull external-feedback circuitry using only two saturable-reactor elements, it is important to stress the following facts:

1. The original d-c-controlled external-feedback circuits, as described in Chaps. 6 and 7, have the unique feature that the fundamental-frequency voltages, which are induced in the series-connected

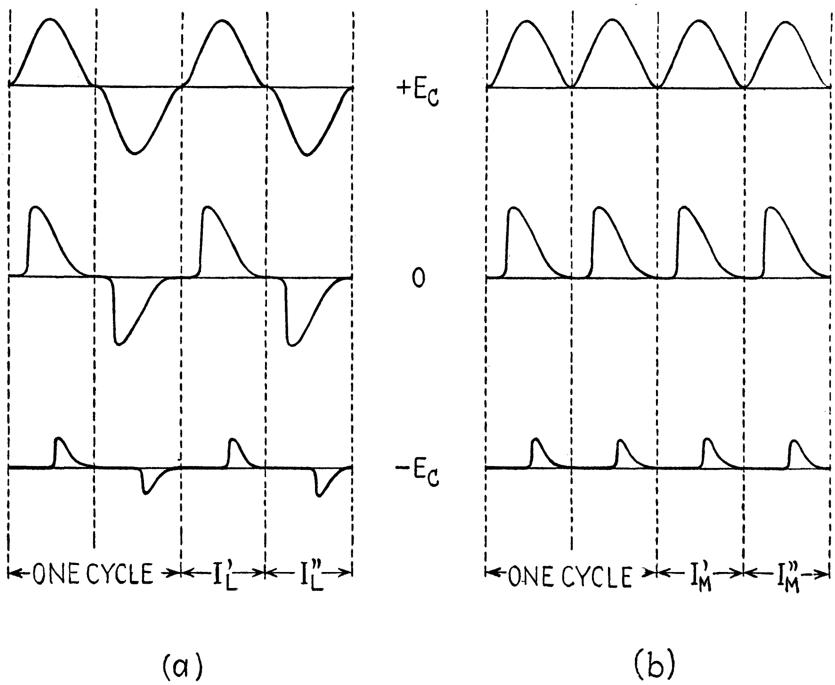

FIG. 8.3. Waveshapes of the output currents of the original external-feeback circuit of Fig. 8.2: (*a*) waveshapes of output current I_L flowing in load component R_L; (*b*) waveshapes of output current I_M flowing in load component R_M.

control winding elements N_C' and N_C'', are always canceling each other completely, not only with no-signal conditions but even when the maximum value of control voltage E_C is applied to the amplifier.
2. However, the a-c controlled push-pull external-feedback circuit with series-connected control winding elements N_C' and N_C'' (Fig. 8.4*b*) has the typical property that the voltages induced in each of the elements N_C' and N_C'' by the fundamental wave and by the odd harmonics of power-supply voltage E_P will cancel each other only for no-signal conditions; otherwise fundamental-frequency voltages having considerable magnitudes may be induced in the control circuit. Moreover, the d-c-controlled circuit of Fig. 8.4*a*

with series-connected control winding elements operates in such a way that a substantially constant fundamental-frequency voltage is induced in the control circuit. Thus, the actual turns ratio $N_C'/N_L' = N_C''/N_L''$ must have a sufficiently small value (about

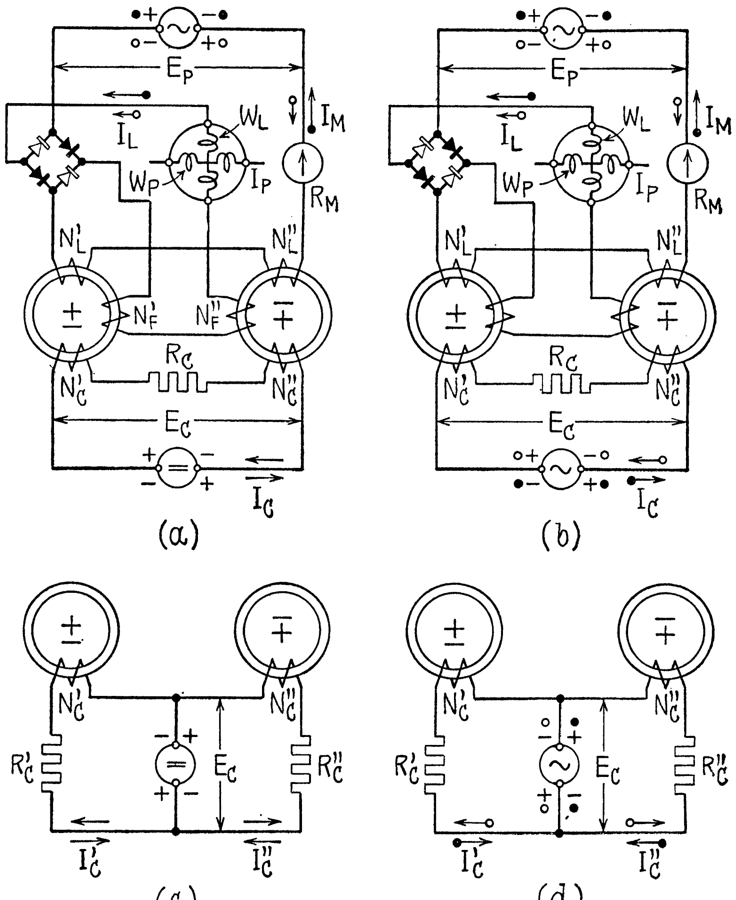

FIG. 8.4. Four diagrams illustrating the mode of operation of the duodirectional external-feedback circuit: (*a*) d-c control with series-connected control windings N_C' and N_C''; (*b*) a-c control with series-connected control windings N_C' and N_C''; (*c*) d-c control with parallel-connected control windings N_C' and N_C''; (*d*) a-c control with parallel-connected control windings N_C' and N_C''.

0.01 to 0.1). Furthermore, it will be necessary to connect a sufficiently high resistance R_C in series with N_C' and N_C'' so that the voltages induced from the load windings N_L', N_L'' into N_C' and N_C'' are unable to produce excessively large currents in the control-circuit loop.

8.3. Practical Applications of Push-Pull External-feedback Circuits with Two Saturable-reactor Elements. Reference to Fig. 8.5 shows that, in both cases *a* and *b*, the output current, being composed of the two half-cycle pulses I'_L and I''_L, has quite unusual waveshapes, which may be illustrated in the following way:

Operating Conditions a. With no-signal conditions ($E_C = 0$), the two half-cycle pulses I'_M and I''_M have exactly the same magnitude.

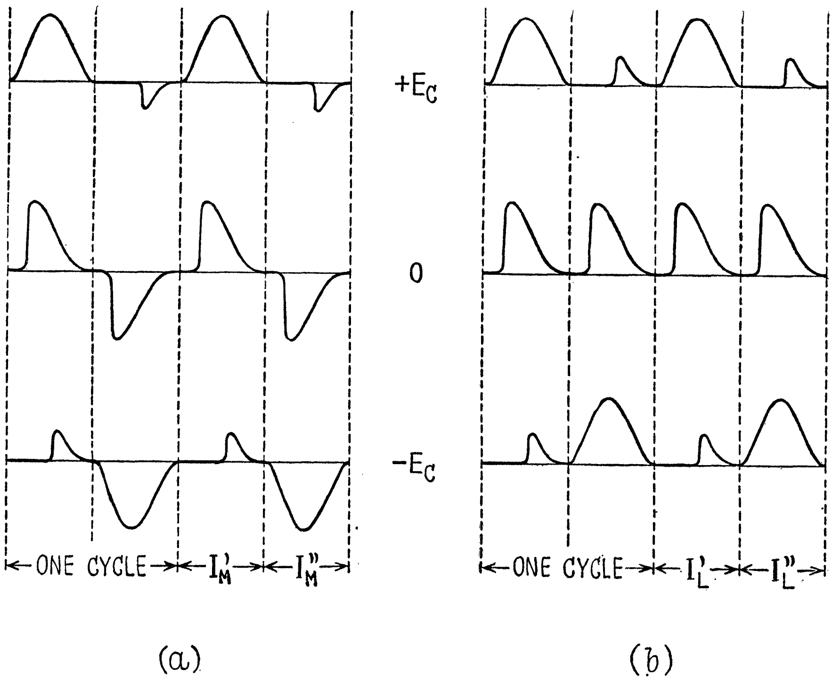

FIG. 8.5. Waveshapes of the output currents of the duodirectional external-feedback circuit of Fig. 8.4: (*a*) waveshapes of output current I_M flowing in load component R_M; (*b*) waveshapes of output current I_L flowing in load component W_L.

Thus, the output current I_M is a pure alternating current having no d-c component. When a positive control voltage $+E_C$ is applied, the first-half-cycle pulses I'_M will increase, while the second-half-cycle pulses I''_M will decrease; and there is a positive d-c component proportional to $I'_M - I''_M$, which is a function of the magnitude of $+E_C$. With negative values, $-E_C$, the first-half-cycle pulses I'_M will decrease, while the second-half-cycle pulses I''_M will increase; and there is a negative d-c component proportional to $I''_M - I'_M$, which is a function of the magnitude of $-E_C$.

The reversible d-c component $\pm I_{dc} = f(\pm E_C)$ may be measured by means of a moving-coil ammeter and can preferably be used for supplying the d-c control windings of a succeeding push-pull magnetic-amplifier stage. The actual a-c component I_{ac}, which is also a function of E_C, may be either eliminated (*e.g.*, by means of a shunt condenser across the moving-coil ammeter) or can be used for supplying the a-c bias windings of such an amplifier stage.

Operating Conditions b. With no-signal conditions ($E_C = 0$), the two half-cycle pulses I'_L and I''_L have exactly the same magnitude. Under these conditions, the output current I_L is a unidirectional current having a d-c component and a superimposed even-harmonic-frequency component, but no fundamental-frequency component. When a positive control voltage $+E_C$ is applied, the first-half-cycle pulses I'_L will increase, while the second-half-cycle pulses I''_L will decrease; and there is a "positive" (0°) fundamental-frequency component, proportional to $I'_L - I''_L$, which is a function of the magnitude of $+E_C$. With negative values, $-E_C$, the first-half-cycle pulses I'_L will decrease, while the second-half-cycle pulses I''_L will increase; and there is a "negative" (180°) fundamental-frequency component, proportional to $I''_L - I'_L$, which is a function of the amplitude of $-E_C$.

The reversible fundamental-frequency component $\pm I_{ac} = f(\pm E_C)$ may be measured by means of a separately excited electrodynamic wattmeter and can preferably be used for controlling a two-phase induction-type reversible motor, as used in servomechanisms. The actual d-c component I_{dc}, which is also a function of E_C, can successfully be used for introducing effective dynamic braking of the two-phase induction motor. This is particularly valuable in those types of reversible a-c motors which have virtually no damping properties, so that additional damping methods must be applied to prevent overshooting or hunting.

8.4. Two-stage Magnetic Servo Amplifier Having Inherent Dynamic-braking Properties. The following illustration of a two-stage magnetic servo amplifier of the push-pull external-feedback type gives a typical example for practical applications of the two basic circuits shown in Fig. 8.4*a* and *b*.

Referring to the output-stage circuit (Fig. 8.4*a*), it is to be noted that the basic idea consists in supplying the amplifier field winding W_L of the two-phase motor from the feedback circuit carrying the unidirectional current I_L having waveshapes as characterized in Fig. 8.5*b*. The line field winding W_P of this motor is supplied with a substantially constant current I_P, which is proportional to the power-supply voltage E_P (power-supply frequency $f_P = 60$ or 400 cps). It is important to stress that this output-stage circuit can be designed so that the d-c component producing dynamic braking decreases when the torque-producing fundamental-frequency (60- or 400-cycle) component of I_L increases, and vice versa.

Referring to the input-stage circuit (Fig. 8.4*b*), it is to be noted that the basic idea consists in using this circuit for various purposes which can be defined as follows:

1. To act as a push-pull type of *magnetic amplifier* to produce a reversible d-c component of load current I_M, a function of the reversible a-c error voltage E_C, for controlling the winding elements N'_C, N''_C of the output-stage circuit (Fig. 8.4*a*)
2. To act as a *phase discriminator* (in a similar way to a phase-sensitive rectifier) so that actual direction of the d-c component of I_M will correspond to the actual phase displacement (0 or 180°) between error voltage E_C and power-supply voltage E_P
3. To act as a *voltage-sensitive bias element* to the output-stage circuit of Fig. 8.4*a* for introducing special voltage-current characteristics (Sec. 7.6)
4. To act as a *decoupling device* in such a way that no appreciable fundamental-frequency voltage will be induced in the synchrotransformer a-c control circuit, which has a total impedance of about 10,000 ohms

The designer, when outlining the circuitry to be used in this special case, faces the following problem: The output-current components of the push-pull external-feedback input-stage circuit, which has only two saturable-reactor elements, must provide:

1. An *a-c bias* component, which is a function of the actual magnitude of the power-supply voltage E_P
2. A reversible *d-c control* component, which is a function of the actual a-c error voltage E_C in the synchrotransformer circuit

Furthermore, it is to be noted that, in contrast to conventional d-c control circuitry, comparatively large fundamental-frequency voltages will be induced in the two control winding elements N'_C, N''_C of the output-stage circuit (Fig. 8.4*a*), even when the actual turns ratio

$$\frac{N'_C}{N'_L} = \frac{N''_C}{N''_L}$$

has a very small value (about $\frac{1}{40}$).

Figure 8.6 shows the complete two-stage circuit, which contains the following components:

1. Two equally rated saturable-reactor elements, preferably having Orthonol tape cores, with
   a. Load windings N'_L, N''_L
   b. Feedback windings N'_F, N''_F

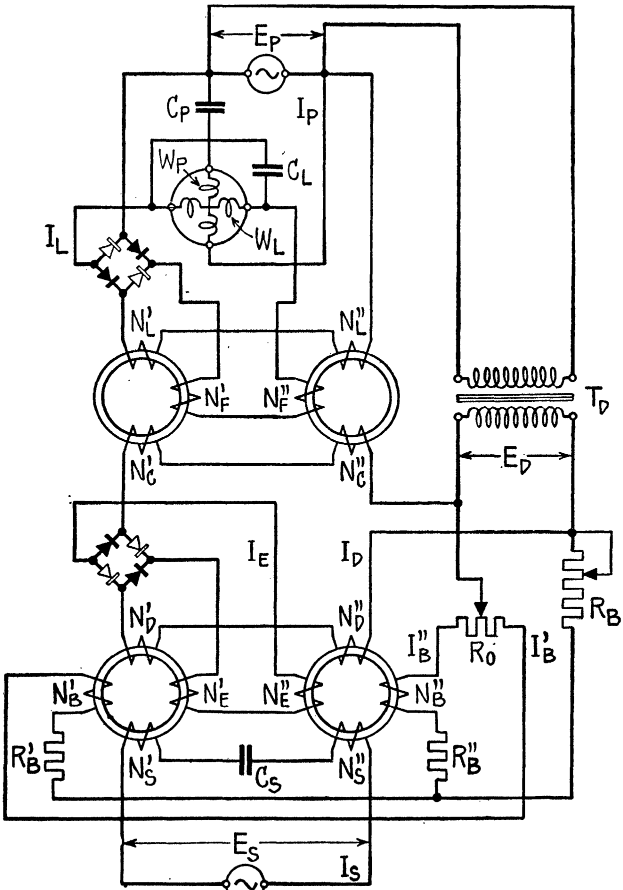

FIG. 8.6. Complete two-stage circuit of a magnetic servo amplifier having inherent dynamic braking characteristics. (*W. A. Geyger.*[1])

   c. D-c control windings N'_C, N''_C acting simultaneously as a-c bias windings
2. A full-wave bridge-type selenium rectifier
3. The amplifier field winding W_L of the two-phase reversible induction motor with shunt condenser C_L
4. The line field winding W_P of this motor with series condenser C_P
5. Two equally rated saturable-reactor elements, preferably having Mumetal tape cores, with
   a. Load windings N'_D, N''_D
   b. Feedback windings N'_E, N''_E
   c. A-c control windings N'_S, N''_S with series condenser C_S

   d. A-c bias windings N'_B, N''_B with fixed series resistors R'_B, R''_B and a potentiometer-type resistor R_0 for zero adjustment
   e. A variable bias resistor R_B for controlling the actual magnitude of the alternating bias currents I'_B and I''_B
6. A full-wave bridge-type selenium rectifier
7. A small transformer T_D (*e.g.*, $E_D = 25$ volts)

Correct and efficient operation of the complete two-stage circuit will be obtained only when the a-c (60- or 400-cycle) components of the currents (I_P, I_L, I_D, I_S) flowing through the various parts of the circuit have a proper phase relationship. The example of Fig. 8.6 shows a simple way for obtaining proper phase displacement by means of three condensers (C_P, C_L, C_S) so that the most favorable working conditions will be secured. This two-stage circuit is the result of a special experimental study, and the optimum values of the condensers, which are not critical, have been determined empirically.

It is to be noted that actual magnitude of the dynamic-braking effect, which is determined by the actual magnitude of the output-stage quiescent current (I_L with $E_S = 0$), can easily be adjusted by means of the bias resistor R_B, which controls the actual magnitude of the a-c bias currents I'_B, I''_B of the input-stage circuit.

Figure 8.7 illustrates three different operating conditions of the full-wave output-stage circuit of the two-stage arrangement of Fig. 8.6 having inherent dynamic-braking characteristics.

Figure 8.7*a*: With $E_C = 0$ the two half-cycle pulses I'_L and I''_L are equal and very small, and there is practically *no damping effect,* because the d-c component of $I'_L + I''_L$ is extremely small. However, $I'_L + I''_L$ and the corresponding damping effect increase when E_C increases. This is generally not desirable, because the speed of the motor is decreased with higher values of error voltage and because there is practically no damping effect in the vicinity of the null point.

Figure 8.7*b*: With $E_C = 0$ the two half-cycle pulses I'_L and I''_L are about 50 per cent of maximum load current, and there will be a large and *constant damping effect* because $I'_L + I''_L$ will have a constant magnitude over the entire range. These working conditions will give satisfactory results in many cases.

Figure 8.7*c*: With $E_C = 0$ the two half-cycle pulses I'_L and I''_L have about their maximum (100 per cent) values, and there will be a very high damping effect. It is to be noted, however, that the damping effect decreases if the error voltage E_C increases. This *variable damping effect,* which is produced by the changing of the d-c component $I'_L + I''_L$, permits faster motor movement toward the null point and slower movement coincident with the lower values of error voltage E_C characterizing the close proximity of the null point. Thus, a very effective dynamic braking of the motor will be obtained in this way. When the a-c (60- or 400-cycle) component of load current $I_L = I'_L - I''_L$ disappears from the amplifier field winding W_L, the comparatively large d-c component of I_L tends to lock the rotor in the position assumed at the instant a-c excitation ceases.

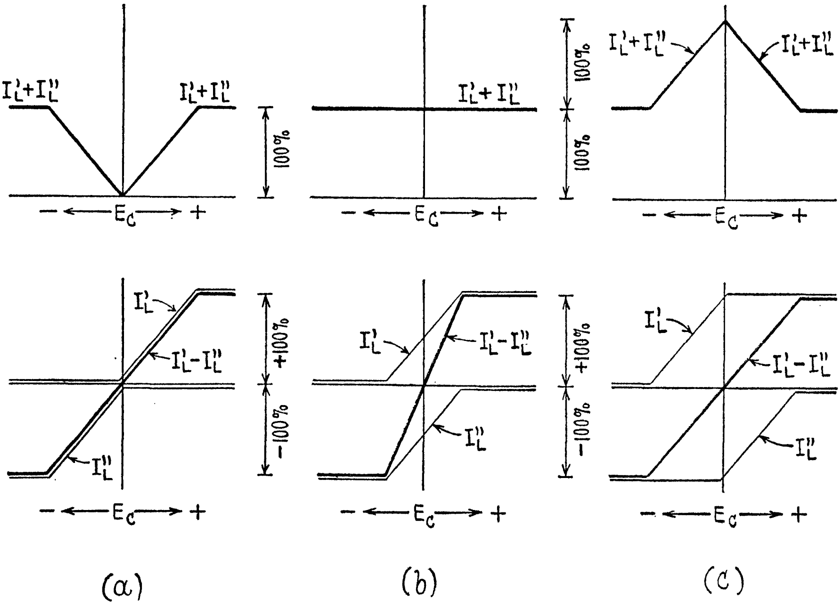

FIG. 8.7. Three different operating conditions of the full-wave output-stage circuit shown in Fig. 8.6. (*a*) There is no damping effect with no-signal conditions ($E_C = 0$). (*b*) There is a constant damping effect. (*c*) There is a variable damping effect in such a way that the d-c component $I_{dc} = I'_L + I''_L$ producing dynamic braking decreases when the torque-producing fundamental-frequency (60- or 400-cycle) component $I_{ac} = I'_L - I''_L$ increases, and vice versa.

Of course, it is possible in special cases to superimpose an additional d-c damping effect in the amplifier field winding (W_L) for further modification of the actual damping conditions.

Referring to the method for obtaining perfect zero adjustment (fundamental-frequency component of $I_L = 0$, if $E_S = 0$), it is to be noted that this can easily be achieved by adjusting the potentiometer resistor R_0, which controls the actual value of current ratio I'_B/I''_B in the input-stage bias circuit.

Application of the small power-supply transformer T_D (*e.g.*, $^{115}\!/_{25}$ volts) makes it possible to reduce the number of cells of the feedback rectifier supplying the feedback windings N'_E, N''_E and to reduce the number of turns of load windings N'_D, N''_D, feedback windings N'_E, N''_E, and output-stage control windings N'_C, N''_C. The transformer T_D, however, may be eliminated by properly varying these circuit components.

Furthermore, in any case, the two input-stage control winding elements N'_S and N''_S, each having its own series condenser, may be connected in parallel to the control voltage E_S of the synchro control transformer to obtain perfect symmetry of the transfer characteristics $I_L = f(E_S)$ by proper rating of these two series condensers.

8.5. Actual Performance Characteristics of a Two-stage Magnetic-servo-amplifier Circuit Having Inherent Dynamic-braking Properties. Figures 8.8 to 8.10 give results of various measurements representing typical examples of actual performance of an output-stage circuit having inherent dynamic-braking characteristics. The two saturable-reactor elements of this circuit (Fig. 8.4*a*) consisted of spirally wound 2-mil Orthonol tape cores having the following dimensions:

Outer diameter, 2.0 in.
Inner diameter, 1⅝ in.
Tape width, 1.0 in.

The tape cores had the following multilayer toroidal windings of enamel-insulated copper wire:

1. Load windings N'_L, N''_L of 4,000 turns each with No. 31 wire (Brown and Sharpe gauge)
2. Feedback windings N'_F, N''_F of 4,000 turns each with No. 31 wire (Brown and Sharpe gauge)
3. Control windings N'_C, N''_C of 100 turns each with No. 31 wire (Brown and Sharpe gauge)
4. Bias windings N'_B, N''_B of 100 turns each with No. 31 wire (Brown and Sharpe gauge)

The magnetic circuit of each core had an effective area of

$$0.168 \text{ in.}^2 = 1.08 \text{ cm}^2$$

and a mean length of 5.7 in. = 14.5 cm.

The test circuit (Fig. 8.4*a*) was d-c controlled by means of a reversible voltage $E_C = 0$ to ± 30 volts, with $R_C = 1{,}000$ ohms. Since no input-stage circuit is used in this case, the afore-mentioned a-c bias windings N'_B, N''_B have been provided for adjusting the quiescent-current values of the two half-cycle pulses I'_L and I''_L. The actual values of I'_L and I''_L can be measured separately by means of an auxiliary "splitting circuit" (Fig. 8.1*b*) containing two half-wave rectifiers and the associated moving-coil milliammeters R'_L and R''_L.

The full-wave bridge-type selenium rectifier contained $4 \times 6 = 24$ cells. The ohmic resistance of the amplifier field winding W_L of the two-phase induction motor was about 1,800 ohms. The line field winding W_P was disconnected in this case.

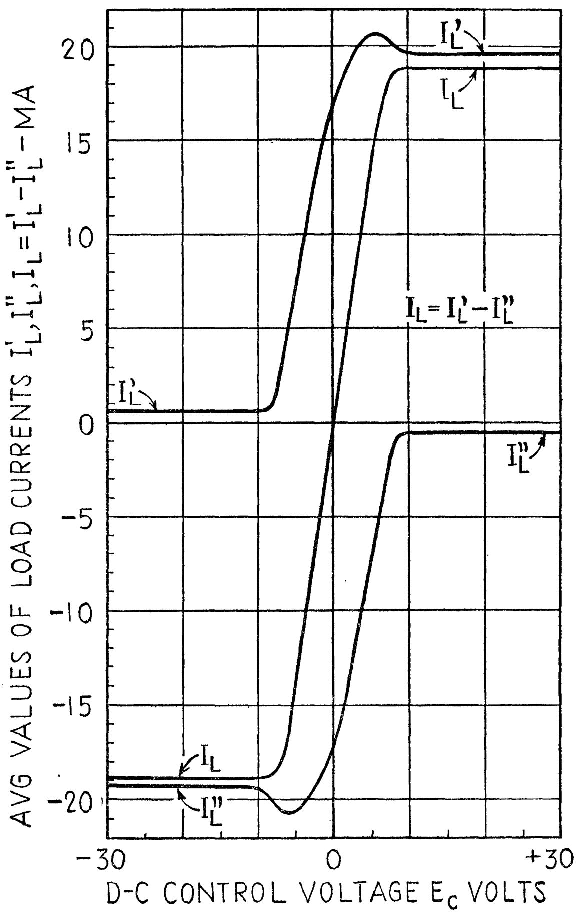

FIG. 8.8. Measured average values of the two half-cycle pulses I'_L and I''_L and resultant a-c component $I_L = I'_L - I''_L$ as a function of d-c control voltage E_C. ($f_P = 60$ cps.)

Figure 8.8 shows the measured values of I'_L and I''_L and $I_{ac} = I'_L - I''_L$ as a function of d-c control voltage E_C, with $R_C = 1{,}000$ ohms. The actual value of quiescent current $I_Q = 17$ ma is so high that the operating conditions characterized in Fig. 8.7*c* are obtained.

Figure 8.9 shows the measured average value of the d-c component $I_{dc} = I'_L + I''_L$ as a function of d-c control voltage E_C, with $R_C = 1{,}000$ ohms. Evidently, with $E_C = 0$, the actual quiescent-current value is

$I_{dc} = 17 + 17 = 34$ ma. It is to be noted that I_{dc} decreases considerably when d-c control voltages, $E_C = \pm 10$ volts ($I_C = \pm 10$ ma), are applied to this circuit.

Figure 8.10 shows the actual voltages of fundamental frequency which are induced from the load windings N'_L, N''_L into two test windings

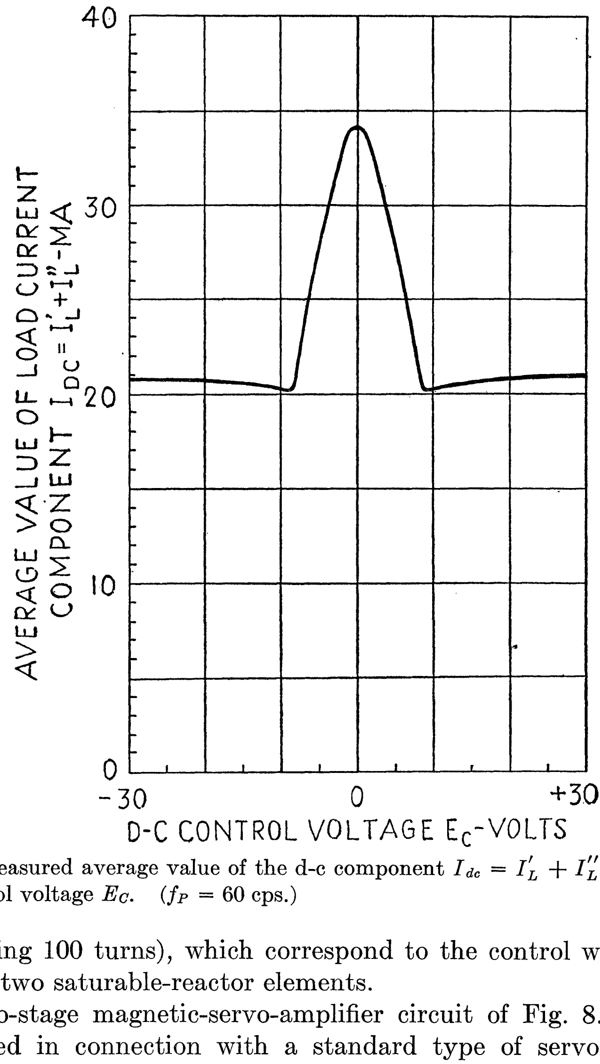

FIG. 8.9. Measured average value of the d-c component $I_{dc} = I'_L + I''_L$ as a function of d-c control voltage E_C. ($f_P = 60$ cps.)

(each having 100 turns), which correspond to the control windings N'_C, N''_C of the two saturable-reactor elements.

The two-stage magnetic-servo-amplifier circuit of Fig. 8.6 has been investigated in connection with a standard type of servomechanism, which is operated by means of a 3.5-watt two-phase reversible induction motor having very poor damping properties and no additional damping means. As a result of various measurements, it was learned that the static error is about 0.05°, while the drift error is about 0.1 to 0.2°.

An important factor governing the satisfactory operation of a closed-loop servomechanism is the proper adjustment of the "sensitivity," or "gain," of the magnetic amplifier. The combination of the error-signal voltage per degree and the actual amplifier gain governs the "tightness"

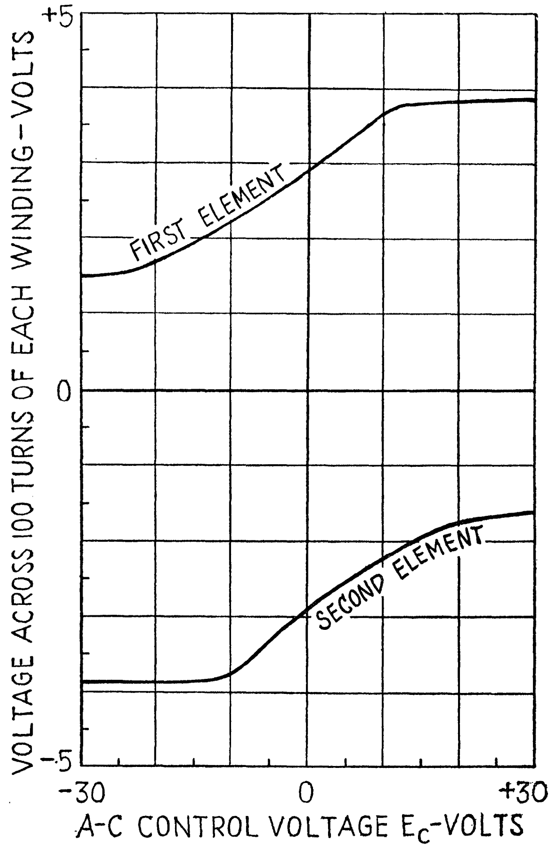

FIG. 8.10. Actual voltages of fundamental frequency which are induced from the load windings N_L', N_L'' into two test windings (each having 100 turns) which correspond to the control windings N_C', N_C'' of the two saturable-reactor elements. (f_P = 60 cps.)

of the system. If the gain is too great, then the servomechanism will oscillate in spite of application of an effective antihunt method. It is often desirable to adjust the gain of the magnetic servo amplifier in such a way that not more than one overshoot, or no overshoot at all, will be obtained. This can be achieved in a simple way by adjusting the bias resistor R_B of the input-stage circuit (Fig. 8.6), as mentioned in Sec. 8.4.

In principle, it will be necessary to eliminate those effects which are caused by the actual differences of the magnetic properties of the saturable-reactor elements and by the actual differences of the dry-disk-rectifier characteristics. It was learned that this may be accomplished by adjusting the potentiometer-type resistor R_0 of the input-stage circuit (Fig. 8.6). This fact means a considerable simplification of the complete circuit for universal application.

REFERENCE

1. W. A. Geyger, "A New Type of Magnetic Servo Amplifier," *AIEE Transactions*, Vol. 71, Part I, pp. 272-280, 1952.

CHAPTER 9

Duodirectional External-feedback Circuits with Four Saturable-reactor Elements

9.1. Classification of Push-Pull External-feedback Circuits with Four Saturable-reactor Elements. It has already been mentioned in Sec. 8.1 that the conventional method for obtaining duodirectional transfer characteristics (Fig. 5.1*c*) on external-feedback-type magnetic amplifiers consists in providing two equally rated polarized units, as described in Chaps. 6 and 7, in a balanced arrangement, preferably a push-pull circuit. In this case, four saturable-reactor elements are incorporated in a symmetrical a-c bridge or differential-type circuit, similar to the duodirectional nonfeedback circuits described in Sec. 5.4.

Generally, these types of magnetic amplifiers represent differential circuits operating on a-c-bridge-network balance conditions: load current $I_L = 0$ if the control current $I_C = 0$. In practice, it is required that fulfillment of this condition (the zero stability) must be mostly independent of changes in magnitude, frequency, and waveshape of the power-supply voltage, major changes in ambient temperature, etc. In other words, the zero drift of the circuit with regard to such changes in operating conditions shall be as small as possible. According to conventional practice in defining the performance of electrical measuring instruments, it is advisable to distinguish between different "drift-rate components," such as "voltage drift," "frequency drift," "temperature drift," etc.

Referring to the voltage drift of push-pull external-feedback circuits with four saturable-reactor elements, it is evident that the quiescent-current values I'_Q and I''_Q of the two polarized magnetic-amplifier units, each having one pair of saturable-reactor elements, must be exactly equal at all times. Thus, actual changes in magnitude of I'_Q and I''_Q, as caused by variations in operating conditions, must also be exactly the same in both amplifier units.

Reference to Fig. 7.10 makes it evident that the actual magnitude of quiescent current I_Q (load current I_L with control voltage $E_C = 0$) will considerably change when E_P is varied (the crossover point on the zero axis, $E_C = 0$, wanders). It has been shown that it is possible to reduce these changes of I_Q to a minimum by applying voltage-sensitive bias circuits which introduce inherent constant-current characteristics (Sec. 7.6). However, with two push-pull-connected external-feedback circuit units operating with very low input-signal level under compensated-feedback conditions ($N_F = N_L$), it will be necessary to match their characteristics over a large part of the working range, to avoid excessive zero drift.

In 1939, the author[1,2] designed another type of push-pull external-feedback circuit with four saturable-reactor elements utilizing "true positive feedback" to obtain better stability with a smaller zero drift, particularly in low-power-level input-stage arrangements. Since, with no-signal conditions, there is no d-c component in the external-feedback windings, the four saturable-reactor elements must be biased to the operating point by means of a properly rated d-c bias circuit. In this case, the actual voltage-current characteristics $I_Q = f(E_P)$ will correspond to nonfeedback conditions so that excellent stability and an extremely small zero drift will be obtained. These arrangements have been characterized as "biased types of push-pull external-feedback circuits." They are also termed "differential-feedback-type circuits," because the actual magnitude of the feedback current I_F (or of the difference of two feedback currents I'_F and I''_F) corresponds to the actual difference between the two load-current components I'_L and I''_L.

Therefore, duodirectional external-feedback circuits with four saturable-reactor elements will be classified in the following way:

Group A: Circuits with two saturable-reactor units, each having its own feedback current, I'_F and I''_F, respectively, corresponding to the basic external-feedback circuits, as described in Chaps. 6 and 7
  1. With a-c load
  2. With d-c load

Group B: Circuits with two saturable-reactor units being additionally controlled by a differential-feedback effect which corresponds to the actual difference between the two load-current components I'_L and I''_L
  1. With a-c load
  2. With d-c load

This chapter describes typical examples of both groups of these circuits.

9.2. Basic Differential Output Circuits for Magnetic Amplifiers of the Push-Pull External-feedback Type with Four Saturable-reactor Elements. These circuits may be classified as follows:

1. Circuits for an a-c load
   a. Circuit with power-supply transformer
   b. Circuit with load transformer
2. Circuits for a d-c load
   a. Series mixing circuit
   b. Parallel mixing circuit
   c. Twin-load circuit

At first, these basic differential output circuits will be briefly described, with reference to the literature (W. F. Horton,[3] F. E. Butcher and

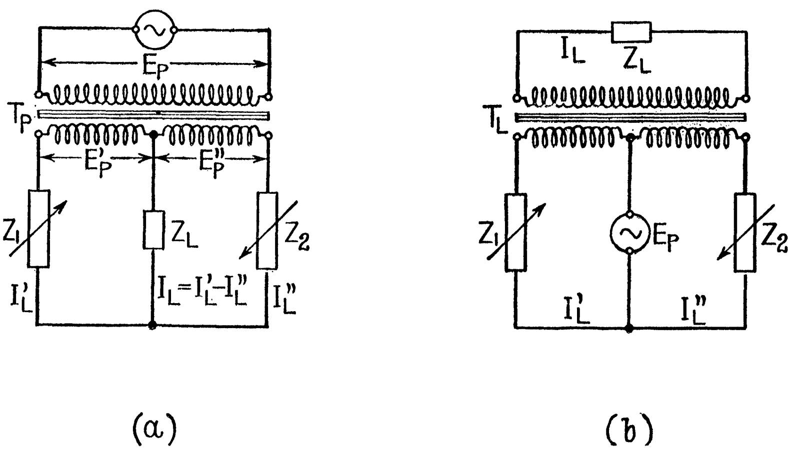

FIG. 9.1. The equivalent circuits of push-pull-type magnetic amplifiers with a-c load Z_L (*a*) with power-supply transformer T_P; (*b*) with load transformer T_L.

R. Willheim[4]) which gives the complete analysis and the condition for maximum power output for each of these circuits.

Circuit with A-C Load and Power-supply Transformer. Figure 9.1*a* shows the equivalent circuit of a magnetic amplifier containing two equally rated saturable-reactor units and a power-supply transformer T_P with center-tapped secondary windings. This type of network, which frequently arises in either power or communication circuits, is that of two a-c generators—voltages E'_P and E''_P—feeding a common load Z_L, the impedances Z_1 and Z_2 representing the actual impedance values of the two saturable-reactor units, which are functions of the control current I_C. With $I_C = 0$, the impedances Z_1 and Z_2 will be equal, and the alternating load current I_L will be zero. But with a "positive" or "negative" control current, $+I_C$ or $-I_C$, an alternating load current $I_L = I'_L - I''_L$ will flow, according to the individual transfer characteristics of the two saturable-reactor units:

$$I_L = E_P \frac{Z_2 - Z_1}{Z_1 Z_2 + (Z_1 + Z_2) Z_L} \tag{9.1}$$

Circuit with A-C Load and Load Transformer. Figure 9.1*b* shows the equivalent circuit of a magnetic amplifier containing two equally rated saturable-reactor units and a load transformer T_L with center-tapped

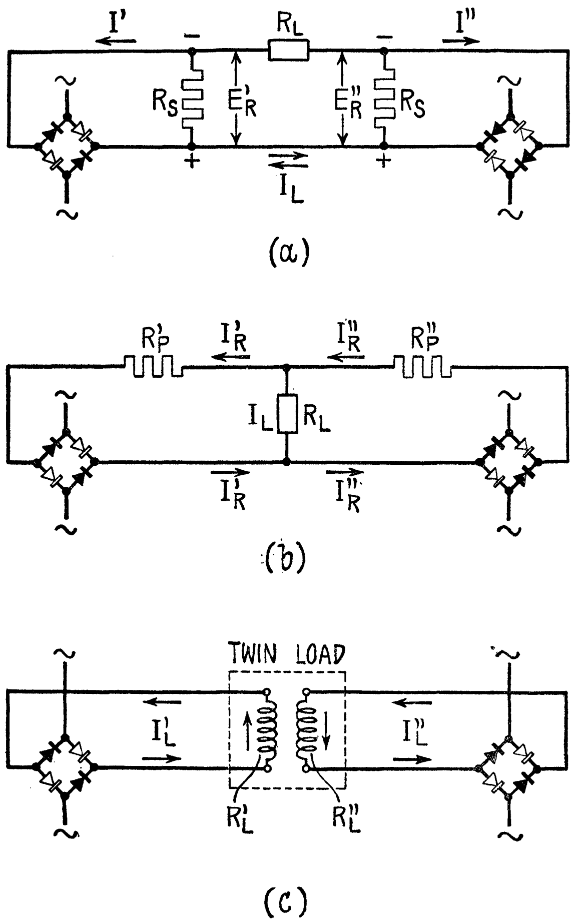

FIG. 9.2. Basic differential output circuits of push-pull-type magnetic amplifiers with d-c load R_L: (*a*) the series mixing circuit based upon the superposition of two voltages E'_R and E''_R; (*b*) the parallel mixing circuit based upon the superposition of two currents I'_R and I''_R; (*c*) the twin-load circuit based upon the superposition of ampere-turns produced by the output-current components I'_L and I''_L in the twin load having the load resistances R'_L and R''_L.

primary windings. This differential transformer feeds the load impedance Z_L with a secondary current I_L corresponding to the actual difference of the two primary currents:

$$I_L = \text{const} \times (I'_L - I''_L) \tag{9.2}$$

Series Mixing Circuit. According to Fig. 9.2*a*, voltage drops E'_R and E''_R, proportional to direct output currents I' and I'', are superimposed, so that the reversible current $\pm I_L = I'_L - I''_L$ flowing through the load resistor R_L corresponds to the reversible control current $\pm I_C$.

In order to find the direct load-current components I'_L and I''_L, each load-current component will be considered separately, according to the superposition theorem:

$$I'_L = I' \frac{R_S}{2R_S + R_L} \tag{9.3a}$$

$$I''_L = I'' \frac{R_S}{2R_S + R_L} \tag{9.3b}$$

$$I_L = I'_L - I''_L = (I' - I'') \left(\frac{R_S}{2R_S + R_L}\right) \tag{9.3c}$$

Parallel Mixing Circuit. According to Fig. 9.2*b*, direct currents I'_R and I''_R, representing converted alternating output currents, are superimposed so that I'_R and I''_R are in opposite direction in the load resistor R_L. For proper operation, ballast resistors R'_P and R''_P, having resistance about three times that of the load resistance R_L, are needed in series with the rectifier outputs to prevent one rectifier from by-passing part of the load current.

Twin-load Circuit. In this case (Fig. 9.2*c*), a differential-type load having two separate load resistors R'_L and R''_L is used, *e.g.*, two equal, separate d-c input control winding arrangements of the next magnetic-amplifier stage (Fig. 9.8). This method not employing passive resistor elements—R_S in Fig. 9.2*a* and R_P in Fig. 9.2*b*—is based upon the superposition of two d-c magnetizations (two d-c ampere-turns) produced by two separate load components acting in opposite directions as a differential-type device. Thus, the twin-load circuit involves no loss of power, since no mixing resistor network is needed, and this method should preferably be used if convenient.

9.3. Push-Pull Types of Magnetic Amplifiers with Two Separate External-feedback Circuits. Figure 9.3 shows a typical example[5] of a magnetic amplifier of the balance-detector type containing:

1. Two saturable-reactor units with separate external-feedback circuits N'_F, N''_F
2. A power-supply transformer T_P with center-tapped secondary windings (Fig. 9.1*a*)
3. A d-c bias circuit with a-c resistors R'_S and R''_S, d-c resistors R'_B and R''_B, and potentiometer resistor R_0 for zero adjustment
4. An a-c load, *e.g.*, the amplifier field winding of a separately excited induction-type two-phase reversible motor

The control winding units N'_C and N''_C carrying the reversible direct control current $I_C = E_C/R_C$ act in such a way that push-pull action is obtained. With "positive" or "negative" control current, $+I_C$ or $-I_C$, feedback current I'_F increases, while feedback current I''_F decreases, or vice versa.

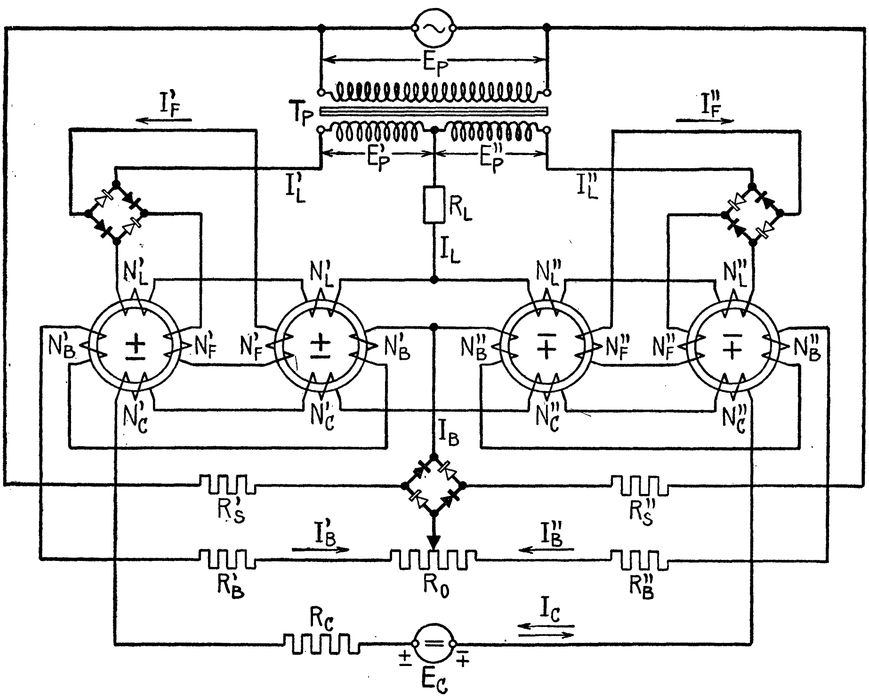

FIG. 9.3. Push-pull external-feedback circuit with a-c load R_L and power-supply transformer T_P corresponding to Fig. 9.1*a*. (*W. A. Geyger, United States patent 2,338,423, Figure* 1, 1938.)

A complete analysis of this circuit was given in 1942 by T. Buchhold.[6] He considers especially the following operating conditions:

1. The actual excursions of control current $\pm I_C$ are very small.
2. The resistance value of the a-c load R_L is very *low*. In this case, the circuits act like a *generator with high internal impedance*, so that the actual magnitude of load current I_L is substantially independent of changes in load resistance R_L.
3. The resistance value of the a-c load is very *high*. In this case, the circuits act like a *generator with low internal impedance*, so that the actual magnitude of load voltage $E_L = I_L R_L$ is substantially independent of changes in load resistance R_L.

4. A properly rated shunt condenser is connected across the a-c load R_L to increase the power amplification of the circuit.
5. A properly rated condenser is series-connected with the a-c load R_L to increase the power amplification of the circuit.
6. Properly rated filter circuits are provided to reduce the waveshape distortion of load current I_L flowing through the a-c load R_L.

Buchhold's analysis is especially interesting with regard to the design procedure when several push-pull magnetic-amplifier circuits are to be connected in cascade.

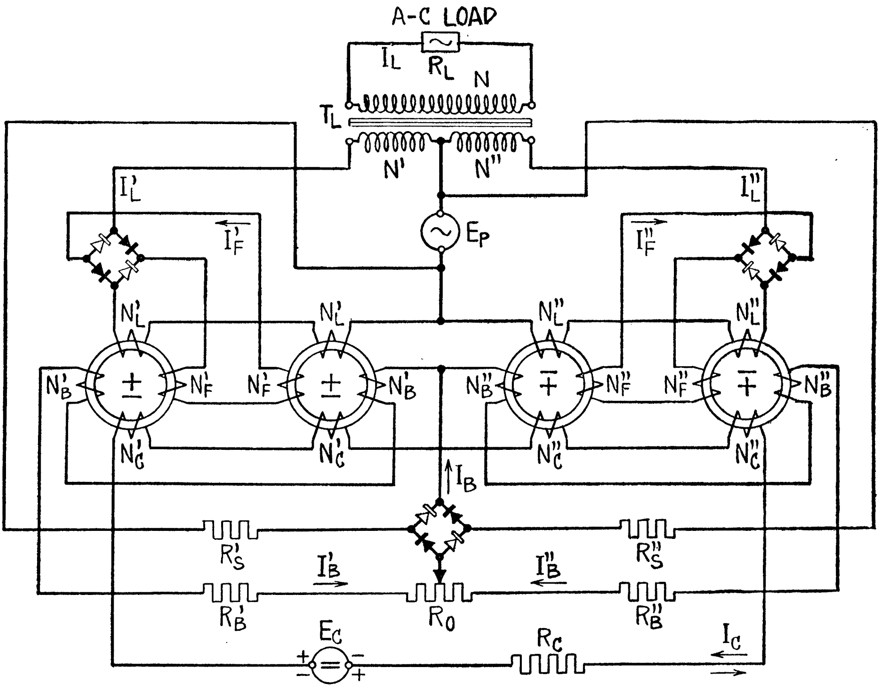

FIG. 9.4. Push-pull external-feedback circuit with a-c load R_L and load transformer T_L corresponding to Fig. 9.1*b*. (*W. A. Geyger, United States patent* 2,338,423, *Figure* 3, 1938.)

Figure 9.4 illustrates a similar arrangement which is based upon the differential transformer circuit of Fig. 9.1*b*.

Figure 9.5 shows a modified push-pull external-feedback circuit with d-c twin load $W'_L W''_L$ and cross-connected winding units $N'_B N''_F$, $N''_B N'_F$, as devised in 1938 by G. Barth.[7] These cross-connected winding units act simultaneously as d-c bias windings and d-c feedback windings in such a way that positive and negative feedback will be obtained. It is possible to bring the quiescent-current values of the two output components I'_L = const $\times$ I'_P and I''_L = const $\times$ I''_P to the desired magnitude by proper rating of the turns ratio $N'_F/N'_B = N''_F/N''_B$. Thus, the additional d-c bias circuit, as used in the arrangements of Figs. 9.3 and 9.4, can be eliminated. This means a substantial simplification of the design of such circuits.

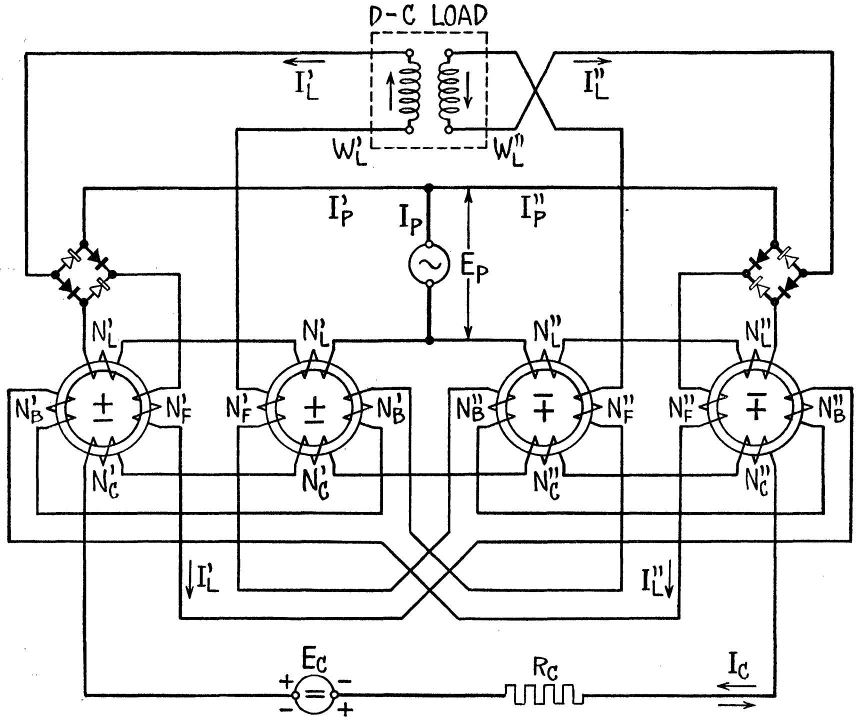

FIG. 9.5. Push-pull external-feedback circuit with d-c twin load $W'_L W''_L$ and cross-connected winding units $N'_B N''_F$ and $N''_B N'_F$ acting simultaneously as d-c bias windings and d-c feedback windings in such a way that positive and negative feedback is obtained. (*G. Barth, United States patent* 2,247,983, *Figure* 6, 1938.)

Figure 9.6 illustrates the possibility of designing an a-c-controlled push-pull external-feedback circuit without providing separate control windings on the two saturable-reactor units. This push-pull circuit represents a modification of the basic a-c-controlled external-feedback circuit of Fig. 7.2*a*, as described by G. Barth.[8] In this case, the alternating control voltage E_C, synchronous with power-supply voltage E_P, is applied directly to the a-c terminals of two feedback-rectifier bridge circuits, through two sufficiently high resistances R'_C, R''_C and a potentiometer resistor R_0 for zero adjustment. Thus, the alternating currents I'_L and I'_C of the first unit and the alternating currents I''_L and I''_C of the second unit are superimposed, so that the actual values of the d-c ampere-turns $I'_F N'_F$ and $I''_F N''_F$ are functions of the reversible control voltage E_C (phase displacement between E_C and E_P, either 0 or 180°). This simple arrangement may be useful for special applications.

Of course, another possibility for application of a-c control on external-feedback-type push-pull circuits, as shown in Figs. 9.3 to 9.5, consists in reversing two of the four control winding elements so that a-c control instead of d-c control will be obtained (Fig. 7.2*b*). It is to be noted,

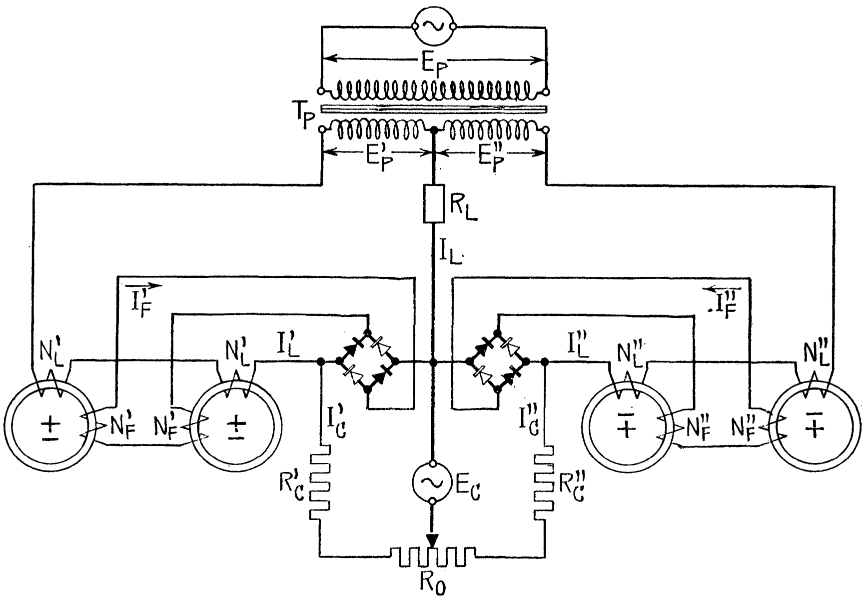

FIG. 9.6. A-c controlled push-pull external-feedback circuit without separate control windings based upon the simple circuit of Fig. 7.2*a*.

however, that the fundamental-frequency voltages induced in the control winding units N'_C and N''_C will cancel each other only for no-signal conditions; otherwise fundamental-frequency voltages having considerable magnitude may be induced in the control circuit. Thus, the actual turns ratio $N'_C/N'_L = N''_C/N''_L$ must have a sufficiently small value (about 0.01 to 0.1). Furthermore, it will be necessary to connect a sufficiently high resistance R_C in series with N'_C and N''_C so that the voltages induced from the load windings N'_L, N''_L into N'_C and N''_C are unable to produce excessively large currents in the control-circuit loop.

9.4. Push-Pull Types of Magnetic Amplifiers with Differential-feedback Circuit. The power amplification of the duodirectional types of nonfeedback magnetic-amplifier circuits (Sec. 5.4), sometimes called "push-pull circuits of the bias excitation type," is comparatively low and not sufficient for most practical purposes. However, it is possible to obtain sufficiently high power amplification by providing true positive feedback, by means of special feedback windings producing additional d-c magnetizations, which are proportional to the actual direct load current $I_L = I'_L - I''_L$. This method, introduced by the author[9] in 1939, may be termed the "differential-feedback method,"

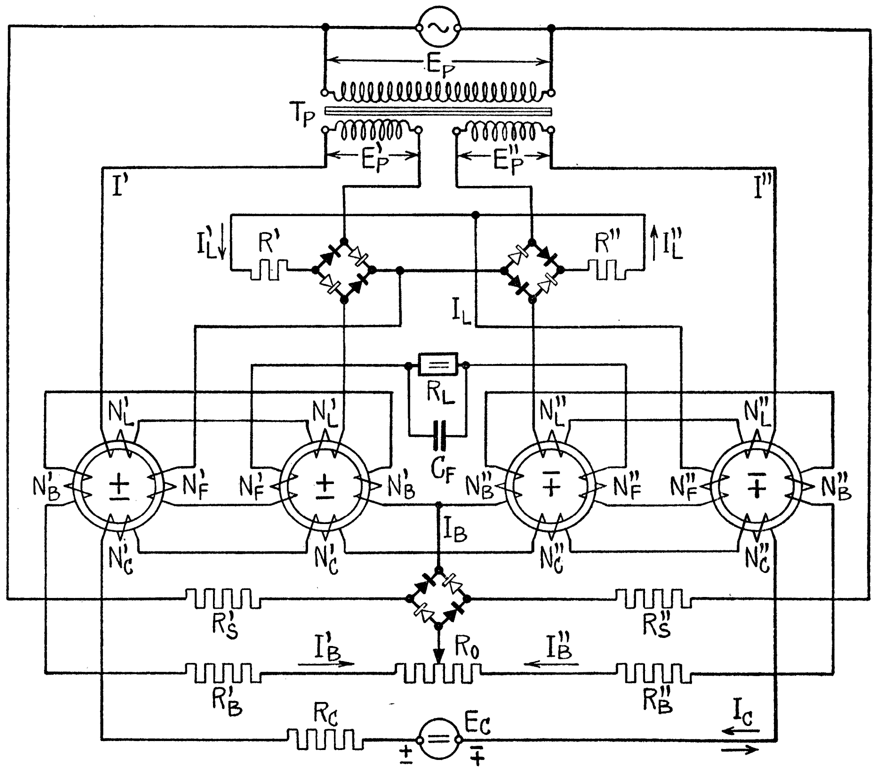

FIG. 9.7. Push-pull external-feedback circuit of the differential-feedback type using "true feedback." (*W. A. Geyger.*[1])

as the feedback effect is proportional to the difference of the two direct output-current components I'_L and I''_L.

In the differential-feedback circuit of Fig. 9.7, the differential output current I_L flowing through the d-c load R_L is fed back through series-connected feedback windings N'_F and N''_F, which aid the control windings N'_C and N''_C. Using this circuit, power amplification of the same order as for the circuits described in Sec. 9.3 can be obtained. Since the additional feedback ampere-turns $I_L N'_F = I_L N''_F$ are proportional to the actual difference of the direct output-current components I'_L and I''_L, these windings may be replaced in special cases (*e.g.*, in a-c load circuits) by two equal feedback windings carrying the output currents I'_L and I''_L separately and acting as differential twin windings (N' and N'' in Fig. 9.8).

In order to reduce the response time of the magnetic amplifier, a derivative-feedback effect may be used. A very efficient method introduced by the author[10] in 1943 consists in providing an additional special feedback circuit containing a large-capacity condenser C_F across the load resistor R_L, as shown in Fig. 9.7. This method is based upon

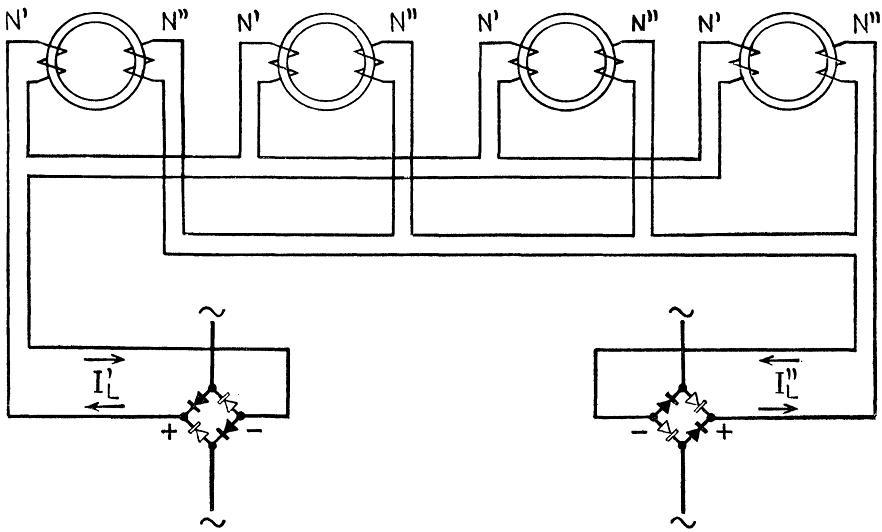

FIG. 9.8. Two equal, separate d-c winding units N' and N'', in a push-pull magnetic amplifier, acting in opposite directions as a differential-type device (twin windings), as used in differential-feedback circuits and in multistage arrangements. (*W. A. Geyger, United States patent* 2,338,423, *Figures* 2 *and* 3, 1938.)

the fact that the actual magnitude of direct feedback current I_L is here a function of time, the shunt condenser C_F acting as a time-variable series impedance on the feedback windings N'_F, N''_F. With rapid changes of I_L, the current flowing through N'_F and N''_F is relatively large, but later on, before steady conditions are reached, this current will slowly decrease to its normal value corresponding to steady conditions. Using this "derivative-feedback method," it is possible to reduce the response time to about one-fifth its original value.

It is to be noted that this circuit utilizes "true positive feedback" and that it operates under compensated-feedback conditions, if $N'_F = 0.5N'_L$ and $N''_F = 0.5N''_L$. Since, with no-signal conditions, there is no d-c component in the external-feedback windings N'_F and N''_F, the four saturable-reactor elements must be biased to the operating point by means of a properly rated d-c bias circuit. In this case, the actual voltage-current characteristics $I_Q = f(E_P)$ will correspond to nonfeedback conditions, so that excellent stability and an extremely small zero drift will be obtained.

9.5. Actual Performance Characteristics of a Push-Pull Type of Magnetic Amplifier with Differential-feedback Circuit. In 1943, the author[11] developed a high-sensitivity magnetic amplifier of the balance-detector type, especially for the purpose of an automatic temperature control with thermocouples. This amplifier with differential-feedback circuit (Fig. 9.7), having one stage of amplification, which operates from a 50-cps power supply, has the following characteristics:

1. Resistance of control windings: $R_C = 15$ ohms
2. Magnitude of control current corresponding to a change in temperature of 1°C: $I_C = 1.13\ \mu a$
3. Magnitude of input voltage corresponding to a change in temperature of 1°C: $E_C = I_C R_C = 17\ \mu v$
4. Magnitude of input power corresponding to a change in temperature of 1°C: $P_C = I_C^2 R_C = 1.9 \times 10^{-11}$ watt
5. Total ampere-turns of control windings ($N_C = 300$) corresponding to a change in temperature of 1°C: $I_C N_C = 3.4 \times 10^{-4}$
6. Ampere-turns per centimeter of control windings (mean length of the magnetic circuit of the Mumetal-core type = 17 cm) corresponding to a change in temperature of 1°C: $AT_C/\text{cm} = 2 \times 10^{-5}$
7. Magnitude of d-c output power corresponding to a change in temperature of 1°C: $P_L = I_L^2 R_L = 10^{-7}$ watt
8. Power gain corresponding to a change in temperature of 1°C: $K_P = P_L/P_C = 10^{-7}\ \text{watt}/(1.9 \times 10^{-11}\ \text{watt}) = 5{,}300$
9. Magnitude of control current corresponding to a change in temperature of 100°C: $I_C = 0.113$ ma
10. Magnitude of input voltage corresponding to a change in temperature of 100°C: $E_C = I_C R_C = 1.7$ mv
11. Magnitude of input power corresponding to a change in temperature of 100°C: $P_C = I_C^2 R_C = 0.19\ \mu w$
12. Magnitude of d-c output power corresponding to a change in temperature of 100°C: $P_L = I_L^2 R_L = 10^{-3}$ watt
13. Power gain corresponding to a change in temperature of 100°C: $K_P = P_L/P_C = 10^{-3}\ \text{watt}/(1.9 \times 10^{-7}\ \text{watt}) = 5{,}300$
14. Drift rate, due to ordinary changes in power-supply voltage, ambient temperature, etc.: each about ±1°C (being ±1 per cent of the 100°C range over a period of 8 hr)

Referring to the zero drift rate of push-pull-type magnetic-amplifier circuits, particularly those operating with such a very low input-power level, it can be said that in these and similar arrangements the entire design revolves around the essential problem that fulfillment of the balance condition $I_L = 0$ if $I_C = 0$ must be mostly independent of changes in magnitude and frequency of the power-supply voltage, changes in ambient temperature, etc. The magnetic amplifiers which were developed in Germany during the period 1937 to 1945 for various servo applications were designed to operate at ambient temperatures between $+70$ and $-60°C$ for power-supply voltage and frequency ranges of 30 to 40 volts and 400 to 600 cps.

In order to avoid excessive temperature drift, various methods for compensating the effects of changes in ambient temperature on the performance of the dry-disk rectifiers have been suggested. One method[12] consists in providing two separate properly rated temperature-sensitive (copper-wire-wound) shunt resistors across the a-c terminals of the two feedback rectifiers. Another way for compensating the temperature drift of push-pull circuits has been used in the differential-feedback arrangement shown in Fig. 9.7: One of the two d-c bias resistors (R'_B or R''_B), acting like a resistance thermometer, adjusts the bias-circuit operating conditions automatically so that the temperature drift is considerably reduced. In a German design, a small heater winding in the spindle of the selenium-rectifier set was provided to improve performance at very low temperatures. For push-pull circuits with selenium rectifiers, the minimum zero drift, or "noise level," is of the order of 10^{-11} watt when differential-feedback circuitry is used.

9.6. Multistage Push-Pull External-feedback Circuits. Magnetic amplifiers involving two or three stages of the push-pull external-feedback type give very high power-gain values. Using a multistage design having two equal stages, a total power gain

$$K_{\text{total}} = \frac{\text{output power of the output stage}}{\text{input power of the input stage}} \tag{9.4}$$

up to 10^7 may be obtained. The over-all power amplification is the product of the stage amplifications, whereas the total time constant is roughly the sum of the individual time constants. It follows that, as the number of stages is increased, there is an increase of the ratio K_P/T between power gain K_P and time constant T. Therefore, in order to obtain a high value of this ratio, the design procedure in many applications will be to connect in cascade two or three stages, each stage having a comparatively low power amplification and yet using the maximum positive feedback that is consistent with the required stability of the amplifier.

The various possibilities for connecting push-pull-type magnetic amplifiers in cascade may be classified, with regard to the example of a two-stage arrangement, in the following way:

1. Combining the d-c control windings (representing a "d-c load") of the second stage with the differential d-c output circuit (Sec. 9.2) of the first stage, using:
   a. A series mixing circuit (Fig. 9.2*a*)
   b. A parallel mixing circuit (Fig. 9.2*b*), or preferably
   c. A twin-load circuit (Fig. 9.2*c*), as shown in Fig. 9.8, N' and N'' representing the twin-type d-c control windings of the second stage[13]
2. Combining the a-c control windings (representing an "a-c load") of the second stage with the a-c output circuit (R_L in Figs. 9.3, 9.4, and 9.6) of the first stage

It is to be noted, however, that in the second case the turns ratio $N'_C/N'_L = N''_C/N''_L$ must have a sufficiently small value to avoid dis-

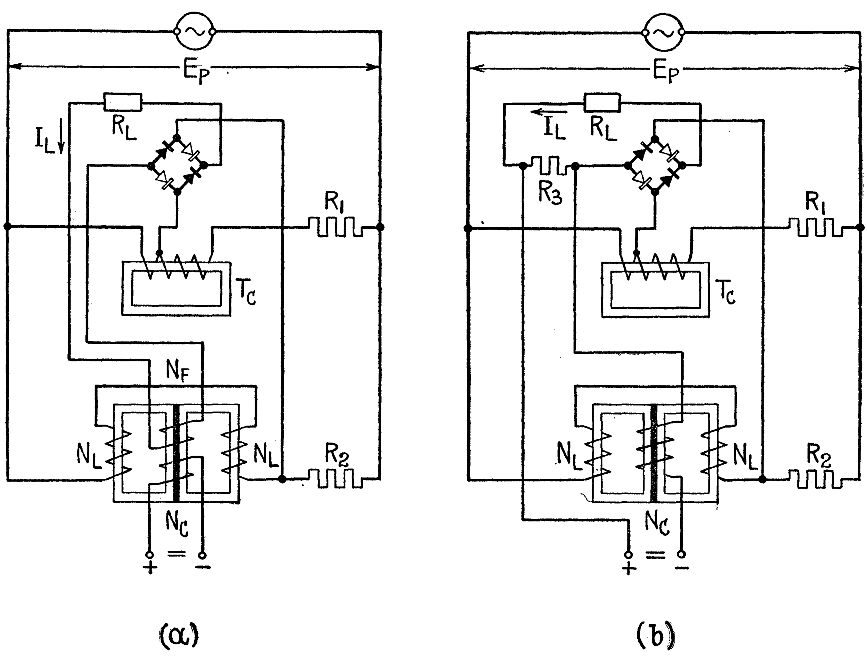

FIG. 9.9. Basic external-feedback circuits using "true feedback." (*A. S. FitzGerald.*[14]) (*a*) Feedback is provided by means of a d-c feedback winding N_F. (*b*) Feedback is provided by an additional resistor R_3 connected in series with both input and output.

turbing effects of the fundamental-frequency voltages which are induced in the a-c control windings N'_C and N''_C of the second stage.

9.7. Other External-feedback Types of Magnetic-amplifier Circuits Using True Feedback. The statement that a magnetic-amplifier circuit uses "true feedback" does not necessarily mean that this amplifier has differential feedback windings. Figures 9.9 and 9.10 illustrate the possibility of applying true feedback by means of an asymmetrical balanced Wheatstone-bridge network, similar to that shown in Fig. 5.7*b*. Such arrangements were devised in 1946 by A. S. FitzGerald.[14]

The basic external-feedback circuits of this type (Fig. 9.9) comprise, in addition to the single saturable-reactor unit with the usual a-c and

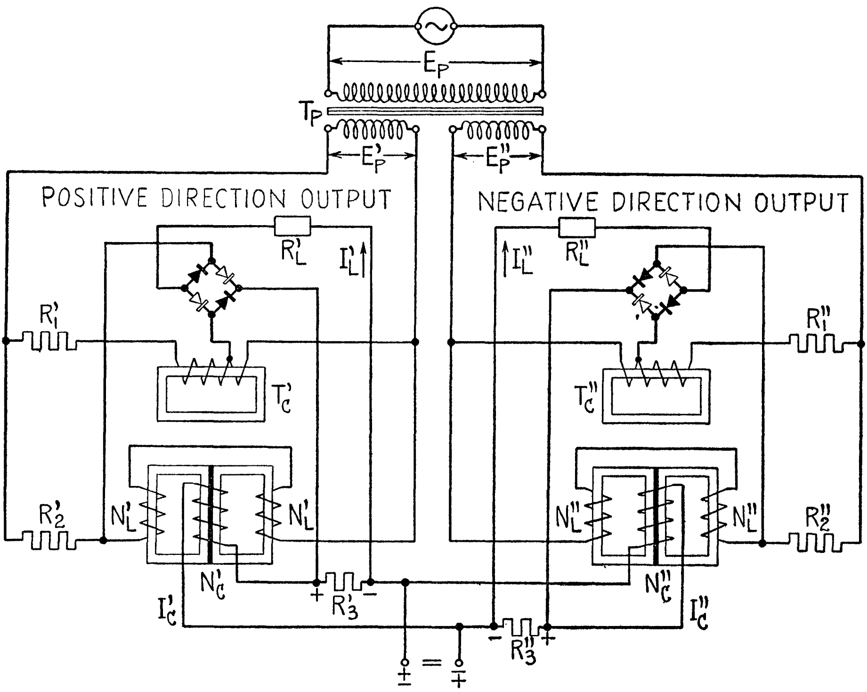

Fig. 9.10. Push-pull feedback arrangement consisting of two separate bridge circuits, as shown in Fig. 9.9*b*. (*A. S. FitzGerald.*[14])

d-c windings N_L and N_C, a transformer T_C and two fixed resistors R_1 and R_2. In the circuit of Fig. 9.9*a*, true feedback is provided by an external-feedback winding N_F. This arrangement can be applied if the input and output-circuit resistance values are incompatible, or if it is desired to keep the input and output circuits electrically separated or insulated. In the circuit of Fig. 9.9*b*, true feedback is provided by an additional resistor R_3 connected in series with both input and output. This is possible because the bridge network is completely balanced, so that, with no-signal conditions, the output current I_L in the load resistor R_L is zero.

Figure 9.10 illustrates the fundamental principle of a push-pull feedback arrangement consisting of two separate bridge circuits, as shown in Fig. 9.9*b*. The connections to the additional resistors R_3' and R_3'' are made so that the feedback is positive on one side when it is negative on the other, and vice versa. In this way, the combined feedback action tends to produce a greater output current on the first unit and a reduced output current on the second unit, in accordance with the actual polarity of the reversible d-c input current. A. S. FitzGerald[15] has stressed that such arrangements with two separated outputs, one for a positive signal and one for a negative signal, have various advantages for special applications, particularly for controlling large-size reversible motors (Sec. 5.5).

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10. Geyger, *op. cit.* (reference 2), pp. 50–53, Fig. 3.
11. See W. A. Geyger, "Magnetic Amplifiers of the Balance Detector Type—Their Basic Principles, Characteristics, and Applications," *AIEE Transactions,* Vol. 70, Part II, pp. 1707–1720, Figs. 32, 33, 1951. See also reference 2.
12. See W. A. Geyger, "Experimentelle Untersuchungen an magnetischen Verstärkern für die Mess- und Regeltechnik" (Experimental investigations on magnetic amplifiers for measurement and control purposes), *Wissenschaftliche Veröffentlichungen aus den Siemens-Werken,* Vol. 20, pp. 43–46, Sec. 5 (The Influence of Changes in Ambient Temperature), 1942.

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CHAPTER 10

Basic Single-core Internal-feedback Magnetic-amplifier Circuits

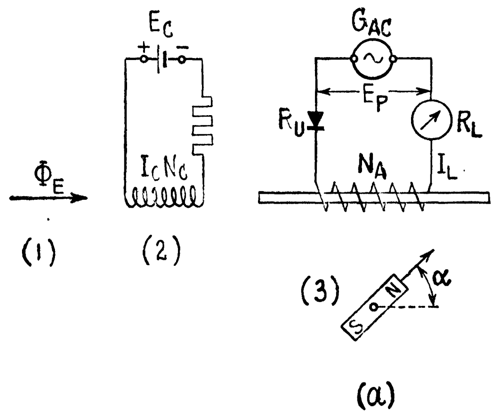
10.1. Introduction. It has already been stressed (Sec. 6.1) that the fundamental principle of regenerative-feedback magnetic-amplifier circuits consists in combining a portion of the output power with the input power in such a way that the actual power gain *with* regeneration is substantially larger than the original power gain *without* regeneration. It has also been mentioned that there are different kinds of regenerative-feedback circuits utilizing either "external" feedback, by means of auxiliary windings (Chaps. 6 to 9), or "internal" feedback, without additional windings ("self-saturation"). Of course, it is possible to combine these two types of feedback circuits to obtain special performance characteristics. Such arrangements, which are based upon the superposition of internal- and external-feedback effects, may be termed "compound feedback circuits."

This chapter describes the basic single-core internal-feedback circuits, which represent fundamental units of various types of more complex internal-feedback circuits (Chaps. 11 to 13) containing two, four, or eight saturable-reactor elements in a single-stage or two-stage arrangement.

10.2. The Fundamental Types of Internal-feedback Circuits. Figure 10.1*a* illustrates the fundamental principle of a simple internal-feedback circuit with a-c generator G_{ac}. This circuit is characterized by the fact that a unidirectional conductor R_U (*e.g.*, a half-wave rectifier), acting as an electric valve, is connected in series with the load winding N_A of a saturable-reactor element so that pulsating, unidirectional load current I_L flows through this winding and the load R_L. Evidently, in this case, the load current I_L causes pulsating, unidirectional magnetic flux in the core of the saturable-reactor element, which flux has a d-c component and an a-c component. Thus, the magnetic flux in the core never falls below a certain minimum value, which depends (1) on the maximum flux density and magnetizing force achieved during the forward alternation of power-supply voltage E_P and positive pulsation of load current

I_L and (2) on the actual value of the magnetomotive force produced by the control means.

Figure 10.1*b* illustrates the possibility of using a d-c generator G_{dc} in connection with a periodically operated contact device S_f, acting as a



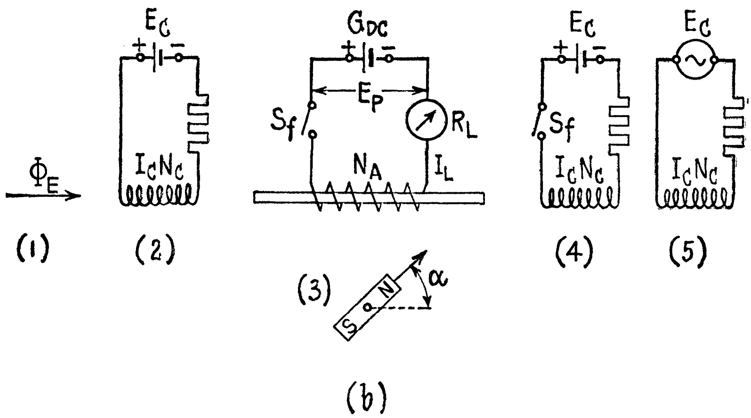


FIG. 10.1. The fundamental types of internal-feedback circuits: (*a*) circuit with a-c generator G_{ac} and half-wave rectifier R_U acting as an electric valve; (*b*) circuit with d-c generator G_{dc} and a periodically operated contact device S_f acting as a chopper or modulator.

chopper or modulator, to obtain similar effects. Such an arrangement may be supplied from a battery, and a mechanically controlled switch with rotating or vibrating contacts may be provided for this purpose.

Figure 10.1 indicates that, in both cases, the "self-saturating circuit" is susceptible to control by any of several means, preferably one of the following:

1. An external magnetic field Φ_E, *e.g.*, the earth's field (flux-gate magnetometer)
2. The magnetic field of a control winding N_C carrying direct current I_C (d-c-controlled magnetic amplifier)
3. An external magnetic field of a permanent magnet having a variable angular position (α) in respect to the saturable-reactor element (telemeter arrangement)
4. The magnetic field of a control winding N_C carrying pulsating, unidirectional current I_C, *e.g.*, using a synchronously operated switch S_f (magnetic amplifier controlled by means of half-cycle pulses)
5. The magnetic field of a control winding N_C carrying a synchronous alternating current I_C, *e.g.*, corresponding to the error-signal voltage of a synchro control transformer (a-c-controlled magnetic servo amplifier)

In both cases, Fig. 10.1*a* and *b*, the saturable-reactor element may have a "closed" magnetic circuit (magnetic amplifier) or an "open" magnetic circuit, as shown in the example of Fig. 10.1 (flux-gate magnetometer).

10.3. Definition of Elementary Operation of the Basic Single-core Self-saturating Circuit with A-C Power Supply. The single-core self-saturating circuit with a-c power supply and "closed" magnetic circuit of high-permeability core material (Fig. 10.2) will serve as a demonstration of the elementary operation of the internal-feedback types of magnetic-amplifier circuits (F. G. Logan[1]). Evidently, in this "half-wave circuit" the power-supply voltage E_P is applied to the load winding N_A only during positive half cycles; during negative half cycles the value of initial flux density (Sec. 4.4) is adjusted by means of the control current I_C, which may be a pure direct current, a pulsating, unidirectional current, or a synchronous alternating current. Thus, during positive half cycles, the core is caused to saturate and allow load current I_L to flow, and, during negative half cycles, the core is caused to reset by an amount dependent upon the actual magnitude of control current I_C. Figure 10.2*b* illustrates this fact by means of corresponding graphical symbols: The half-wave pulses of load current I_L will flow during the "conducting" portion of the cycle ("black" half-wave rectifier), and the adjustment of the value of initial flux density will occur during the "nonconducting" portion of the cycle ("white" half-wave rectifier), *i.e.*, during the reverse alternation of power-supply voltage E_P.

In this circuit I_L is blocked by the rectifier for one half cycle. This means no flux is produced in the core in this half cycle. Therefore, the resulting sequence of flux changes in the core is the same as that with "compensated" external feedback, where the feedback current I_F

eliminates the flux in one half cycle by opposing the load current I_L. This *inherent* feedback effect characterizes the fundamental principle of self-saturating circuits having no additional feedback windings.

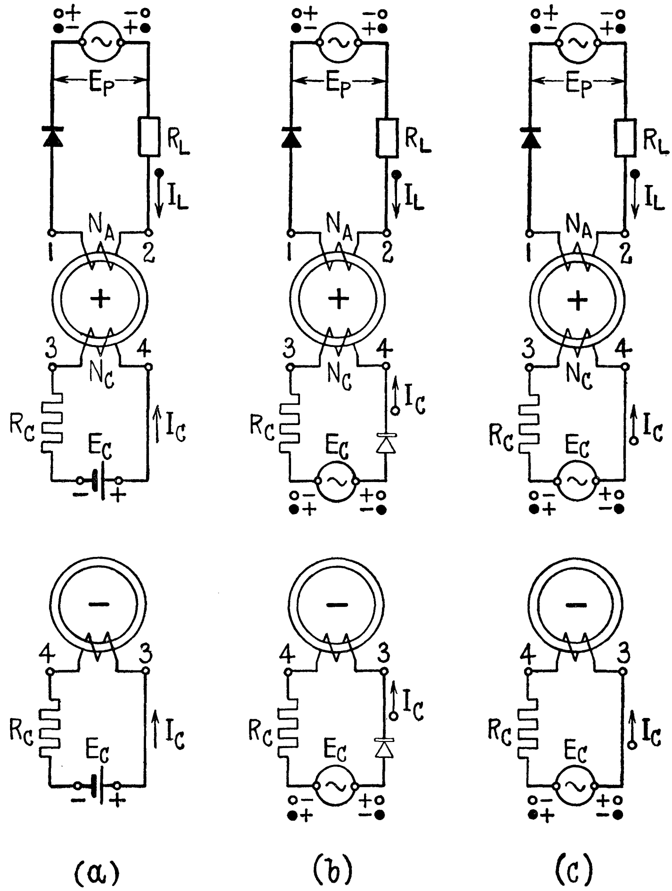


FIG. 10.2. The single-core self-saturating circuit with a-c power supply (E_P) and a "closed" magnetic circuit of high-permeability core material. This circuit may be operated by means of (*a*) a pure direct control current I_C, supplied from a battery; (*b*) a pulsating, unidirectional control current I_C, derived from a half-wave-rectifier circuit (synchronous half-wave pulses); (*c*) a synchronous alternating control current I_C, adjustable either in amplitude or in phase (*e.g.*, by means of a phase-shifting network).

Application of such "internal-feedback" circuits gives the possibility of obtaining very high values of power gain (up to about 10^6 to 10^7).

The load winding N_A itself produces a d-c magnetization, which is proportional to the average value of load current I_L. It is to be noted that the direction of the d-c component of I_L and of the corresponding additional d-c magnetization is always the same and especially is not dependent upon the direction of the reversible control current $\pm I_C$. However, the resultant control action will be greatly influenced by changes in direction of the control current I_C, the additional control action of the d-c component of I_L being either aiding or opposing to the control action of I_C and consequently producing either an increase or a decrease of load current I_L flowing through the load resistor R_L. Therefore, such an arrangement represents a polarized magnetic-amplifier circuit, since its operation will be influenced by changes in direction of the control current I_C. Figure 10.2 illustrates this fact by means of corresponding graphical symbols: When the control action of the load-circuit ampere-turns I_LN_A is aiding that of the control-circuit ampere-turns I_CN_C, then the actual impedance of the load circuit will decrease and the magnitude of load current I_L and load voltage I_LR_L will correspondingly increase (+). However, when the control action of I_LN_A is opposing to that of I_CN_C, then the actual impedance of the load circuit will increase and the magnitude of load current I_L and load voltage I_LR_L will correspondingly decrease (−). In Fig. 10.2, reversing of control current I_C has been indicated by reversing the terminals 3 and 4 of control winding N_C, while the polarity of the terminals 1 and 2 of load winding N_A remains unchanged in all cases.

Referring to the special operating conditions of the single-core self-saturating circuit (Fig. 10.2), it is to be noted that fundamental-frequency voltages having considerable magnitudes may be induced by transformer action into the control circuit. Thus, the actual turns ratio N_C/N_A must have a sufficiently small value (about 0.01 to 0.1). Furthermore, it will be necessary to connect a sufficiently high resistance R_C in series with N_C so that the voltages induced from the load winding N_A into N_C are unable to produce excessively large currents in the control-circuit loop.

Comparison between magnetic amplifiers utilizing "external" and "internal" feedback shows that the self-saturating circuit is equivalent to an external-feedback circuit working under "compensated-feedback" conditions. When the control current I_C is zero, then the magnetizing current produces a large d-c flux in the core, which results in the load current I_L having a correspondingly large quiescent-current value I_Q. Therefore, the resultant $I_LN_A = f(I_CN_C)$ characteristic of the self-saturating circuit is similar to that of an equivalent external-feedback circuit working under compensated-feedback conditions. According to Fig. 6.2, with $I_C = 0$, the load current I_L has a certain value I_Q, which is dependent upon the size of the core, magnetic properties of the core material, magnitude and frequency of the applied power-supply voltage E_P, and departure of the half-wave rectifier from the ideal characteristic. Obviously, it is possible to reduce I_Q to any desired value by means of a bias circuit producing an additional d-c magnetization, which is opposed to the d-c magnetization produced by the d-c component of the load current I_L and is independent of the control magnetization. Alternatively, an a-c bias circuit may be provided for this purpose.

The upper part of Fig. 10.3, showing the sinusoidal power-supply voltage E_P and the actual load voltage I_LR_L as functions of time, indicates

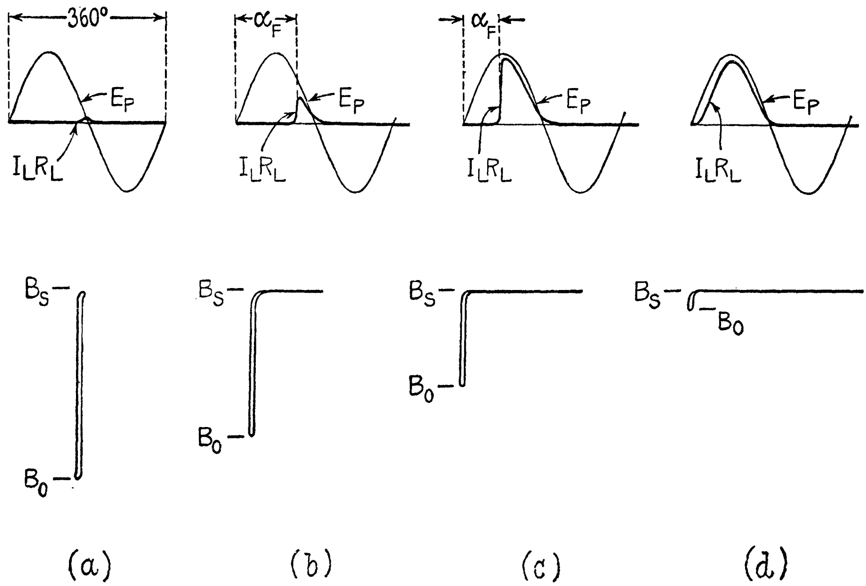


FIG. 10.3. Diagrams illustrating the elementary operation of the basic single-core self-saturating circuit with high-permeability core material (*a*) with minimum value of control current I_C; (*b*) with low value of I_C; (*c*) with medium value of I_C; (*d*) with high value of I_C. The upper part shows the sinusoidal power-supply voltage E_P and the actual load voltage I_LR_L as functions of time; the lower part shows the corresponding dynamic minor hysteresis loops of the saturable-reactor element. (*F. G. Logan.*[2])

the operating conditions of a properly biased single-core self-saturating circuit with various values of control current I_C. This example indicates the manner in which the actual magnitude of the half-wave pulses of load current I_L and the voltage drop I_LR_L can be smoothly adjusted by means of changes of control current I_C.

Corresponding magnetic conditions of the core are shown in the lower part of Fig. 10.3, which gives the dynamic minor hysteresis loops associated with the four cases *a* to *d*. The effect of control current I_C on the minimum or initial flux density B_0 is indicated. It is to be noted that the maximum flux density at the minimum-current or cutoff condition is so adjusted that the hysteresis loop is nearly symmetrical and shows a total excursion of flux density of just less than twice the saturation flux density B_S of the magnetic material. F. G. Logan[2] has stressed that this is one of the more important design features of such circuits, especially where it is desired to secure the optimum range of output for a given reactor structure and control signal.

Referring to Fig. 10.3, it is evident that the actual output-voltage waves (I_LR_L) are similar to those of Fig. 4.1 corresponding to a saturable reactor with high-permeability core material working under nonfeedback conditions. These voltage waves, resembling the output of a gas-filled grid-controlled electronic tube, contain a portion with very great slope, corresponding to "firing" of the saturable-reactor element. The time at which firing occurs depends on the actual magnetomotive force of the control ampere-turns I_CN_C. Thus, actual magnitude of the "firing angle" α_F depends on the initial magnetic flux density B_0 associated with this magnetomotive force: For *increased* values of B_0, firing occurs at *earlier* instants, *i.e.*, *smaller* firing angles α_F (Sec. 4.4). Before firing (during the "cutoff" period), there is substantially no voltage drop I_LR_L across the load resistance R_L, the whole of the applied voltage E_P being sustained by the load windings N_A. After firing, a large part of E_P appears across the load R_L, the maximum value of load voltage I_LR_L being dependent on the actual values of the ohmic resistances in the load circuit.

In 1949, W. J. Dornhoefer[3] described self-saturation in magnetic-amplifier circuits and proposed a simplified but approximate method of calculating the performance of this type of circuit. This treatise includes considerations on the effect of control premagnetization and a method of calculation giving the approximate firing angle and output voltage. It is shown there (1) "that, in a typical case, the prefiring skirt of the voltage wave represents about 3 per cent of the voltage-time integral causing flux-linkage changes in the reactor," and (2) that, "consequently, an assumption that the voltage-time integral absorbed by the reactor is given by the integral of the supply voltage from zero to the firing instant is a relatively good one." This assumption simplifies the analysis greatly. Furthermore, transfer characteristics and oscillograms taken in such a circuit are compared with calculated approximate curves for the same conditions. The effects of the distributed magnetic circuit and of eddy-current shielding are given for a particular case. Finally, salient features of the external-feedback method are compared with self-saturation, and various methods of control of these circuits are discussed.

10.4. Development of Single-core Self-saturating Types of Magnetic-amplifier Circuits. It has already been stressed that the half-wave self-saturating circuits, as shown in Figs. 10.1 and 10.2, are susceptible to control by any of several means. The fundamental principle of operation consists in the adjustment of the value of initial flux density B_0 during the nonconducting portion of the cycle, *i.e.*, during the reverse alternation of the power-supply voltage E_P. Besides the most obvious method, consisting in the application of variable control currents I_C, control may also be achieved by changing the reverse current in the load winding N_A during the nonconducting portion of the cycle by means

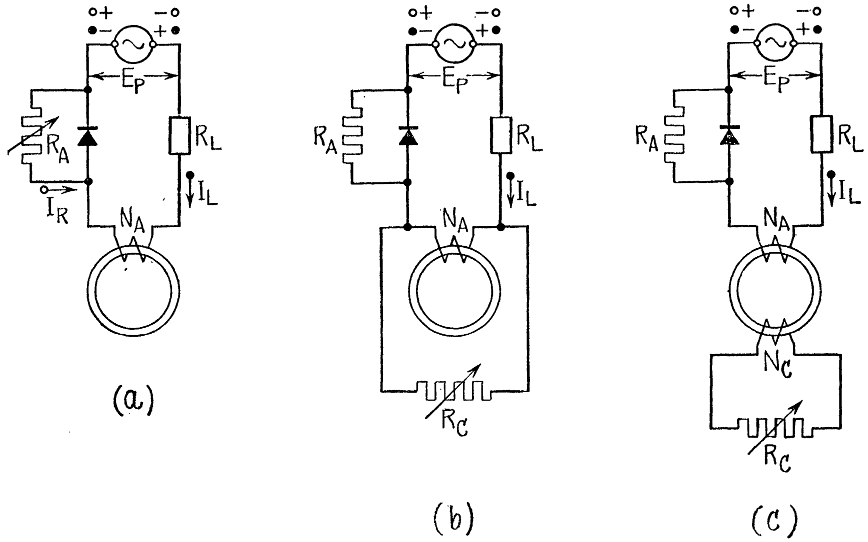


FIG. 10.4. Impedance-controlled single-core self-saturating circuits: (*a*) changing the reverse current I_R in the load winding N_A during the nonconducting half cycle by means of a variable shunt resistor R_A across the half-wave rectifier; (*b*) impedance control by means of a variable shunt resistor R_C across the load winding N_A; (*c*) impedance control by means of a variable resistor R_C across an additional winding N_C of the saturable-reactor element.

of adjustable impedance elements (*e.g.*, variable resistors, condensers, or reactors) across the terminals of the half-wave rectifier, across the load winding N_A, or across an additional winding of the saturable-reactor element. In any case, the elementary operation of such an arrangement is that of a switch, making at a controllable instant of the supply-voltage wave and breaking at or near the zero passage of the current wave. This performance is achieved by utilizing the saturable-reactor unit as a circuit element of alternately very high and very low impedance, an effect facilitated by the existence of a point of rapid change in slope of the magnetizing characteristic, particularly with rectangular-hysteresis-loop core material.

a. Impedance-controlled Single-core Self-saturating Circuits. Figure 10.4*a* illustrates the possibility of changing the reverse current I_R in the load winding N_A during the nonconducting portion of the cycle by means of a variable shunt resistor R_A across the terminals of the half-wave rectifier. In this case, adjustment of the value of initial flux density B_0 during the nonconducting portion of the cycle is achieved by application of the variable reverse current I_R, which is derived during the reverse alternation from the power-supply voltage E_P itself; and the load winding N_A carrying this variable control current I_R acts simultaneously as a control winding.

The basic circuit of Fig. 10.4*a* makes it evident that the characteristics of rectifiers employed for producing regenerative-feedback effects in magnetic amplifiers are quite important, if not critical. In self-saturating circuits, limited reverse resistance (R_A) of the half-wave rectifiers causes a certain back current (I_R) and a corresponding demagnetization of the core during the nonconducting half cycle, with consequent reduction of the internal-feedback effect.

Figure 10.4*b* illustrates a single-core self-saturating circuit, impedance-controlled by means of a variable resistor R_C across the terminals of load winding N_A. Figure 10.4*c* shows a similar arrangement with a variable resistor R_C across an additional winding N_C of the saturable-reactor element. In both cases, a fixed shunt resistor R_A across the rectifier may be provided to adjust the actual magnitude of the internal-feedback effect.

b. Biasing of Single-core Self-saturating Circuits. Obviously, it is possible to reduce the quiescent-current value I_Q of single-core self-saturating circuits, as shown in Figs. 10.1, 10.2, and 10.4, by providing d-c or a-c bias. Application of voltage-sensitive bias circuitry (Sec. 7.6) offers the important advantage of obtaining inherent constant-current characteristics, so that I_Q will be practically independent of changes in power-supply voltage E_P within the limits from ± 5 to 10 per cent.

Figure 10.5*a*, corresponding to Fig. 10.2*b*, shows a single-core self-saturating circuit with bias winding N_B carrying the bias current

$$I_B = \text{const} \times E_P$$

during the reverse alternation of E_P. The bias resistor R_B can be adjusted so that I_Q has the desired value. The separate control winding N_C may be supplied with a variable current I_C, as illustrated in Fig. 10.2*a* to *c*. Of course, a single winding acting simultaneously as bias and control winding may be used in some cases.

Application of a-c bias makes it possible to eliminate the additional bias rectifier used in the circuit of Fig. 10.5*a*. Figure 10.5*b* shows a single-core self-saturating circuit with a separate bias winding N_B carrying the alternating bias current $I_B = \text{const} \times E_P$. This current produces an additional a-c flux, which is superimposed on the a-c flux produced by the a-c component of the half-wave load current I_L. Two resistors R'_B, R''_B and a properly rated shunt condenser C_B are provided (Sec. 7.5) to obtain optimum value of phase displacement between I_B and E_P. In this way, the quiescent current I_Q can be reduced to the desired value. The separate control winding N_C may be supplied with a variable current I_C, as illustrated in Fig. 10.2*a* to *c*. Of course, in this

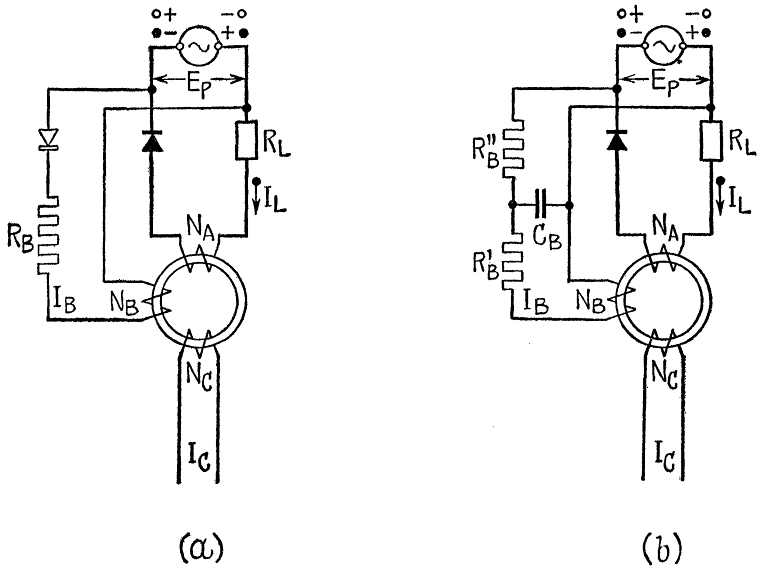


FIG. 10.5. Biasing of single-core self-saturating circuits (*a*) by means of an additional winding N_B carrying the half-wave bias current I_B = const $\times$ E_P during the reverse alternation of E_P; (*b*) by means of an additional winding N_B carrying the alternating bias current I_B = const $\times$ E_P. Two resistors R'_B, R''_B and a properly rated shunt condenser C_B are provided to obtain the optimum value of phase displacement between I_B and E_P.

case too, a single winding acting simultaneously as bias and control winding may be used.

c. Single-core Self-saturating Circuit with Auxiliary Transformer Acting as a Magnetizing Element. In 1951, R. A. Ramey[4] described a special single-core self-saturating circuit (Fig. 10.6) with an auxiliary transformer ("magnetizing element") T_M, which derives a properly rated synchronous voltage E_M from the power-supply voltage E_P and applies it to the control circuit, which contains an additional half-wave rectifier. This rectifier is placed in series opposition to the (pure or pulsating) d-c control voltage E_C, which is in series with the so-called magnetizing voltage E_M.

Figure 10.6 illustrates the possibility of using either (*a*) a single winding N_A, acting simultaneously as load and control winding, or (*b*) two separated windings N_A and N_C. In any case, the auxiliary transformer T_M is rated so that its actual voltage ratio E_M/E_P is exactly equal to the actual turns ratio N_C/N_A of the saturable-reactor element. The purpose of the magnetizing voltage E_M can be defined in the following way: During "positive" half cycles ("black" polarities) the voltage E_M will prevent flow of current in the control circuit due to winding N_C; and during "negative" half cycles ("white" polarities) the voltage E_M must

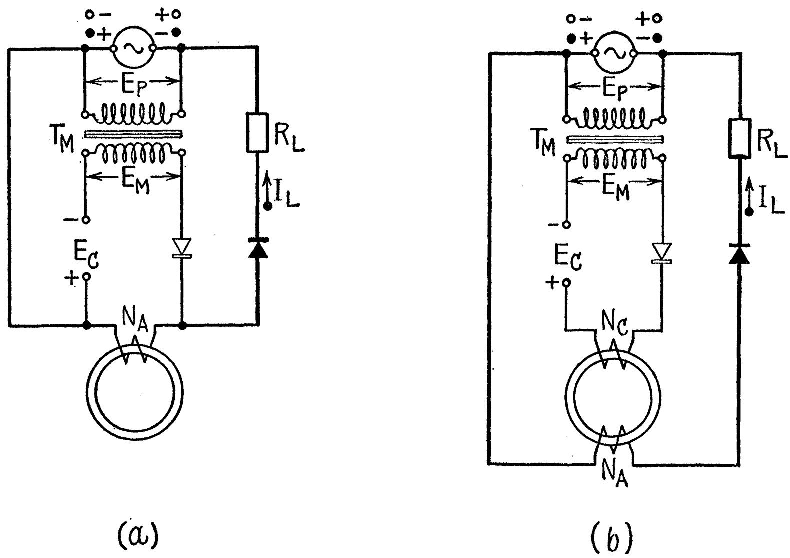


FIG. 10.6. Single-core self-saturating circuit with auxiliary transformer T_M acting as a "magnetizing element" (*R. A. Ramey*[4]): (*a*) using a single winding N_A acting simultaneously as load and control winding; (*b*) using two separated windings N_A and N_C.

accomplish the appropriate magnetization of the core of the saturable-reactor element.

L. J. Johnson[5] described interesting results of various experimental investigations concerning the effect of four different core materials on the circuit of Fig. 10.6*b*. He illustrated the correlation between hysteresis-loop characteristics and transfer characteristics. In these investigations "an ideal transfer characteristic is assumed to be one which exhibits a linear relationship of average load current vs. average control voltage with the load current ranging from zero to some maximum value called saturation. Ideally, independence of supply frequency and voltage, rectifier leakage, and core hysteresis-loop width is implied."[5] Furthermore, "it is assumed that the control voltage is derived from a voltage similar in waveshape and of the same phase as the a-c supply voltage to the control circuit."[5] Therefore, the control voltage E_C is obtained by rectifying the 60-cycle a-c supply through a variable-voltage transformer by a full-wave dry-disk rectifier. By this means the control-voltage waveform (a continuous series of half sinusoids) is made the same as that of the a-c supply (E_M) to the control circuit and 180° out of phase with it during the reset half cycle. Since the actual current

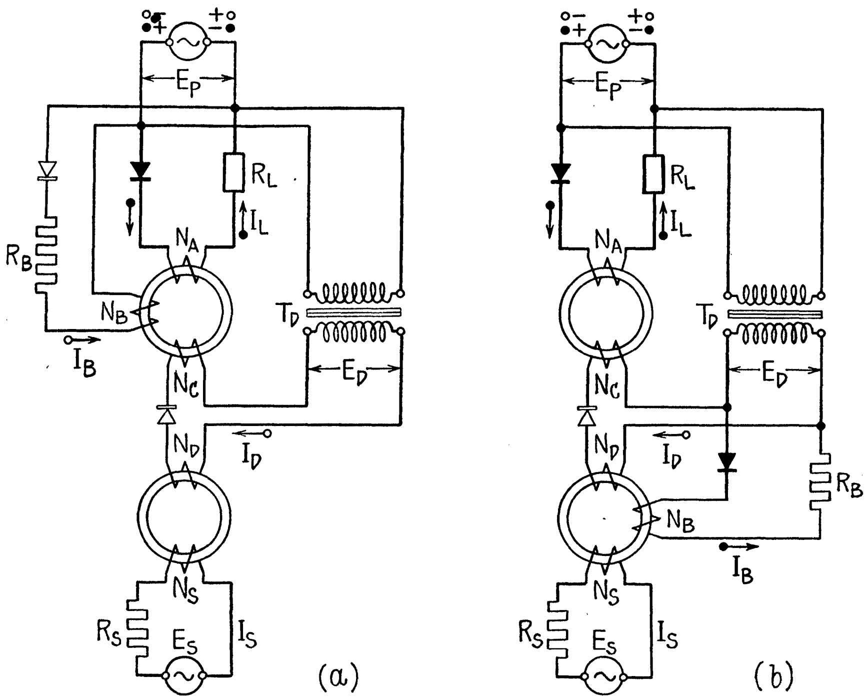


FIG. 10.7. Two-stage arrangements with two single-core self-saturating circuits: (*a*) using additional biasing of the output-stage circuit; (*b*) using additional biasing of the input-stage circuit.

flowing through the control circuit is equal to the magnetizing current minus the load-circuit rectifier leakage current times the turns ratio, "the criterion for selecting an appropriate rectifier to obtain an optimum transfer characteristic is only that the leakage must not exceed the magnetizing current requirement." The basic operating principles of this special single-core circuit are the same as those of the more complex circuits which can be built with it.

d. Two-stage Arrangement with Two Single-core Self-saturating Circuits. Figure 10.7*a* shows such an arrangement, which has the following components: The output stage is a single-core self-saturating circuit containing:

1. The load winding N_A with series-connected half-wave rectifier and load resistor R_L (load current I_L)
2. The bias winding N_B with series-connected half-wave rectifier and bias resistor R_B (bias current I_B)
3. The control winding N_C carrying the output current I_D of the input stage

The input stage is also a single-core self-saturating circuit containing:

4. The load winding N_D with series-connected half-wave rectifier and power-supply transformer T_D (voltage E_D), the control winding N_C of the output stage representing the load of this input-stage circuit
5. The control winding N_S (signal current I_S) with series resistor R_S and signal-voltage source producing the variable alternating signal voltage E_S

Evidently, in this two-stage arrangement, the power-supply voltage E_P is applied to N_A only during "positive" half cycles ("black" polarities); during "negative" half cycles ("white" polarities) the secondary voltage E_D of transformer T_D is applied to N_D, and the value of initial flux density in the output-stage core is adjusted by means of the input-stage load current I_D, which is a function of the signal current

$$I_S = \text{const} \times E_S$$

It is to be noted that the input stage acts simultaneously as a voltage-sensitive bias circuit (Sec. 8.4), so that the quiescent-current value I_Q of the output-stage load current I_L will be considerably reduced. This reduction of I_Q may be excessive. For this reason, the output stage is additionally biased by means of the bias winding N_B carrying the adjustable current I_B, and the series resistor R_B can be adjusted so that I_Q has the desired value.

Figure 10.7*b* shows a similar arrangement employing an additional bias winding N_B on the core of the input stage. Evidently, in this case, the actual quiescent-current value I_Q of the output-stage load current I_L is adjusted by controlling the actual magnitude of the quiescent-current value of the input-stage load current I_D flowing through the control winding N_C of the output stage. In this case, too, the series resistor R_B of the bias winding N_B can be adjusted so that I_Q flowing through the load resistor R_L has the desired magnitude.

10.5. Literature Concerning Theoretical Investigations on the Basic Single-core Self-saturating Circuit. Besides the authors already referred to in this chapter, contributions describing theoretical investigations on the operating conditions of the basic single-core self-saturating circuit have been made by A. U. Lamm,[6] U. H. Krabbe,[7] D. W. VerPlanck and others,[8] W. H. Esselman,[9] E. J. Smith,[10] and H. Lehmann.[11]

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CHAPTER 11

Basic Two-core Internal-feedback Types of Magnetic-amplifier Circuits

11.1. Classification of Internal-feedback Magnetic-amplifier Circuits. The basic single-core self-saturating circuits, as described in the preceding chapter, represent "polarized" magnetic amplifiers, which are able to discriminate between "positive" and "negative" control currents. Referring to the asymmetrical transfer characteristic of Fig. 6.2 of the basic external- and internal-feedback circuits, it is evident that a positive control current $+I_C$ causes an increase of load current I_L, while a negative control current $-I_C$ causes a decrease of I_L. However, I_L will never change its direction.

In order to obtain internal-feedback-type magnetic amplifiers having duodirectional transfer characteristics, as shown in Fig. 5.1*c*, it is necessary to apply certain modifications of the basic self-saturating circuits.

Classification of internal-feedback magnetic-amplifier circuits and of those modified internal-feedback circuits having additional external-feedback windings ("compound" feedback circuits) may be outlined as follows:

Group 1: Simple *polarized* circuits
- *a.* With a-c load
- *b.* With d-c load

Group 2: Duodirectional (push-pull) circuits
- *a.* With a-c load
- *b.* With d-c load

This chapter describes typical examples of the first group of these circuits.

11.2. Basic Two-core Self-saturating Circuits with Pure Internal Feedback. These circuits are characterized by the fact that the feedback effect is inherently produced by self-saturation alone, without any additional feedback windings. However, other kinds of additional windings may be provided for special purposes, *e.g.*, d-c or a-c bias windings (Fig. 10.5), additional d-c or a-c control windings, etc. Furthermore, these circuits may be modified in various ways, *e.g.*, by providing shunt resistors across the two half-wave-rectifier units (Fig. 10.4) or additional impedance elements (*e.g.*, special types of variable resistors, condensers, or reactors) across the already available windings, or across additional windings of the two saturable-reactor elements.

a. The Basic Two-core Internal-feedback Circuit with A-C Load. Figure 11.1 shows the basic self-saturating circuit with a-c load R_L, sometimes called a "doubler" circuit (F. G. Logan,[1] 1936), which may be either

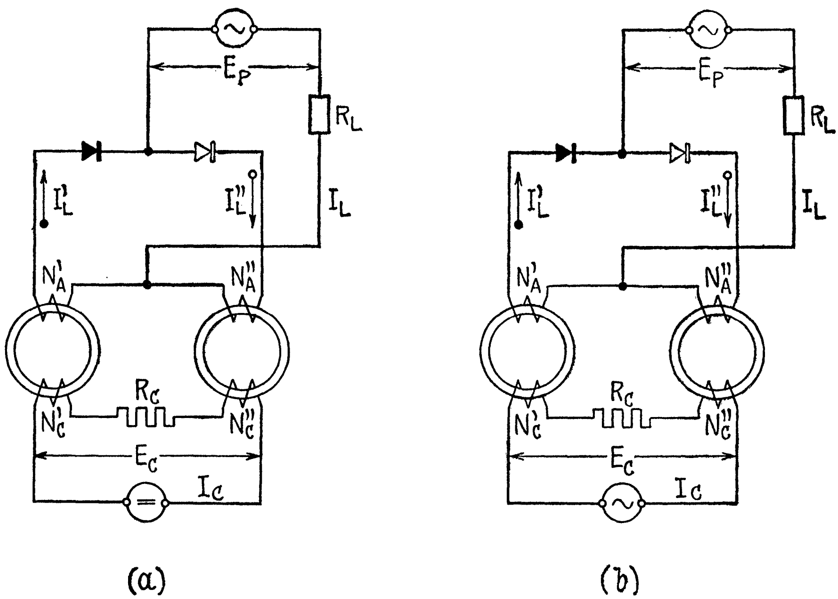


FIG. 11.1. The basic self-saturating circuit with a-c load, or "doubler" circuit (*F. G. Logan.*[1]): (*a*) d-c controlled; (*b*) a-c controlled.

d-c controlled or a-c controlled.[2] The two half-wave load-current components I'_L ("positive" half cycles, indicated by "black" polarities) and I''_L ("negative" half cycles, indicated by "white" polarities) are superimposed in such a way that pure alternating current I_L flows through the load resistor R_L. The actual magnitude of I_L is a function of the variable control current $I_C = E_C/R_C$ in the control windings N'_C and N''_C. Internal feedback is produced by the d-c components of the two pulsating, unidirectional currents I'_L and I''_L flowing through the load windings N'_A and N''_A of the two saturable-reactor elements.

Comparison between magnetic amplifiers utilizing external and internal feedback makes it evident that the internal-feedback circuit of Fig. 11.1 is equivalent to the external-feedback circuit of Fig. 6.1, working under compensated-feedback conditions. Thus, the transfer characteristic representing the load-circuit ampere-turns as a function of the control-circuit ampere-turns (Fig. 6.2) is similar to that of an equivalent external-feedback circuit working under compensated-feedback conditions. With $I_C = 0$, the load current I_L has a certain value I_Q, which is dependent upon the size of the cores, magnetic properties of the core material,

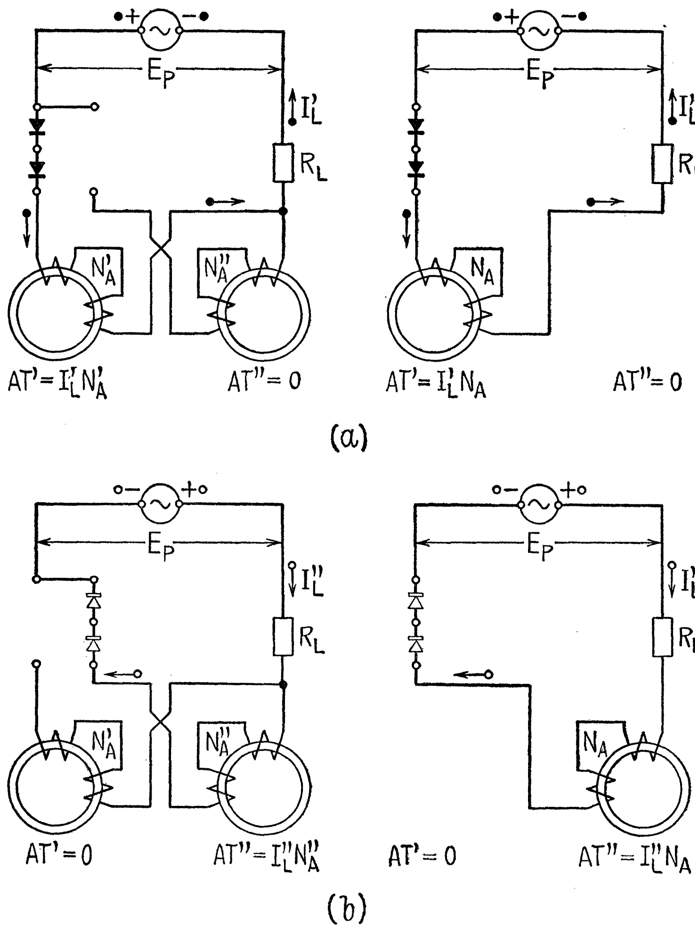


FIG. 11.2. The fundamental principle of the basic self-saturating circuit with a-c load, as shown in Fig. 11.1. (*a*) Operating conditions during the first half-cycle period; (*b*) operating conditions during the second half-cycle period.

magnitude and frequency of the applied power-supply voltage E_P, and departure of the half-wave rectifiers from the ideal characteristic. Of course, it is possible to reduce the quiescent current I_Q to any desired value by means of a bias circuit producing an additional d-c magnetization, which is opposed to the d-c magnetization produced by the d-c components of the two pulsating, unidirectional currents I'_L, I''_L and is independent of the control magnetization. Alternatively, an a-c bias circuit may be used for this purpose.

Figure 11.2, illustrating the fundamental principle of the basic internal-feedback circuit with a-c load (Fig. 11.1), shows that:

a. During the *first* (positive) half-cycle period

$$AT' = I'_L N'_A = I'_L N_A \tag{11.1a}$$
$$AT'' = 0 \tag{11.1b}$$

b. During the *second* (negative) half-cycle period

$$AT' = 0 \tag{11.2a}$$
$$AT'' = I''_L N''_A = I''_L N_A \tag{11.2b}$$

Comparison between Fig. 6.3 (external feedback) and Fig. 11.2 (internal feedback) makes it evident that the total copper resistance of the load windings is

$$R_{\text{total}} = 4R_N \text{ in the external-feedback circuit}$$

and

$$R_{\text{total}} = 2R_N \text{ in the internal-feedback circuit}$$

Evidently, with maximum value of load current I_L, the power dissipation in the load and feedback windings, N_L and N_F, of the external-feedback circuit is twice that in the internal-feedback circuit, assuming that identical magnetic-core structures and the same value of power-supply voltage E_P are provided in both cases. Consequently,[3] for the same temperature rise (corresponding to full-load-current conditions), the output power of the external-feedback circuit must be limited to 50 per cent of that allowable in the internal-feedback circuit using the same weight of copper and magnetic material. Therefore, to reach the same output power with the external-feedback circuit with operation on the same power-supply voltage E_P, the copper section must be approximately doubled. This means (Dornhoefer[3]) that "twice the window area will be required for the circuit using external feedback. If a square window and the same stack is assumed, doubling the winding means an increase of approximately 41 per cent in the mean length of the magnetic path and weight of core material, as well as the 100 per cent increase of copper weight."

b. The Basic Two-core Internal-feedback Circuit with D-C Load. Figure 11.3 shows the basic self-saturating circuit with d-c load R_L, sometimes called a "full-wave" circuit (F. G. Logan,[4] 1931), which may be either d-c controlled or a-c controlled. In this case, a center-tapped power-supply transformer T_P is provided; and the two half-wave load-current components I'_L and I''_L are superimposed in such a way that full-wave rectified direct load current $I_L = I'_L + I''_L$ flows through the load resistor

R_L. The actual magnitude of I_L is a function of the variable control current $I_C = E_C/R_C$ in the control windings N'_C and N''_C. Internal feedback is produced by the d-c components of the two pulsating, unidirectional currents I'_L and I''_L flowing through the load windings N'_A and N''_A of the two saturable-reactor elements. The transfer characteristic of this circuit is similar to that of the basic a-c load circuit, shown in Fig. 11.1.

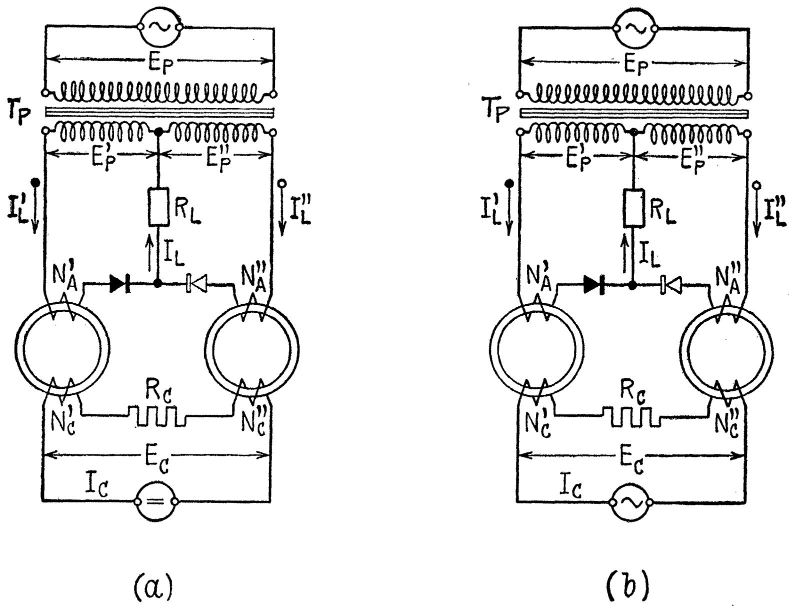
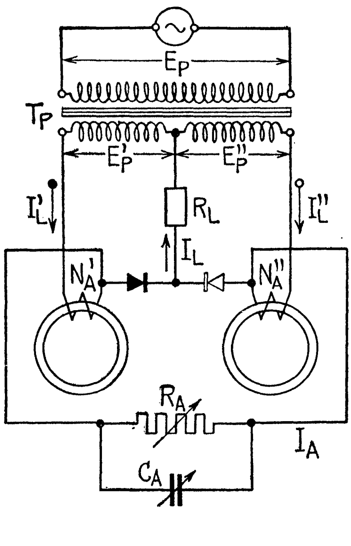


FIG. 11.3. The basic self-saturating circuit with d-c load, or "full-wave" circuit (*F. G. Logan*[4]): (*a*) d-c controlled; (*b*) a-c controlled.

In 1919, J. Jonas[5] devised various saturable-reactor circuits "for regulating the voltage of metal-vapor-rectifier installations" and for supplying controllable d-c power from single-, three-, and six-phase networks. The single-phase circuit, as described by Jonas, employing two saturable-core-reactor elements which operate with self-saturation, is similar to the self-saturating circuit of Fig. 11.3*a*. However, it is to be noted that in Jonas's voltage-regulator circuit the two saturable-reactor elements are additionally excited by the variable direct control current always "in the same direction" as by the load ("anode") current, whereas in the *polarized* full-wave magnetic-amplifier circuit of Fig. 11.3*a*, the d-c control windings are excited by means of a *reversible* direct control current, so that the actual ampere-turns $I'_L N'_A = I''_L N''_A$ and $I_C N'_C = I_C N''_C$ will be either aiding or opposing each other. Thus, in this case, the desired "polarized" transfer characteristic (Fig. 5.1*b*) will be obtained.

F. G. Logan[6,7] has stressed that the circuit of Fig. 11.3 may also be impedance-controlled by means of variable impedance elements (resistors, condensers, or reactors) across various parts of the circuit. Figure 11.4 illustrates this possibility by presenting two typical examples.

c. The Bridge-type Two-core Internal-feedback Circuit with D-C Load.[8] This circuit (Fig. 11.5) makes it possible to eliminate the power-supply



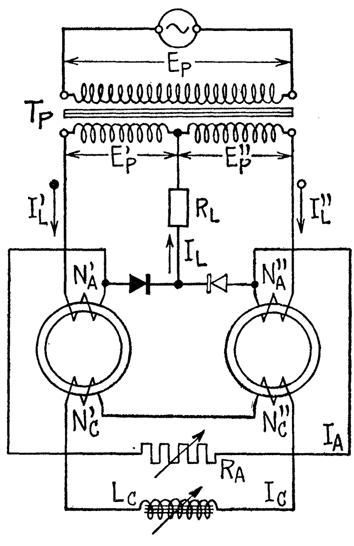


FIG. 11.4. Impedance-controlled self-saturating "full-wave" circuit: (*a*) using a variable shunt resistor R_A and/or a variable shunt condenser C_A across two terminals connecting the rectifier units with the load windings N_A' and N_A''; (*b*) using a variable shunt resistor R_A across these two terminals and/or a variable reactor L_C (*e.g.*, a magnetically controlled saturable-reactor device) across additional series-connected windings N_C' and N_C''.

transformer (T_P in Fig. 11.3) by dividing the load windings N_L', N_L'' into two equally rated parts, which are insulated from each other. It is to be noted that the total number of dry-disk-rectifier cells remains unchanged, so that, for example, 4×12 cells may be used in the bridge circuit (Fig. 11.5), whereas 2×24 cells have to be provided in the transformer circuit (Fig. 11.3). Obviously, the circuit of Fig. 11.5 may be a-c controlled after reversing one of the two control winding elements N_C', N_C''; and a d-c or a-c bias arrangement may be provided for controlling the quiescent-current value I_Q of the direct load current $I_L = I_L' + I_L''$.

11.3. Two-core Self-saturating Circuits with Additional Feedback Windings. These circuits are characterized by the fact that the actual feedback effect is the sum of two components, which may be either

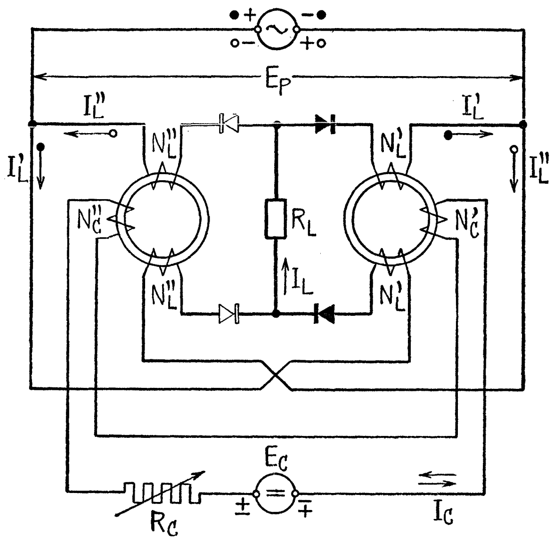


FIG. 11.5. The bridge-type self-saturating circuit with d-c load.

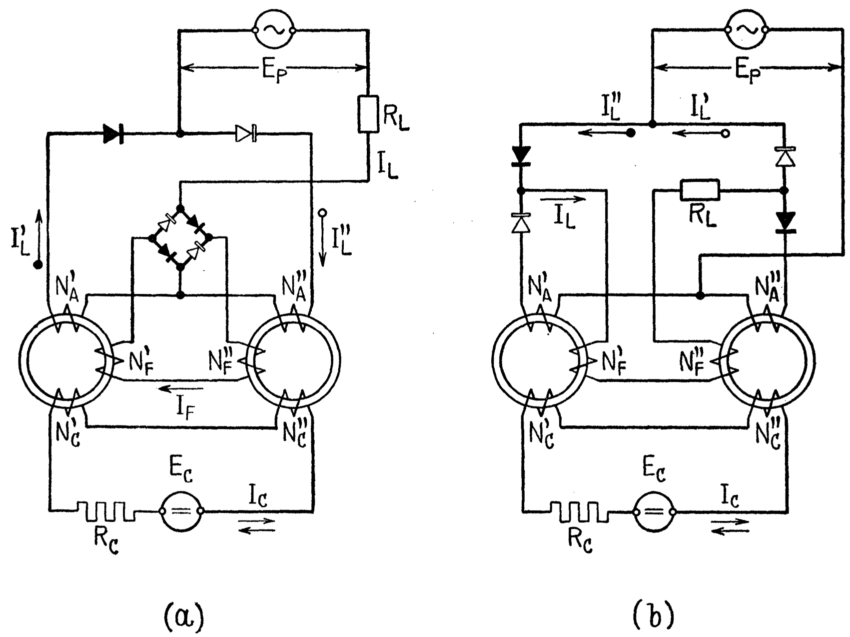
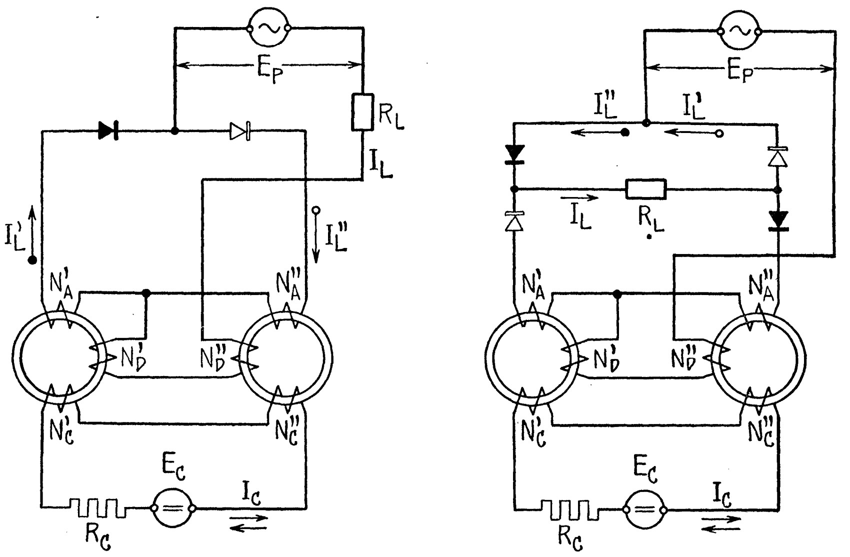


FIG. 11.6. Self-saturating circuits with additional d-c feedback windings N'_F, N''_F (*a*) with a-c load; (*b*) with d-c load.

aiding or opposing. The first component is due to self-saturation, and the second component is produced by means of additional d-c or a-c windings. Figures 11.6 to 11.9 illustrate some typical examples of such "compound" feedback circuits.[9] All these circuits, of course, may be modified in various ways, *e.g.*, by providing shunt resistors across the two half-wave-rectifier units (Fig. 10.4) or additional impedance elements



(a) (b)

FIG. 11.7. Self-saturating circuits with additional a-c series windings N'_D, N''_D (*a*) with a-c load; (*b*) with d-c load.

across the already available windings, or across additional windings of the two saturable-reactor elements.

a. Internal-feedback Circuits with Additional D-C Windings. Figure 11.6 shows (*a*) an a-c load circuit, and (*b*) a d-c load circuit with additional d-c feedback windings N'_F, N''_F. The magnitude of the resultant feedback effect depends upon the number of turns of the load windings N_A and of the additional d-c windings N_F carrying the full-wave rectified load current I_L. In this case, the current gain I_L/I_C is given by the equation

$$\frac{I_L}{I_C} = 2\frac{N_C}{N_A}\frac{1}{1 \mp (N_A - 2N_F)/N_A} \tag{11.3}$$

However, the actual number of turns N_F of the additional d-c feedback windings has no influence upon the magnitude of the power-supply voltage E_P which can be applied to the arrangement.

b. Internal-feedback Circuits with Additional A-C Windings. The current gain I_L/I_C of the circuits of Fig. 11.7 with a-c load or with d-c load is given by the following relationship:

$$\frac{I_L}{I_C} = \frac{2N_C}{N_A + 2N_D} \frac{1}{1 \mp N_A/(N_A + 2N_D)} \tag{11.4}$$

It is to be noted that the actual number of turns N_D of the additional a-c series windings carrying the pure alternating current I_L has a certain

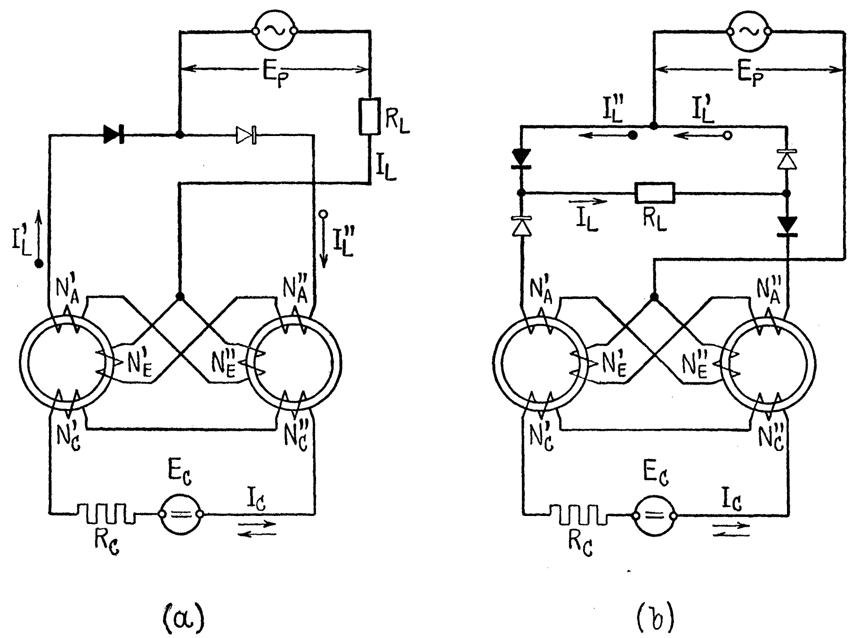


FIG. 11.8. Self-saturating circuits with additional cross-connected a-c windings N'_E, N''_E (*a*) with a-c load; (*b*) with d-c load.

influence upon the magnitude of the power-supply voltage E_P which can be applied to the circuit.

c. Internal-feedback Circuits with Additional Cross-connected A-C Windings. Figure 11.8 shows further modifications of the basic two-core internal-feedback circuits with a-c or d-c load having cross-connected a-c windings N_E. In this case, the current gain is

$$\frac{I_L}{I_C} = \frac{2N_C}{N_A + N_E} \frac{1}{1 \mp (N_A - N_E)/(N_A + N_E)} \tag{11.5}$$

The actual number of turns N_E of the additional windings carrying the half-wave currents I'_L and I''_L has a certain influence upon the magnitude of the power-supply voltage E_P which can be applied to the arrangement.

d. Internal-feedback Circuits with Center-tapped Power-supply Transformer and Additional D-C or A-C Windings. Figure 11.9 illustrates

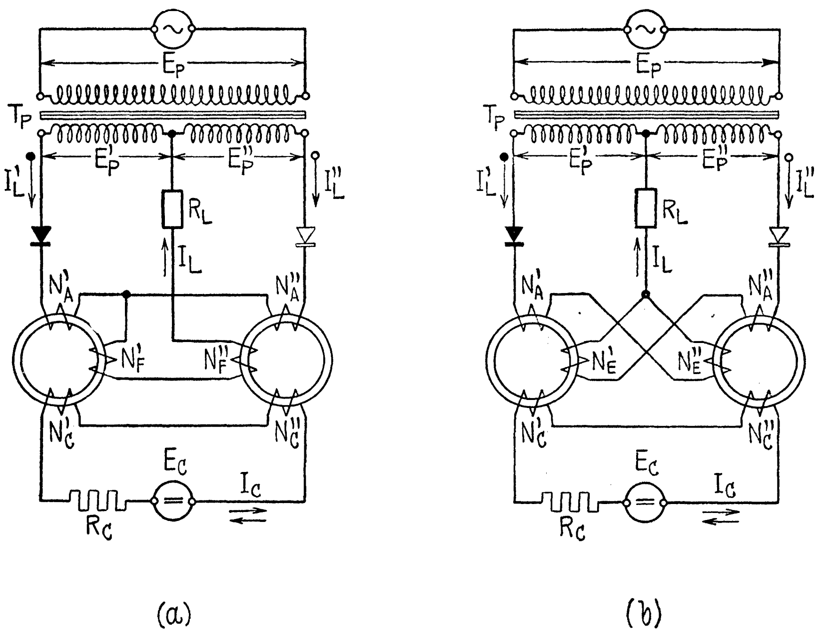


FIG. 11.9. Self-saturating circuits with center-tapped power-supply transformer T_P and additional d-c or a-c windings (*a*) with series-connected d-c feedback windings N_F', N_F''; (*b*) with cross-connected a-c feedback windings N_E', N_E''.

the possibility of applying additional d-c or a-c windings on the basic "full-wave" circuit with d-c load, as shown in Fig. 11.3:

a. The direct load current $I_L = I_L' + I_L''$ flowing through the external-feedback windings N_F produces an additional d-c feedback effect, proportional to the ampere-turns $I_L N_F$; and the actual current gain I_L/I_C is given by Eq. (11.3).

b. The cross-connected feedback windings N_E carrying the half-wave currents I_L' and I_L'' produce an additional feedback effect, proportional to the ampere-turns $I_L'N_E'' = I_L''N_E'$; and the actual current gain I_L/I_C is given by Eq. (11.5).

11.4. Two-core Self-saturating Circuit with Auxiliary Transformer Acting as a Magnetizing Element. In 1951, R. A. Ramey[10] described the full-wave circuit (Fig. 11.10) with an auxiliary transformer ("magnetizing element") T_M, which derives a properly rated synchronous voltage E_M from the power-supply voltage E_P and applies it to the control circuit, which contains an additional bridge-type half-wave-rectifier combination. The two control winding elements N'_C, N''_C are connected to this second rectifier combination just as the load winding elements N'_A, N''_A are connected to the first rectifier combination, with the exception that the d-c load R_L is replaced by the d-c control voltage E_C. Evidently, this two-core arrangement is a further development of the single-core

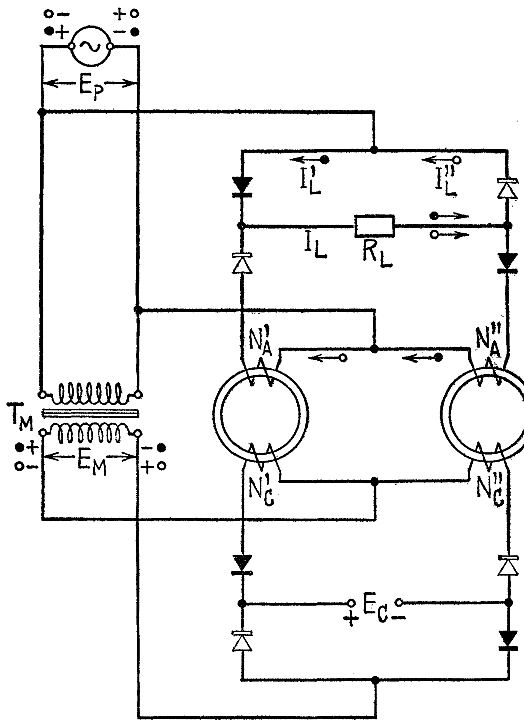


FIG. 11.10. Self-saturating circuit with two saturable-reactor elements and an auxiliary transformer T_M acting as a magnetizing element. (*R. A. Ramey.*[10])

self-saturating circuit with magnetizing element, as described in Sec. 10.4*c* (Fig. 10.6*b*).

In the output circuit the full load current I_L flows in alternate half cycles through the load windings N'_A, N''_A. The existence of the four half-wave rectifiers in the output circuit prevents the application of the power-supply voltage E_P to the saturable-reactor element, which is not to conduct during any particular half cycle. In the input circuit the "magnetizing voltage" E_M is chosen to be of the same phase as the power-supply voltage E_P and of magnitude $E_M = E_P N_C/N_A$. The control voltage E_C is obtained by rectifying E_P through a variable-voltage transformer by a full-wave rectifier. By this means the control-voltage waveform (a continuous series of half sinusoids) is made the same as that of the a-c supply (E_M) to the control circuit and 180° out of phase during the reset half cycle. In this arrangement too (L. J. Johnson[11]) "the criterion for selecting an appropriate rectifier to obtain an optimum transfer characteristic is only that the leakage must not exceed the magnetizing current requirement."

11.5. Literature Concerning Theoretical Investigations on the Basic Two-core Self-saturating Circuits with Additional D-C or A-C Windings. A complete analysis of various two-core self-saturating circuits with additional d-c or a-c feedback windings has been given by U. H. Krabbe[12] and by L. A. Finzi, G. F. Pittman, and H. L. Durand.[13] Further contributions concerning development and special applications of self-saturating circuits with two saturable-reactor elements have been made by M. A. Edwards,[14] R. A. Andrews,[15] A. U. Lamm,[16] S. E. Hedström and A. Söderholm,[17] S. Steinitz,[18] R. E. Morgan,[19] and H. M. Ogle.[20]

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CHAPTER 12

Full-wave Types of Duodirectional Internal-feedback Circuits with Two Saturable-reactor Elements

12.1. Classification of Duodirectional Internal-feedback Circuits. The duodirectional (push-pull) internal-feedback types of magnetic-amplifier circuits may be classified in a manner similar to the duodirectional external-feedback circuits (Sec. 8.1):

Group 1: Push-pull circuits with *two* saturable-reactor elements
- *a.* With a-c load
- *b.* With d-c load

Group 2: Push-pull circuits with *four* saturable-reactor elements
- *a.* With a-c load
- *b.* With d-c load

However, it is to be noted that there are two different kinds of duodirectional internal-feedback circuits operating with two saturable-reactor elements:

1. *Full*-wave push-pull circuits, which are characterized by the fact that *both* half cycles of power-supply voltage E_P are used for supplying the load windings and the a-c or d-c load
2. *Half*-wave push-pull circuits, which are characterized by the fact that only *one* half cycle of E_P is used for supplying the load windings and the a-c or d-c load

This chapter describes the full-wave circuits, while the half-wave types will be treated in the next chapter.

12.2. Push-Pull Circuits with Two Saturable-reactor Elements. The basic idea consists in feeding the two control winding elements N'_C and N''_C of the basic two-core internal-feedback circuits having a-c or d-c load (Figs. 11.1 and 11.3) in such a way that push-pull action is obtained. These winding elements may be incorporated into two separate (d-c or a-c) control circuits, or they may be series-connected or parallel-connected in a common (d-c or a-c) control circuit.

According to the simple procedure, presented in Sec. 8.2, the fundamental principle of the duodirectional internal-feedback circuit with two saturable-reactor elements will be explained by comparing its mode of operation with that of the basic "doubler" circuit, as shown in Fig. 11.1.

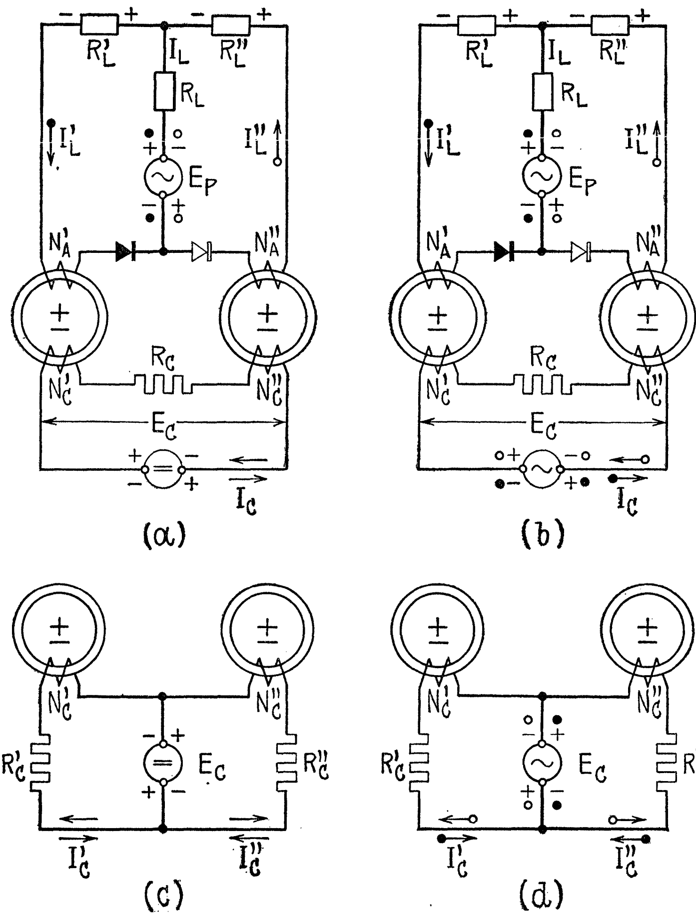


FIG. 12.1. Four diagrams illustrating the mode of operation of the basic self-saturating "doubler" circuit shown in Fig. 11.1: (*a*) d-c control with series-connected control windings N_C' and N_C''; (*b*) a-c control with series-connected control windings N_C' and N_C''; (*c*) d-c control with parallel-connected control windings N_C' and N_C''; (*d*) a-c control with parallel-connected control windings N_C' and N_C''.

a. Mode of Operation of the Original Internal-feedback Circuit. Figure 12.1 illustrates four different possibilities for controlling the basic self-saturating circuit by means of a reversible control voltage E_C. In this case, both the half-cycle pulses I_L'' and I_L' will increase or decrease (Fig. 8.3) when the d-c (or a-c) control voltage E_C is changed from zero to positive or negative values (or from zero to values having 0 or 180°

phase displacement between E_C and E_P, respectively). Therefore, the actual impedance values of the two saturable-reactor elements will at all times decrease or increase simultaneously, as indicated in Fig. 12.1 by the same symbol ($\pm$) in both core elements.

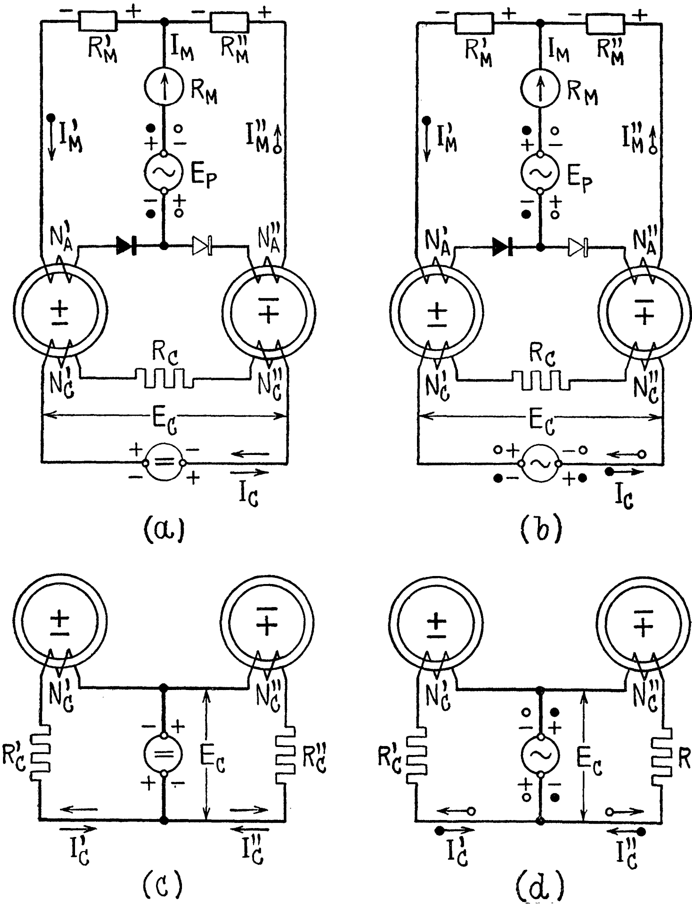


FIG. 12.2. Four diagrams illustrating the mode of operation of the duodirectional self-saturating full-wave circuit: (*a*) d-c control with series-connected control windings N'_C and N''_C; (*b*) a-c control with series-connected control windings N'_C and N''_C; (*c*) d-c control with parallel-connected control windings N'_C and N''_C; (*d*) a-c control with parallel-connected control windings N'_C and N''_C.

b. Mode of Operation of the Duodirectional Internal-feedback Circuit. Figure 12.2 illustrates four different possibilities for controlling the self-saturating circuit with two saturable-reactor elements in such a way that push-pull action is obtained, using a reversible control voltage E_C, which may be either d-c or a-c. In this case, according to Fig. 8.5, the half-cycle pulses I'_L will increase, and the half-cycle pulses I''_L will decrease (and vice versa), when the d-c (or a-c) control voltage E_C is changed from zero to positive or negative values (or from zero to values having 0 or 180° phase displacement between E_C and E_P, respectively). Therefore, the actual impedance values of the two saturable-reactor elements will always change in opposite directions, as indicated in Fig. 12.2 by opposite symbols ($\pm$ and $\mp$) in the two core elements.

Referring to the special problems which are involved in push-pull feedback circuitry using only two saturable-reactor elements (Sec. 8.2*b*), it is important that the actual turns ratio $N'_C/N'_A = N''_C/N''_A$ have a sufficiently small value (about 0.01 to 0.1). Furthermore, it will be necessary to connect a sufficiently high resistance R_C in series with N'_C and N''_C so that the fundamental-frequency voltages induced from the load windings N'_A, N''_A into N'_C and N''_C are unable to produce excessively large currents in the control-circuit loop.

12.3. Practical Applications of Push-Pull Internal-feedback Circuits with Two Saturable-reactor Elements. Reference to Fig. 8.5 shows that there are two different operating conditions, *a* and *b*, which have already been described in Sec. 8.3.

Operating Conditions a. Figure 12.2 illustrates the possibility of measuring the reversible d-c component $\pm I_{dc} = f(\pm E_C)$ of the load current I_M by means of a moving-coil ammeter (d-c load R_M). This d-c component can preferably be used for supplying the d-c control windings of a succeeding push-pull magnetic-amplifier stage. The actual a-c component I_{ac}, which is also a function of E_C, may be either eliminated (*e.g.*, by means of a shunt condenser across the moving-coil ammeter) or used for supplying the a-c bias windings of such an amplifier stage.

Operating Conditions b. Figure 12.2 illustrates also the possibility of providing twin-type load resistors R'_M, R''_M, which may be represented by two equally rated parts of the control windings of a separately excited, two-phase, induction-type reversible motor, as used in high-performance servomechanisms. In this case, the actual d-c component I_{dc}, which is also a function of E_C, can successfully be used for introducing effective dynamic braking of the two-phase motor. This artifice has proved to be particularly valuable with regard to those types of reversible a-c motors which have virtually no damping properties, so that additional damping methods must be applied to prevent overshooting or hunting.

12.4. Two-stage Magnetic Servo Amplifier Having Inherent Dynamic-braking Properties. The following illustration of a two-stage magnetic servo amplifier of the full-wave, push-pull internal-feedback type gives a representative example for practical applications of two modified circuits, as shown in Figs. 12.3 and 12.4. This arrangement was devised by the author[1] in 1952.

a. Output-stage Circuit. According to Fig. 12.3, the basic idea consists in supplying the amplifier field winding W_L of the two-phase induction motor from a *full-wave* self-saturating circuit carrying the unidirectional current I_L having waveshapes as characterized in Fig. 8.5*b*. The line field winding W_P of this motor is supplied with a substantially constant current, which is proportional to the power-supply voltage E_P (f_P = 60 or 400 cps). This output-stage circuit can be designed so that

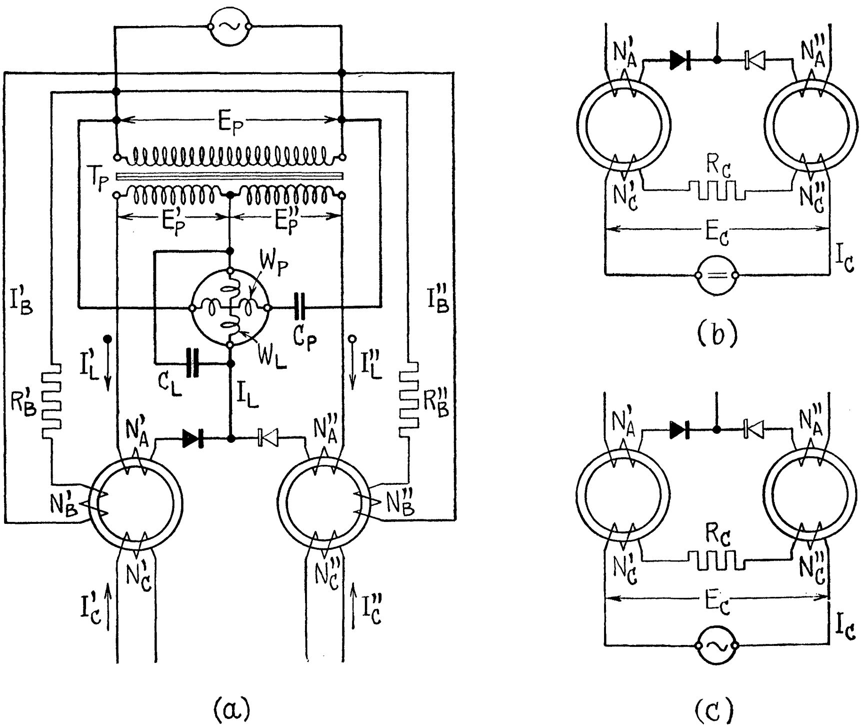


FIG. 12.3. Full-wave output-stage circuit of the magnetic servo amplifier having inherent dynamic-braking properties: (*a*) with two separate control winding elements N_C', N_C''; (*b*) with series-connected d-c control winding elements N_C', N_C''; (*c*) with series-connected a-c control winding elements N_C', N_C''.

the d-c component producing dynamic braking decreases when the torque-producing fundamental-frequency component of I_L increases, and vice versa.

Comparison between Figs. 12.3 and 11.3 shows that the *duodirectional* full-wave output-stage circuit (Fig. 12.3) is a modification of the original full-wave circuit (Fig. 11.3) representing a simple *polarized* magnetic amplifier. The essential difference between these two circuits consists in the fact that, in the output-stage circuit of Fig. 12.3, the two control winding elements N_C' and N_C'' are fed by two separate control currents I'_C and I''_C (Fig. 12.3*a*) or by a common control current I_C (Fig. 12.3*b* and *c*) in such a way that push-pull action is obtained. Thus, these winding elements may be incorporated into two separate (d-c or a-c) control circuits (Fig. 12.3*a*), or they may be series-connected or parallel-connected in a common (d-c or a-c) control circuit. Figure 12.3*b* and *c* illustrates the examples that the series-connected control

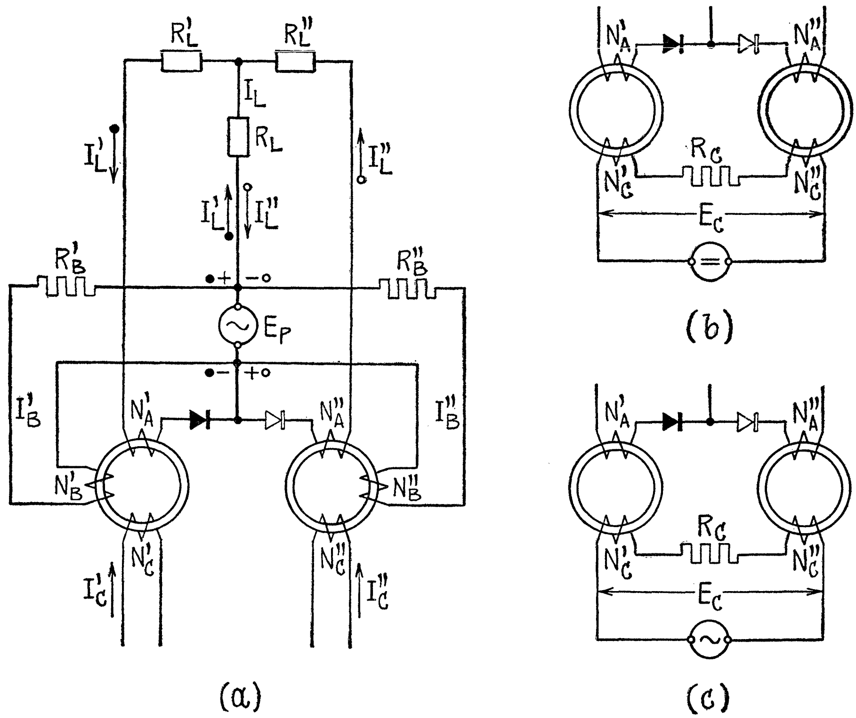


FIG. 12.4. Full-wave input-stage circuit of the magnetic servo amplifier having inherent dynamic-braking properties: (*a*) with two separate control winding elements N'_C, N''_C; (*b*) with series-connected d-c control winding elements N'_C, N''_C; (*c*) with series-connected a-c control winding elements N'_C, N''_C.

winding elements N'_C, N''_C with series resistor R_C are d-c controlled or a-c controlled by means of a reversible control voltage $E_C = I_C R_C$.

b. Input-stage Circuit. Figure 12.4 illustrates the mode of operation of the duodirectional full-wave input-stage circuit which is used for the following purposes:

1. It acts as a push-pull type of *magnetic amplifier* to produce a reversible d-c component of load current I_L, a function of the reversible a-c error voltage, for controlling the d-c control winding elements N'_C, N''_C of the output-stage circuit (Fig. 12.3*b*).

2. It acts as a *phase discriminator* (in a similar way to a phase-sensitive rectifier) so that actual direction of the d-c component of I_L will correspond to the actual phase displacement (0 or 180°) between error voltage E_C and power-supply voltage E_P (Fig. 12.4*c*).
3. It acts as a *voltage-sensitive bias element* to the output-stage circuit of Fig. 12.3*b* for introducing special voltage-current characteristics (Sec. 7.6).
4. It acts as a decoupling device in such a way that no appreciable fundamental-frequency voltages will be induced into the synchrotransformer circuit (Fig. 12.4*c*), which has a total impedance of about 10,000 ohms, or more.

Thus, the output current I_L of the input-stage circuit of Fig. 12.4 will provide:

1. An *a-c bias* component, which is a function of the actual magnitude of the power-supply voltage E_P
2. A reversible *d-c control* component, which is a function of the actual a-c error voltage E_C in the synchrotransformer circuit (Fig. 12.4*c*)

c. Full-wave Two-stage Circuits. Correct and efficient operation of the complete two-stage circuit will be obtained only when the a-c (60- or 400-cycle) components of the currents flowing through the various parts of this circuit have a proper phase relationship. The following two examples show a simple way for obtaining proper phase displacement, by means of several condensers, so that the most favorable working conditions will be secured. The final two-stage circuits are results of a special experimental study, and the optimum values of the condensers, which are not critical, have been determined empirically.

Figure 12.5 shows a full-wave two-stage circuit which contains:

1. The output-stage circuit, as shown in Fig. 12.3*b*
2. The input-stage circuit, as shown in Fig. 12.4*c*
3. The power-supply transformer T_P having a center-tapped secondary winding (voltages E_P', E_P'') and an additional secondary winding supplying the input-stage circuit

Proper phase relationship of the currents flowing through the various parts of this two-stage circuit and maximum torque conditions of the two-phase motor will be obtained by means of the condensers C_P, C_L, C_D, and C_S.

It is to be noted that the actual values of output-stage quiescent currents I_L', I_L'' can easily be controlled by adjusting the variable a-c bias resistor R_B. In this way, the actual magnitude of the dynamic-braking effect can be adjusted so that not more than one overshoot, or no overshoot at all, will be obtained.

The bridge-type output-stage circuit in Fig. 12.6 makes it possible to eliminate the comparatively large power-supply transformer (T_P in Fig. 12.5) and to reduce size and weight of the complete magnetic servo

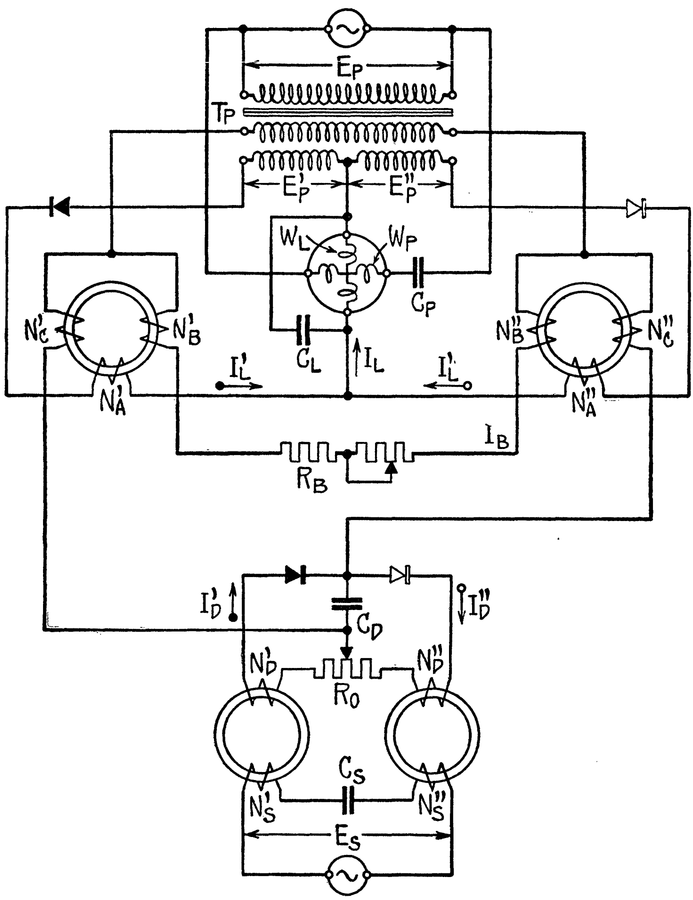


FIG. 12.5. Complete two-stage full-wave circuit of a magnetic servo amplifier with center-tapped power-supply transformer T_P. (*W. A. Geyger.*[1])

amplifier by dividing the load windings N'_A, N''_A into two equally rated parts, which are insulated from each other. Note that the total number of dry-disk rectifier cells remains unchanged, so that, for example, 4×12 cells may be used in the bridge-type circuit (Fig. 12.6), whereas 2×24 cells have to be provided in the transformer-type circuit (Fig. 12.5).

It is important to note that application of the small power-supply transformer T_D of the input-stage circuit makes it possible to reduce the number of cells of the half-wave rectifiers in this circuit and to reduce the number of turns of input-stage load windings N'_D, N''_D and output-

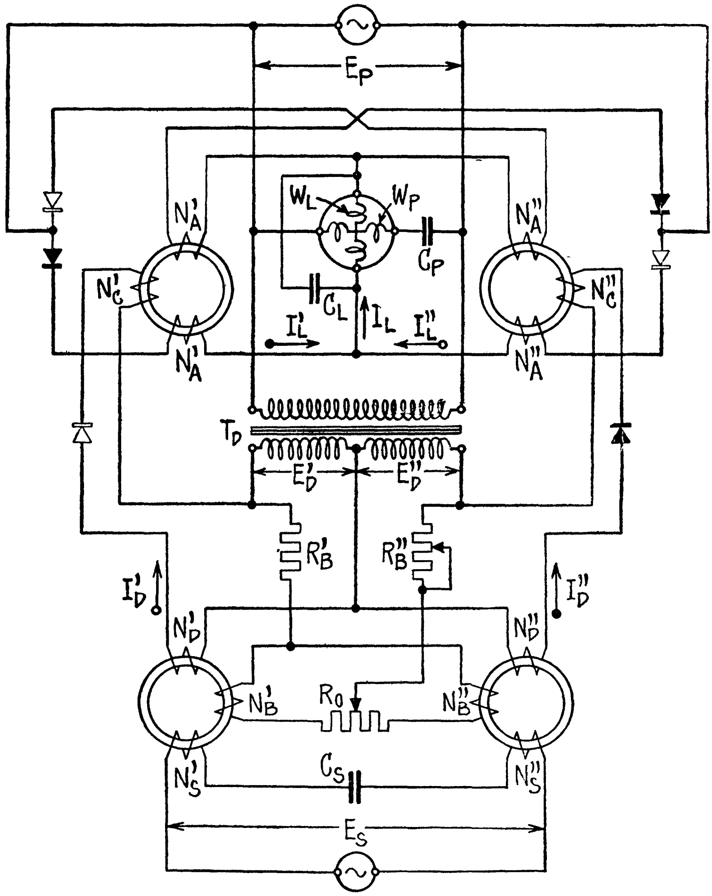


FIG. 12.6. Complete two-stage full-wave circuit of a magnetic servo amplifier without power-supply transformer. (*W. A. Geyger.*[1])

stage control windings N'_C, N''_C. The transformer T_D, however, may also be eliminated by properly varying these circuit components.

Furthermore, in any case, the two input-stage control windings N'_S and N''_S, each having its own series condenser, may be connected in parallel to the control voltage E_S of the synchro control transformer to obtain perfect symmetry of the transfer characteristic $I_L = f(E_S)$ by proper rating of these two series capacitors.

In this two-stage circuit (Fig. 12.6) the dynamic-braking effect, which is determined by the actual magnitude of the output-stage quiescent current (I_L with $E_S = 0$), can easily be adjusted by means of the bias resistor R''_B, which, in this case, controls the actual magnitude of the a-c bias currents of the input-stage circuit.

Referring to the method for obtaining perfect zero adjustment

$$I_L = I'_L - I''_L = 0 \text{ if } E_S = 0$$

it is to be noted that this condition can easily be achieved by adjusting the potentiometer-type resistor R_0 (Figs. 12.5 and 12.6), which controls the actual value of current ratio I'_D/I''_D of the input-stage circuit.

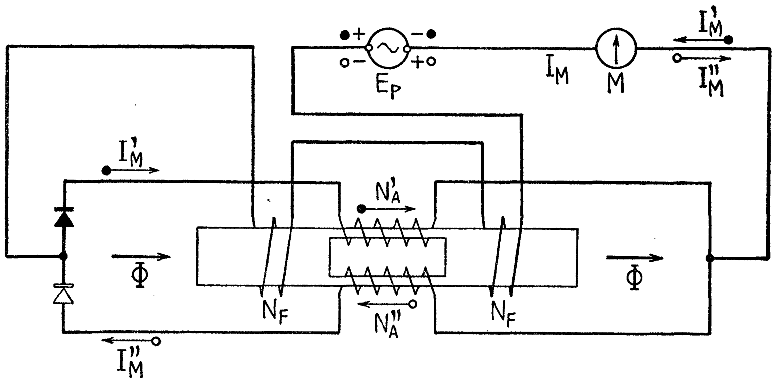


FIG. 12.7. Full-wave self-saturating type of flux-gate-magnetometer circuit with two half-wave dry-disk rectifiers and additional (regenerative) feedback windings N_F. (*G. Barth, Swiss patent* 224,992, 1942.)

The actual performance characteristics of the *internal*-feedback-type two-stage circuits of Figs. 12.5 and 12.6 are similar to those (Sec. 8.5) of the equivalent *external*-feedback-type circuit, as shown in Fig. 8.6.

12.5. Full-wave Internal-feedback-type Flux-gate-magnetometer Circuits. G. Barth introduced various types of flux-gate-magnetometer circuits utilizing "external," "internal," or "compound" feedback. These arrangements, operated either by an a-c generator in connection with rectifiers or by a d-c generator in connection with mechanical choppers or modulators (Fig. 10.1), may have either an "open" or a "closed" magnetic circuit of high-permeability core material, preferably Permalloy or Mumetal.

Figure 12.7 illustrates a typical example of a simple full-wave internal-feedback circuit with additional d-c external-feedback windings, operating as a self-saturating flux-gate magnetometer.[2] This arrangement contains:

1. An a-c generator producing the power-supply voltage E_P (frequency f_P preferably about 400 to 600 cps)

2. The load windings N'_A, N''_A of a special construction of saturable reactor that is magnetically controlled by the magnetic field Φ to be investigated, *e.g.*, by the earth's field
3. Two half-wave rectifiers producing self-saturating effects
4. Additional d-c external-feedback windings N_F
5. A moving-coil instrument M (or another d-c device) representing the d-c load of the arrangement

The load windings N'_A, N''_A are connected with the rectifiers in such a way that the d-c components of the two pulsating, unidirectional currents I'_M and I''_M produce, in the two central parts of the core having reduced cross-sectional area, d-c magnetizations of opposite directions. Consequently, an *additive* magnetization caused by one of the windings (N'_A) is superimposed in one of these two parts on the magnetization caused by the magnetic field Φ to be investigated; and the magnetization caused by the winding (N''_A) of the other part is *subtractive* with respect to the magnetization effected by the field Φ.

It is to be noted that, with no-signal conditions (with $\Phi = 0$), the two half-cycle pulses I'_M and I''_M have exactly the same magnitude. Thus, the load current I_M flowing through the moving-coil galvanometer M is a pure alternating current having no d-c component. When a magnetic-field component arises, as indicated in Fig. 12.7, then the first-half-cycle pulses I'_M will increase, and the second-half-cycle pulses I''_M will decrease. Thus, there will be a *positive* d-c component, proportional to $I'_M - I''_M$, which is a function of Φ. However, if Φ has the opposite direction, the first-half-cycle pulses I'_M will decrease, while the second-half-cycle pulses I''_M will increase; and there is a *negative* d-c component, proportional to $I''_M - I'_M$, which is a function of Φ. Because the impedance of each of the windings N'_A, N''_A is varied in accordance with the resultant magnetization in the appertaining part of the core, the moving-coil galvanometer responds to these half-wave currents and indicates their differential effect.

Evidently, the fundamental mode of operation of the flux-gate-magnetometer circuit (Fig. 12.7) is similar to that of the magnetic-amplifier circuit, as shown in Figs. 12.2 and 12.4.

Referring to the possibility of employing additional d-c feedback windings N_F, it may be noted that additional external-feedback effects produced by these windings may be either regenerative or degenerative. Figure 12.7 illustrates the case that regenerative (positive) feedback is provided to increase the sensitivity of the arrangement.

Obviously, the d-c component $\pm I_{dc} = \pm\Phi$ of the load current I_M may be used for supplying the d-c control windings of a succeeding push-pull magnetic-amplifier stage. The actual a-c component I_{ac}, which is also a function of Φ, may be either eliminated (by means of a shunt condenser across the moving-coil galvanometer) or used for supplying the a-c bias windings of such an amplifier stage.

Figure 12.8 shows a full-wave compound-feedback type of flux-gate-magnetometer circuit with a mechanical rectifier *MR* and d-c external-feedback windings N_F, which, in this case, produce an additional degenerative (negative) feedback effect.[2] This may be advisable, because the actual magnitude of the regenerative (positive) internal feedback effect,

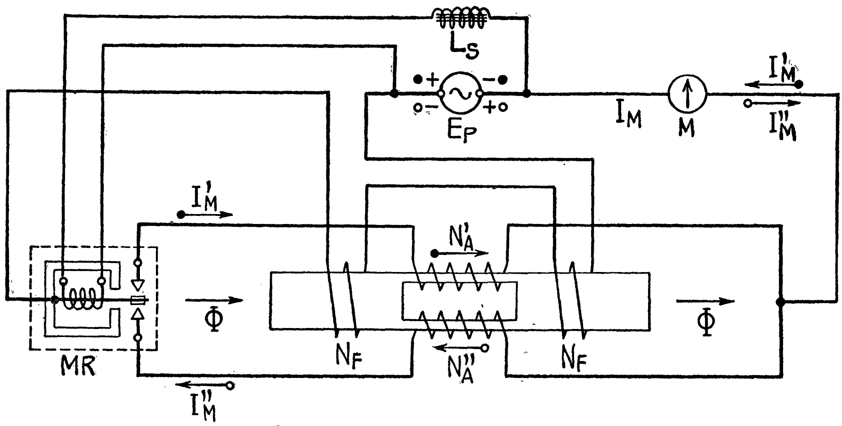


Fig. 12.8. Full-wave self-saturating type of flux-gate-magnetometer circuit with a mechanical rectifier *MR* and additional (degenerative) feedback windings N_F. (*G. Barth, United States patent* 2,390,051, *Figure* 2, 1941.)

produced by the mechanical rectifier *MR*, which has ideal characteristics (forward resistance = 0, reverse resistance = ∞), may be too large. The mechanical rectifier is separately excited by the power-supply voltage E_P, and a properly rated inductive series impedance L_S is provided to obtain the desired phase displacement between E_P and the exciting current of the rectifier *MR*.

REFERENCES

1. W. A. Geyger, "A New Type of Magnetic Servo Amplifier," *AIEE Transactions*, Vol. 71, Part I, pp. 272–280, 1952.
2. See "Einrichtung zum Bestimmen der Grössen eines magnetischen Feldes" (Arrangement for measuring the magnitudes of a magnetic field), Swiss patent 224,992, issued December 31, 1942 (application filed February 16, 1942). Also G. Barth, "Means for Measuring Magnetic Fields," United States patent 2,390,051, issued December 4, 1945 (application filed July 15, 1941).

CHAPTER 13

Half-wave Types of Duodirectional Internal-feedback Circuits with Two Saturable-reactor Elements

13.1. Fundamental Principle of Half-wave Push-Pull Circuits. It has been mentioned in Chap. 12 that the half-wave push-pull circuits are characterized by the fact that only one half cycle of the power-supply voltage (E_P) is used for supplying the load windings and the a-c or d-c load.

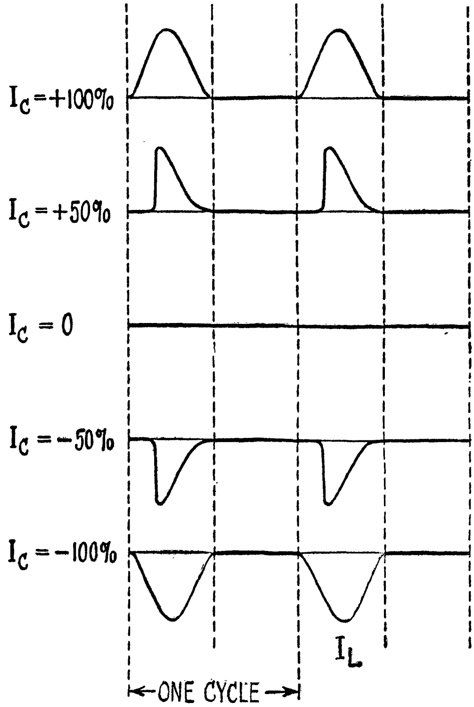


FIG. 13.1. Waveshapes of the reversible output current I_L of half-wave push-pull circuits.

Figure 13.1, illustrating the fundamental principle of these circuits, shows that magnitude and direction of the half-wave load current I_L is a function of the actual value (+100, +50, 0, −50, −100 per cent) of the reversible control current I_C, which may be derived from a d-c or a-c circuit.

Various types of flux-gate magnetometers and magnetic amplifiers are based upon this fundamental principle. This chapter describes

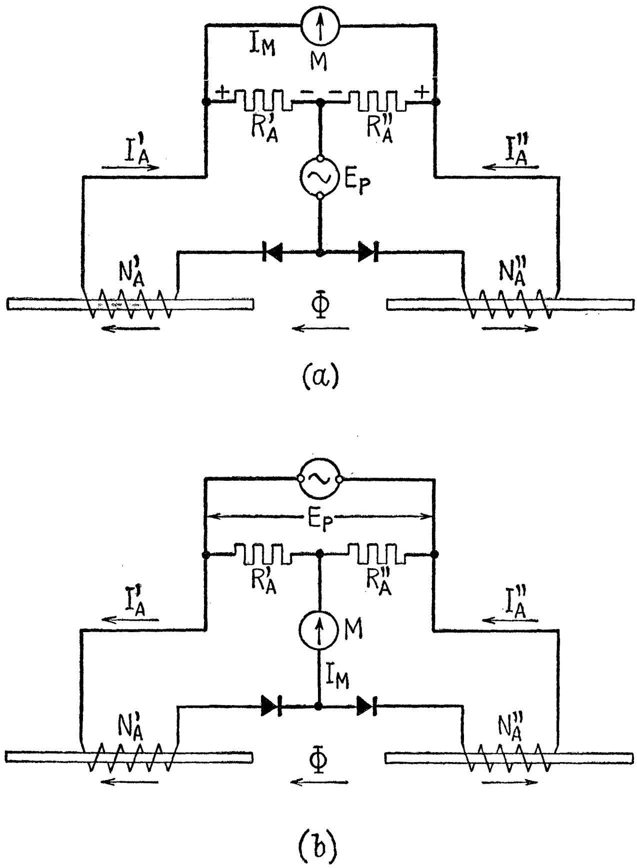


FIG. 13.2. Bridge-type half-wave flux-gate-magnetometer circuit with two dry-disk-rectifier elements. (*G. Barth, United States patent* 2,390,051, *Figure* 3, 1941.) (*a*) The moving-coil galvanometer M is connected across the resistors R'_A, R''_A. (*b*) The a-c generator (voltage E_P) is connected across the resistors R'_A, R''_A.

some typical examples of such arrangements, which may be either of the single-stage or the multistage type.

13.2. Half-wave Types of Flux-gate-magnetometer Circuits. G. Barth[1,2] introduced various types of half-wave flux-gate-magnetometer circuits, as shown in Figs. 13.2 to 13.4. These arrangements employ either two separated magnetic cores in the form of a rod ("open" magnetic circuits, Figs. 13.2 and 13.3), or only one saturable-reactor element ("closed" magnetic circuit, Fig. 13.4). High-permeability core material, preferably Permalloy or Mumetal, is provided in both cases to obtain a very high sensitivity of the self-saturating circuits.

Figure 13.2 shows two examples of a bridge-type half-wave flux-gate magnetometer containing:

1. An a-c generator producing the power-supply voltage E_P (frequency f_P preferably about 400 to 600 cps)
2. The load windings N'_A, N''_A of two equally rated saturable-reactor elements, each having a high-permeability core in the form of a rod, the longitudinal dimension being great in proportion to the transverse dimension
3. Two half-wave rectifiers producing self-saturating effects
4. Two series resistors R'_A, R''_A
5. A moving-coil instrument M (or another d-c device) representing the d-c load of the arrangement

The windings N'_A, N''_A are connected with the half-wave rectifiers in such a way that the d-c components of the pulsating, unidirectional currents I'_A, I''_A produce in the magnetic cores d-c magnetizations of opposite directions. Consequently, an *additive* magnetization caused by one of the windings (N'_A) is superimposed in one core on the magnetization caused by the magnetic field Φ to be investigated, while the magnetization caused by the winding (N''_A) of the other core is *subtractive* with respect to the magnetization effected by the field Φ.

It is to be noted that only the positive (or negative) half waves of current from the a-c generator flow through the windings N'_A and N''_A. Thus, the polarities of the voltage drops across the series resistors R'_A, R''_A in Fig. 13.2*a* will be opposite, and actual magnitude of the load current I_M flowing through the moving-coil instrument (d-c load resistor) M will be proportional to the actual difference of the half-wave currents I'_A and I''_A. Because the impedance of each of the windings N'_A, N''_A is varied in accordance with the resultant magnetization in the appertaining core, the moving-coil instrument responds to these half-wave currents and indicates their differential effect. Assuming symmetrical circuit conditions ($R'_A = R''_A = R_A$) in the arrangement of Fig. 13.2*a*, the actual load current flowing through M is

$$I_M = (I'_A - I''_A)\frac{R_A}{2R_A + R_M} = \text{const} \times (I'_A - I''_A) \qquad (13.1)$$

where R_M represents the resistance of the d-c load M. Evidently, in the similar arrangement of Fig. 13.2*b*, the actual load current I_M is proportional to the difference of the two half-wave currents:

$$I_M = \text{const} \times (I'_A - I''_A) \tag{13.2}$$

Figure 13.3*a* and *b* illustrates two possibilities for using a d-c power-supply voltage E_P in connection with periodically operated contact devices acting as choppers or modulators in such arrangements.

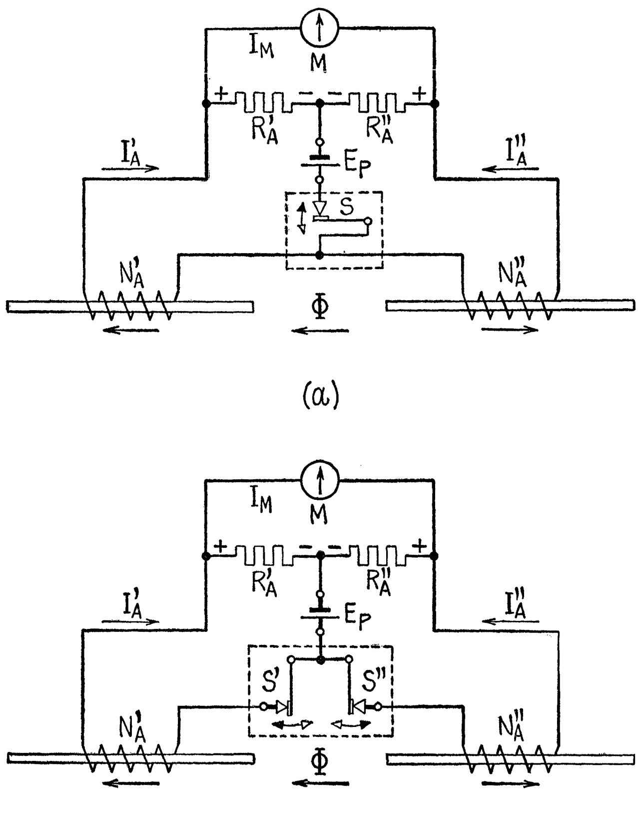


FIG. 13.3. Bridge-type half-wave flux-gate-magnetometer circuit with d-c power-supply voltage E_P and periodically operated contact devices acting as choppers or modulators. (*G. Barth, United States patent* 2,390,051, *Figure* 4, 1941.) (*a*) The contact device S is series-connected with the battery circuit. (*b*) The contact devices S', S'' are series-connected with the load windings N'_A, N''_A.

Figure 13.4 shows an improved bridge-type half-wave flux-gate magnetometer with a special saturable-reactor construction providing a "closed" magnetic circuit.[3] Each of the load windings N'_A and N''_A is divided into two equally rated parts, each part having a series-connected half-wave rectifier producing self-saturation. The four load winding units are connected with the associated four rectifier units in such a way that the d-c components of the two half-wave currents I'_M and I''_M produce, in the two central parts of the core having reduced cross-sectional area, d-c magnetizations of opposite directions. Consequently, an *additive* magnetization caused by one of the windings (N'_A) is superimposed in one of these two parts on the magnetization caused by the magnetic field Φ to be investigated; and the magnetization caused by the winding (N''_A) of the other part is *subtractive* with respect to the magnetization effected by the field Φ. Comparison between Figs. 12.7 and

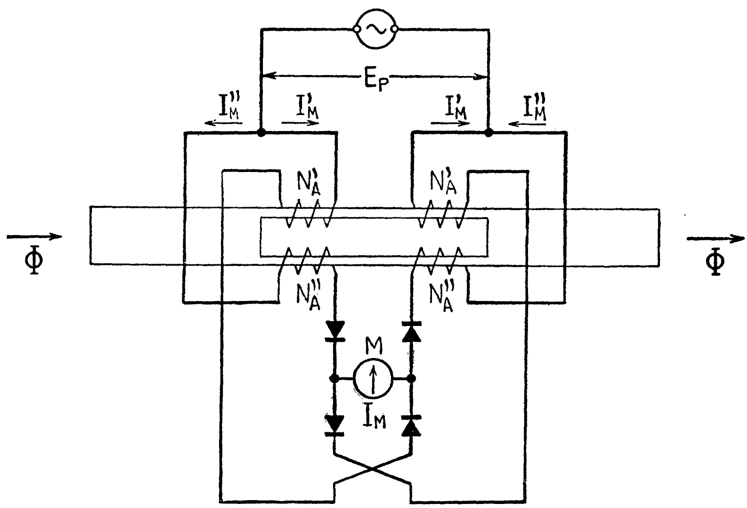


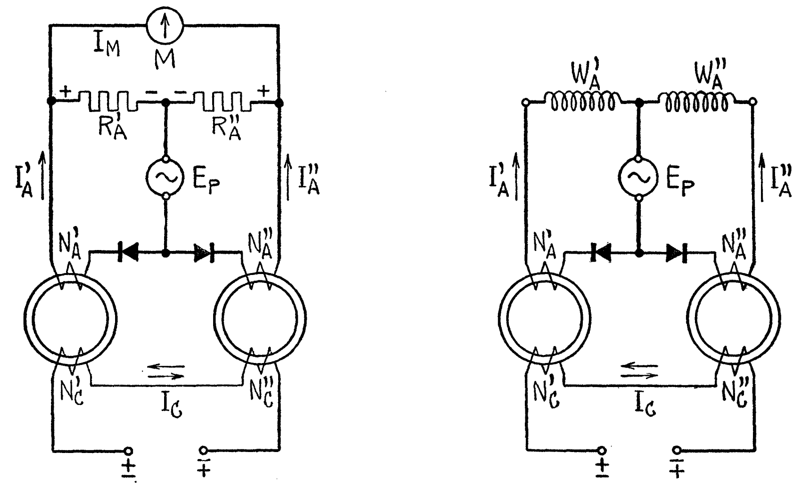
FIG. 13.4. Bridge-type half-wave flux-gate magnetometer with a special saturable-reactor construction providing a "closed" magnetic circuit. (*G. Barth, Swiss patent* 224,992, *Figure* 4, 1942.)

13.4 makes it evident that the fundamental mode of operation of the half-wave circuit is similar to that of the full-wave arrangement.

13.3. Half-wave Types of Push-Pull Magnetic-amplifier Circuits. In 1938, A. L. Whiteley and L. C. Ludbrook[4] devised several basic half-wave push-pull types of magnetic-amplifier circuits. Figure 13.5*a*, illustrating the operation of one of these circuits, shows that a reversible-polarity half-wave load current I_M (Fig. 13.1) can be obtained and controlled by means of a relatively small (d-c or a-c) control current I_C which flows through the series-connected control winding elements N'_C and N''_C. The actual magnitude of I_M is given by Eq. (13.1).

Figure 13.5*b* shows a modified arrangement.[5] The series resistors R'_A, R''_A have been replaced by windings W'_A, W''_A, which may be considered the operation windings of conventional relays or contactors, for example. In this manner an arrangement is produced which is the equivalent of a very sensitive polarized relay, although the actual relays which are employed are inexpensive and rugged standard nonpolarized relays or contactors.

It is to be noted that the basic half-wave push-pull circuits, as shown in Figs. 13.2 to 13.5, use only the d-c component of the reversible half-wave load current $I_M = \text{const} \times (I'_A - I''_A)$ flowing through the d-c devices. Therefore, shunt condensers may be connected across the series resistors (Fig. 13.5*a*) or operating windings (Fig. 13.5*b*) so as to smooth out the voltage drop across these circuit elements.



(a) (b)

FIG. 13.5. Half-wave push-pull magnetic-amplifier circuits (*A. L. Whiteley and L. C. Ludbrook, United States patent* 2,229,952, *Figures* 3 *and* 5, 1938): (*a*) with two resistors R'_A, R''_A and d-c load (moving-coil instrument) M; (*b*) with two operating windings W'_A, W''_A of relays, contactors, or other d-c devices.

L. R. Crow[6] has stressed that, since the output of this type of magnetic amplifier is zero when the input is zero, several similar stages may be easily cascaded to increase the sensitivity to any reasonable degree. He has also shown[7] that, if the two series resistors (R'_A, R''_A in Fig. 13.5*a*) are replaced by the differential twin-type control windings (W'_L, W''_L in Fig. 13.6*a*) of a separately excited reversible a-c motor, the arrangement results in a sensitive reversible a-c motor control.

Figure 13.6*b* shows another half-wave reversible a-c motor-control circuit, described in 1945 by C. S. Hudson.[8] A bridge-type arrangement with center-tapped power-supply transformer T_P is provided here, so that the a-c component of the reversible half-wave load current

$$\pm I_L = \text{const} \times (I'_L - I''_L)$$

is a function of the reversible (d-c or a-c) control current I_C. Evidently, the half-wave circuit of Fig. 13.6*b* is similar to the full-wave circuit of Fig. 12.3. It is to be noted, however, that, in the half-wave circuit of Fig. 13.6*b*, the torque-producing a-c (60- or 400-cycle) component and the associated d-c component of the half-wave load current

$$\pm I_L = \text{const} \times (I'_L - I''_L)$$

will always increase or decrease at the same rate. Thus, with no-signal conditions ($I_C = 0$), the two load currents I'_L and I''_L are equal, and there

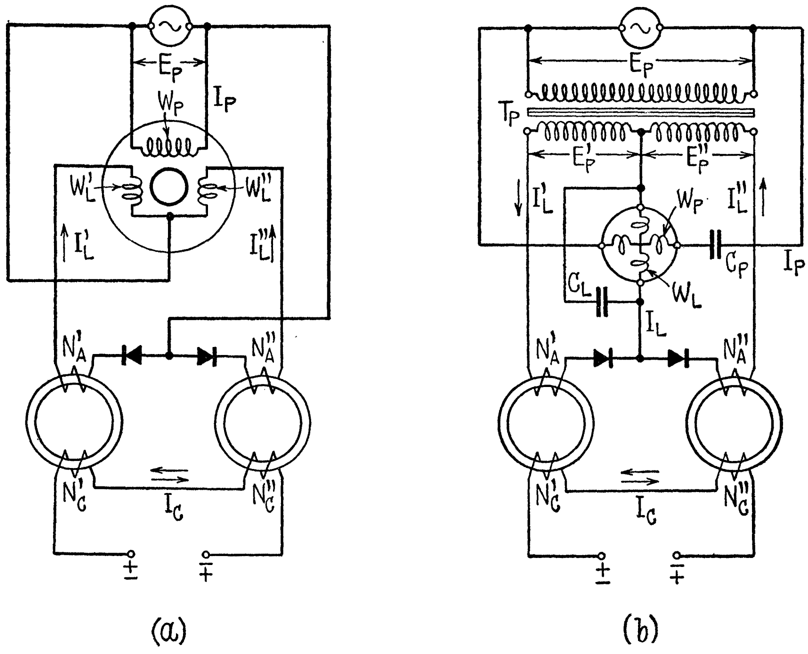


FIG. 13.6. Half-wave push-pull a-c motor-control circuits: (*a*) differential-type half-wave circuit for reversible a-c motor control (*L. R. Crow*[7]); (*b*) bridge-type half-wave push-pull circuit for reversible a-c motor control with center-tapped power-supply transformer T_P (*C. S. Hudson*[8]).

is no damping effect on the motor. This is generally not desirable in connection with two-phase induction motors which have very poor damping properties, because there is no dynamic-braking effect in the vicinity of the null point. However, these half-wave push-pull circuits have proved to be very useful with a reversible a-c motor having inherent damping properties or special damping means.

Figure 13.7 illustrates the possibility of eliminating the power-supply transformer (T_P in Fig. 13.6*b*) by providing a simple symmetrical bridge-type circuit, similar to that of Fig. 13.4. This arrangement, devised by

G. Barth[9] in 1942, may be modified in various ways.[10] Each of the load windings N'_A and N''_A is divided into two equally rated parts, each part having a series-connected half-wave rectifier producing self-saturation. The four load winding units are connected with the associated four rectifier units in such a way that the d-c components of the two half-wave currents I'_L and I''_L produce in the central parts of the two core elements d-c fluxes of opposite directions. Consequently, an *additive* d-c flux caused by one of the load windings (N'_A) is superimposed in one of these

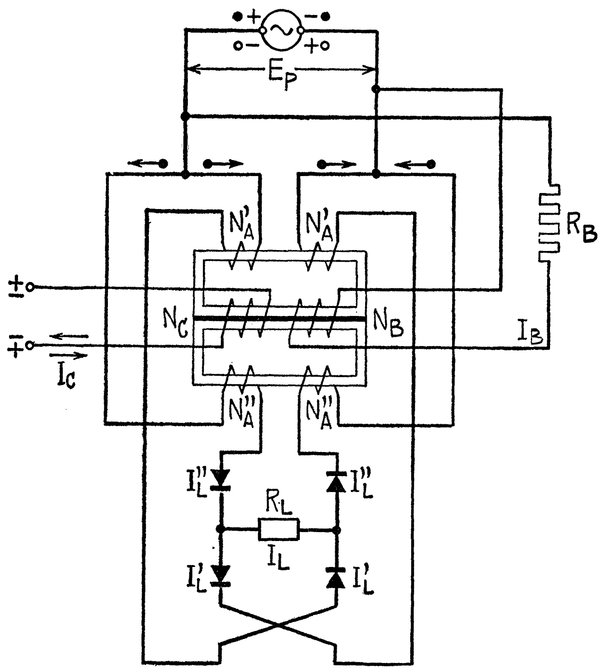


FIG. 13.7. Bridge-type half-wave push-pull circuit without power-supply transformer. (*G. Barth, Swiss patent* 233,014, 1942.)

two parts on the d-c flux caused by the reversible control current I_C; and the d-c flux caused by the other load winding (N''_A) is *subtractive* with respect to the d-c flux effected by I_C flowing through the common control winding N_C. The two saturable-reactor elements may preferably be biased, *e.g.*, by means of an alternating current I_B flowing through the common bias winding N_B with series resistor R_B.

13.4. Two-stage Half-wave Magnetic-servo-amplifier Circuits. Half-wave, push-pull, self-saturating circuits, as shown in Figs. 13.2 to 13.7, represent the basic units of various types of two-stage arrangements that are especially suitable for designing of high-performance magnetic servo amplifiers. The following illustration of a two-stage half-wave type of magnetic servo amplifier (Fig. 13.8), devised in 1952 by C. W. Lufcy[11] and his staff, gives a typical example of such an arrangement.

a. Output-stage Circuit. The amplifier field winding W_L of the two-phase induction motor is supplied from a bridge-type half-wave circuit

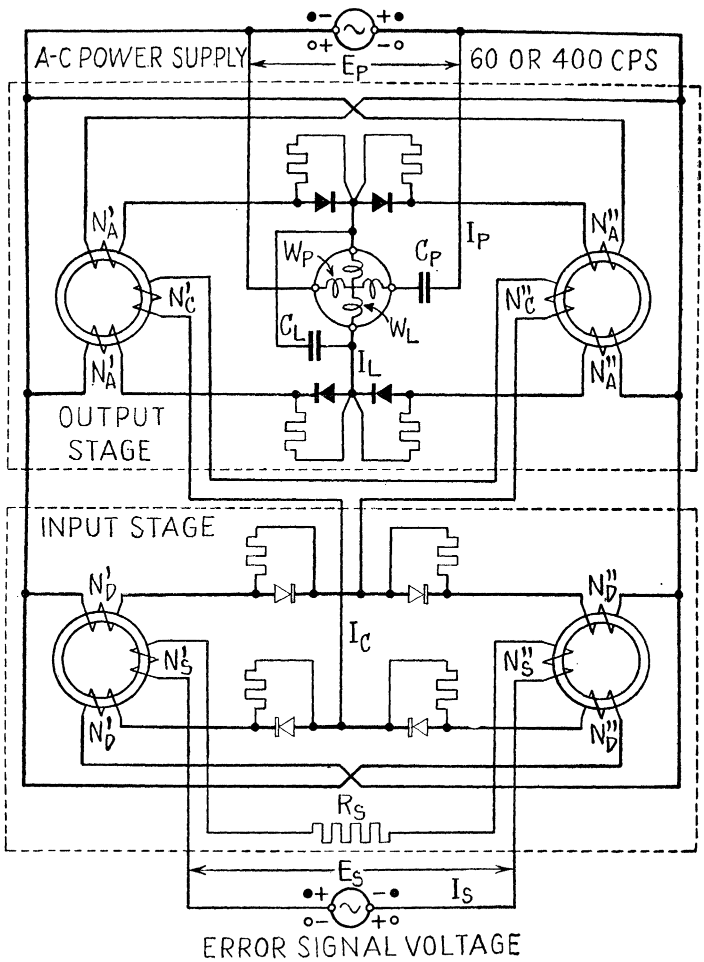


FIG. 13.8. Two-stage magnetic servo amplifier with two bridge-type half-wave push-pull circuits. (*C. W. Lufcy, A. E. Schmid, and P. W. Barnhart.*[11])

with the pulsating, unidirectional current I_L having waveshapes as characterized in Fig. 13.1. The line field winding W_P of this motor is fed with a substantially constant current I_P, which is proportional to the power-supply voltage E_P (frequency f_P = 60 or 400 cps). Each of the load windings N'_A and N''_A is divided into two equally rated parts, which are insulated from each other; and four series-connected half-wave-rectifier units are provided to produce self-saturation of the two saturable-reactor elements, which preferably have Orthonol tape cores with multilayer toroidal windings. The output current I_C of the input stage flowing through the control winding units N'_C, N''_C produces, in the two cores of the output stage, d-c fluxes of opposite directions. Consequently, an *additive* d-c flux caused by I_C is superimposed in one of the two cores on the d-c flux produced by the first load winding (*e.g.*, N'_A); and a *subtractive* d-c flux caused by I_C is superimposed in the other core on the d-c flux produced by the second load winding (*e.g.*, N''_A). The two saturable-reactor units may be biased by means of additional bias circuits. Another way of obtaining the desired working conditions consists in providing properly rated shunt resistors across the rectifiers, as indicated in Fig. 13.8.

b. Input-stage Circuit. This bridge-type arrangement, similar to that of the output stage, contains the four parts of the divided load windings N'_D, N''_D, four half-wave rectifiers with shunt resistors, and the series-connected control windings N'_S, N''_S carrying the alternating control (error-signal) current $I_S = E_S/R_S$ of the synchro-control-transformer circuit. Thus, the two saturable-reactor elements of the input-stage circuit are a-c controlled by means of the actual error-signal voltage E_S.

c. Two-stage Circuit. It is to be noted that the output stage is supplied with "positive" half-cycle pulses ("black" polarities) from power-supply voltage E_P, whereas the input stage carries the "negative" half-cycle pulses ("white" polarities), so that the cores of the output stage are controlled by means of the reversible current I_C during the "reset" half-cycle periods (Sec. 10.3). Proper phase relationship between the a-c components of the various currents in this two-stage circuit and maximum torque conditions of the two-phase motor are obtained by proper rating of the condensers C_P and C_L.

C. W. Lufcy, A. E. Schmid, and P. W. Barnhart[11] have shown that various refinements of this circuitry are possible and that excellent performance characteristics can be obtained with such arrangements,[12] especially when using those improved types of two-phase induction motors which have inherent damping properties.

13.5. Special Combinations of Half-wave Push-Pull Circuits. In 1942, G. Barth[13] devised various magnetic servo amplifiers with two separate half-wave push-pull arrangements, operating independently from each other, to obtain special performance characteristics and full-wave d-c or a-c output of the output-stage circuit.

Figure 13.9 shows such an arrangement, which contains two separate half-wave push-pull circuits, as already illustrated in Fig. 13.7, and two load units R'_L, R''_L, which may be, for example, the separate control winding units of a succeeding magnetic-amplifier stage, which also contains two separate half-wave push-pull circuits. It is to be noted that R'_L is fed from the circuit carrying the "positive" half-cycle pulses ("black" polarities), while R''_L carries the "negative" half-cycle pulses ("white" polarities).

Figure 13.10 illustrates the possibility of providing cross-connected external-feedback windings N'_F, N''_F and a common (a-c or d-c) load R_L, so that full-wave output is obtained (G. Barth,[14] 1942). It is to be

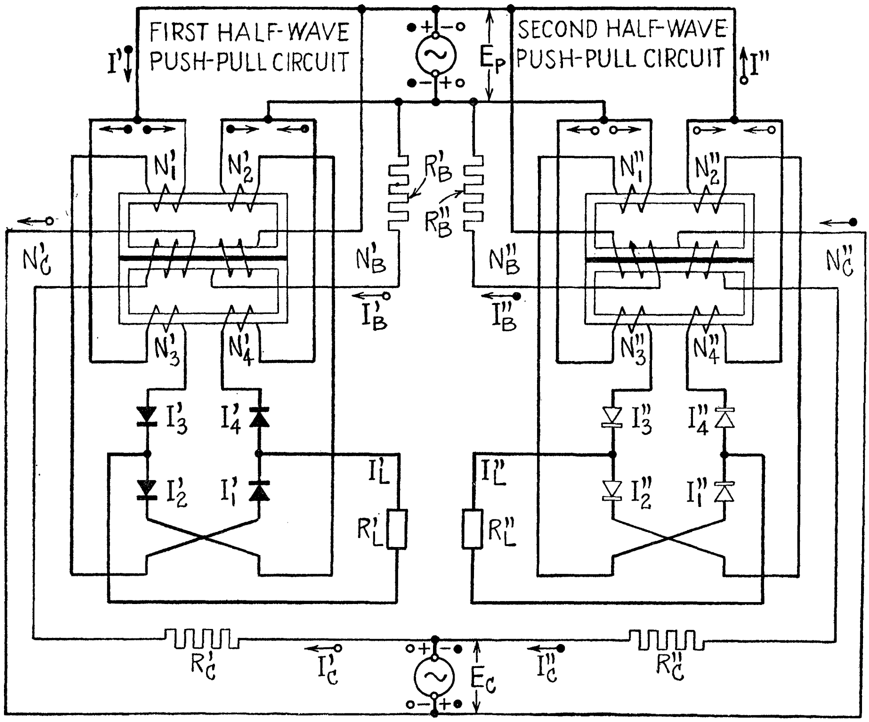


FIG. 13.9. Combination of two half-wave push-pull circuits with two separate load units R'_L and R''_L. (*G. Barth, Swiss patent* 233,014, 1942.)

noted that, in this case, the "black half-cycle circuit" is additionally influenced by the load current component I''_L of the "white half-cycle circuit," and vice versa. Obviously, the additional effects of the external-feedback windings N'_F and N''_F may be either regenerative or degenerative. Barth has also shown that these arrangements may be modified so that special "current-feedback," "voltage-feedback," or "derivative-feedback" effects may be obtained. He also pointed out that in such arrangements, provided with series-connected d-c control winding units N'_C and N''_C (Fig. 13.10), the fundamental-frequency voltages which are induced in these units are always canceling each other completely, not only with no-signal conditions but even when the maximum value of control voltage E_C is applied to the amplifier.

13.6. Modifications of the Basic Half-wave Push-Pull Circuits. C. S. Hudson[15] has shown that half-wave push-pull circuits may be used in such a way that an a-c load impedance is connected across series-connected secondary windings of two saturable-transformer elements.

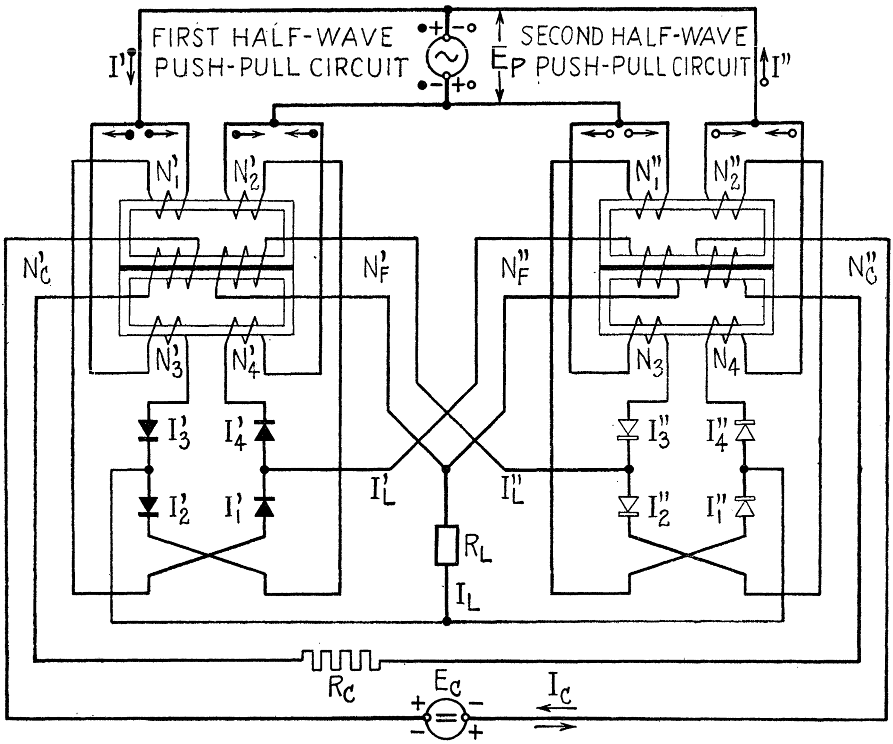
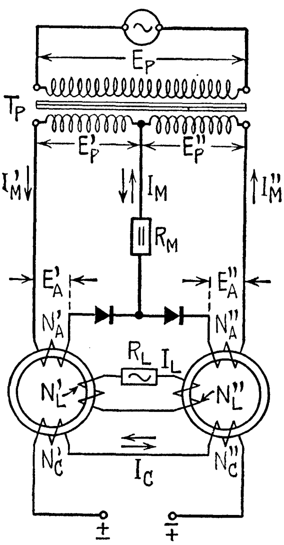


FIG. 13.10. Combination of two half-wave push-pull circuits with a common a-c or d-c load R_L and additional external-feedback windings N'_F, N''_F producing regenerative or degenerative feedback. (*G. Barth, Swiss patent* 233,962, 1942.)

Figure 13.11*a*, a modification of the original self-saturating circuit in Fig. 13.6*b*, illustrates the possibility of supplying simultaneously a d-c load R_M and an a-c load R_L which is series-connected with the secondary windings N'_L, N''_L of two saturable-transformer elements. The pure alternating load voltage $I_L R_L = \text{const} \times (E'_A - E''_A)$, a function of the reversible control current I_C, has 0 or 180° phase displacement with regard to the power-supply voltage E_P. This a-c load voltage may be applied to the input of a push-pull tube circuit or to the control windings of a second push-pull magnetic-amplifier stage, which may have a d-c or a-c load.

Figure 13.11*b* shows another modification of Hudson's half-wave push-pull circuit (Fig. 13.6*b*), supplying only an a-c load R_L. In this example, a single half-wave-rectifier unit is provided, and the saturable-transformer elements are separately biased by means of two a-c bias circuits, so that the a-c load voltage $I_L R_L = \text{const} \times (E'_A - E''_A)$ is zero, if $I_C = 0$. In this case, too, the a-c load R_L represents the input



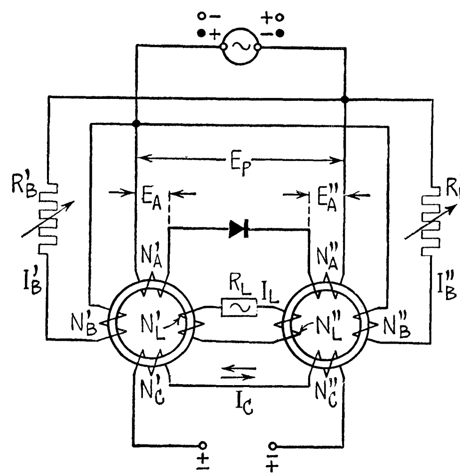


FIG. 13.11. Modifications of the basic half-wave push-pull circuits (*C. S. Hudson*[8]): (*a*) self-saturating saturable-transformer circuit supplying simultaneously a d-c load R_M and an a-c load R_L which is series-connected with the secondary windings N'_L, N''_L of the two saturable-transformer elements; (*b*) self-saturating saturable-transformer circuit supplying only an a-c load R_L.

of a push-pull tube circuit or the control windings of a second push-pull magnetic-amplifier stage with d-c or a-c load.

REFERENCES

1. G. Barth, "Method and a Device for Determining the Magnitudes of Magnetic Fields," United States patent 2,252,059, issued August 12, 1941 (application filed December 24, 1937; in Germany, December 24, 1936), Figure 13.
2. G. Barth, "Means for Measuring Magnetic Fields," United States patent 2,390,-051, issued December 4, 1945 (application filed July 15, 1941), Figures 3 and 4.
3. See "Einrichtung zum Bestimmen der Grössen eines magnetischen Feldes" (Arrangement for measuring the magnitudes of a magnetic field), Swiss patent 224,992, issued December 31, 1942 (application filed February 16, 1942), Figure 4.

4. A. L. Whiteley and L. C. Ludbrook, "Magnetic Amplifier," United States patent 2,229,952, issued January 28, 1941 (application filed October 19, 1938; in Great Britain, June 9, 1938).
5. *Ibid.*, Figure 5.
6. L. R. Crow, "Saturating Core Devices, Operating Principles and Applications," p. 282, The Scientific Book Publishing Company, Vincennes, Ind., 1949.
7. *Ibid.*, pp. 294–295.
8. C. S. Hudson, "Improvements in or Relating to Magnetic Amplifiers," British patent 598,285, issued February 13, 1948 (application filed October 26, 1945), Figure 2 and claim 2.
9. See "Schaltung mit einem Magnetverstärker" (Circuit arrangement with a magnetic amplifier), Swiss patent 233,014, issued June 30, 1944 (application filed May 4, 1942).
10. See M. Bracutt, "Direct Current Control System," United States patent 2,622,-239, issued December 16, 1952 (application filed March 18, 1950).
11. C. W. Lufcy, A. E. Schmid, and P. W. Barnhart, "An Improved Magnetic Servo Amplifier," *AIEE Transactions*, Vol. 71, Part I, pp. 281–289, 1952. Also "Half-wave Magnetic Servo Amplifier," *Electronics*, Vol. 25, No. 8, pp. 124–125, August, 1952.
12. See also H. F. McKenney and H. Kaplan, "Magnetic Amplifier System," United States patent 2,636,150, issued April 21, 1953 (application filed March 30, 1951).
13. See Swiss patent 233,014, Figures 3–23.
14. See "Schaltung mit einem Magnetverstärker" (Circuit arrangement with a magnetic amplifier), Swiss patent 233,962, issued August, 31, 1944 (application filed May 4, 1942), Figures 1–11.
15. C. S. Hudson, British patent 598,285.

CHAPTER 14

Duodirectional Internal-feedback Circuits with Four Saturable-reactor Elements

14.1. Classification of Push-Pull Internal-feedback Circuits with Four Saturable-reactor Elements. The special combinations of two half-wave push-pull circuits with eight rectifier units (Sec. 13.5) present a convenient transition to various other types of duodirectional internal-feedback circuits with two pairs of saturable reactors. These circuits containing only four rectifier units may be classified as follows:

Group 1: Circuits with only one load element
  a. With a-c load
  b. With d-c load
Group 2: Circuits with several load elements
  a. With an a-c load element and additional d-c load elements
  b. With a d-c load element and additional d-c load elements

Both groups, d-c or a-c controlled, may represent either bridge circuits or differential-transformer circuits. This chapter describes some representative examples of such arrangements, which, of course, may be used for designing single-stage or multistage types of magnetic servo amplifiers.

14.2. The Basic Self-saturating Push-Pull Circuit. In 1940, the author[1] described the control of two-phase induction-type reversible motors with a magnetic servo amplifier employing the simple self-saturating push-pull circuit of Fig. 14.1. In this arrangement, one "doubler" circuit (Sec. 11.2*a*) is used for each direction of rotation of the two-phase motor. The alternating load current $I_L = I_L' - I_L''$ flowing through the a-c load R_L (control winding of the separately excited two-phase motor) is a function of the reversible direct control current

$$I_C = \frac{E_C}{R_C}$$

which may be derived from any suitable d-c control circuit. Preferably, a properly rated d-c bias circuit is provided to adjust the quiescent-current values of I'_L and I''_L so that the most favorable operating conditions of the two pairs of saturable-reactor elements will be secured. This d-c bias circuit, similar to that of the nonfeedback arrangement (Fig. 5.14), contains (1) bias windings N'_B, N''_B; (2) a full-wave dry-disk rectifier; (3) fixed series resistors R'_S, R''_S, R'_B, R''_B; and (4) a potentiometer-type resistor R_0 for zero adjustment: $I_L = 0$, if $E_C = 0$. In the arrangement of Fig. 14.1, the four winding units N'_C, N''_C act simultaneously

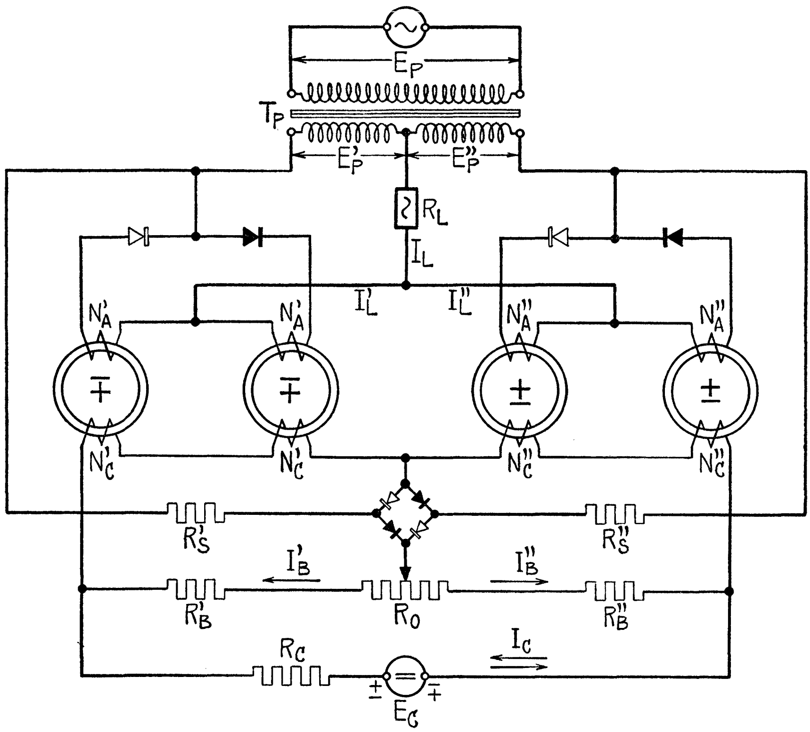


FIG. 14.1. Basic self-saturating push-pull circuit with two pairs of saturable-reactor elements and an a-c load. (*W. A. Geyger.*[1])

as d-c bias windings and d-c control windings, the bias currents I'_B, I''_B and the control currents I_C being superimposed in these winding units.

Evidently, the self-saturating circuit of Fig. 14.1 is much more sensitive than the nonfeedback circuit of Fig. 5.11. F. S. Malick[2] illustrated this fact in a dramatic way by comparing these two types of push-pull circuits in the special case that the d-c components flowing through the control windings of the four saturable-reactor elements are supplied from the plate circuits of two 6AQ5 tubes operating in push-pull.

In the first nonfeedback design (corresponding to the nonfeedback arrangement of Fig. 5.11) "60 ma were required in 22,000 turns to obtain the desired 2-amp motor current. The 6AQ5 tubes which were used required a ± 25-volt d-c signal on the grids." However, "because of the larger power gain of the self-saturating circuit, a control current of 6 ma in 1,200 turns is all that is required for full output. The 5744 subminiature control tubes required only ± 1 volt on the grid. This circuit, in common with the simple reactor circuit, was unaffected by large amounts of ripple on the signal applied to the tube grid."

14.3. Modified Self-saturating Push-Pull Circuits with Center-tapped Power-supply Transformer and Several Load Elements. In 1943, G. Barth[3] described various modified types of self-saturating push-pull circuits, similar to the so-called "ring-modulator" arrangements, which are suitable for supplying several load elements, preferably an a-c load and a twin-type d-c load.

Figure 14.2*a* to *d* illustrates the possibility of using such a bridge circuit with center-tapped power-supply transformer T_P in the following four special cases:

a. D-c control combined with application of an a-c load resistor R_L (carrying pure alternating current I_L) and two separate d-c load resistors R'_L, R''_L (carrying the half-wave current components I'_L and I''_L)

b. A-c control combined with application of a d-c load resistor R_L (carrying full-wave rectified current I_L) and two separate d-c load resistors R'_L, R''_L (carrying the half-wave current components I'_L and I''_L)

c. D-c control combined with application of a d-c load resistor R_L (carrying full-wave rectified current I_L) and two separate d-c load resistors R'_L, R''_L (carrying the half-wave current components I'_L and I''_L)

d. A-c control combined with application of an a-c load resistor R_L (carrying pure alternating current I_L) and two separate d-c load resistors R'_L, R''_L (carrying the half-wave current components I'_L and I''_L)

In any case, the four saturable-reactor elements are properly biased by means of four separate a-c bias circuits, so that each of these elements can be separately adjusted, with regard to the actual quiescent-current values of the four half-wave current components, by means of the associated series resistors R'_B and R''_B. In the arrangements of Fig. 14.2*a* and *c* having series-connected d-c control windings N'_C, N''_C, the fundamental-frequency voltages which are induced from the load windings N'_A, N''_A into the control circuit are always canceling each other completely, not only with no-signal conditions but even when the maximum value of control voltage E_C is applied to the amplifier. However, in the

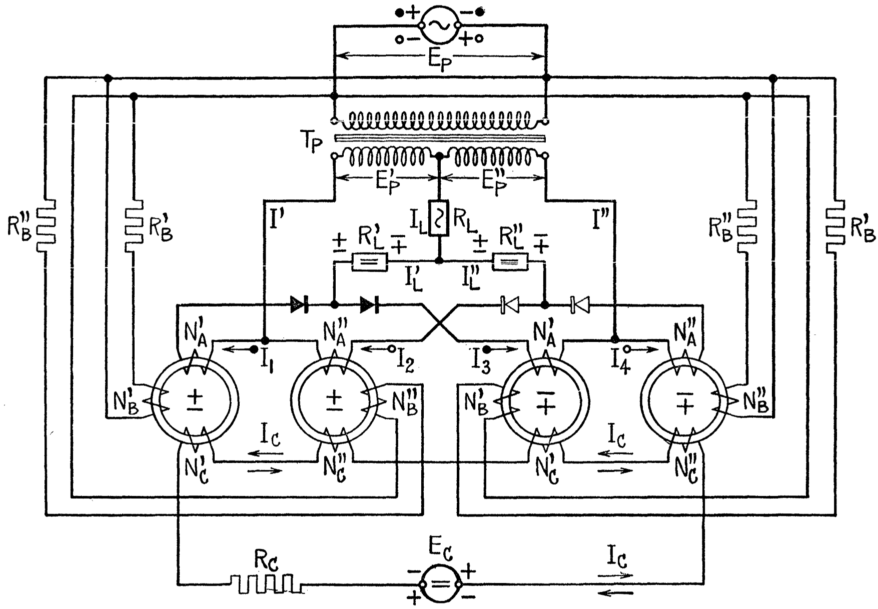


FIG. 14.2*a*. Modified d-c-controlled bridge-type self-saturating push-pull circuit with an a-c load R_L and two additional d-c load resistors R_L' and R_L''. (*G. Barth, Swiss patent* 235,142, 1943.)

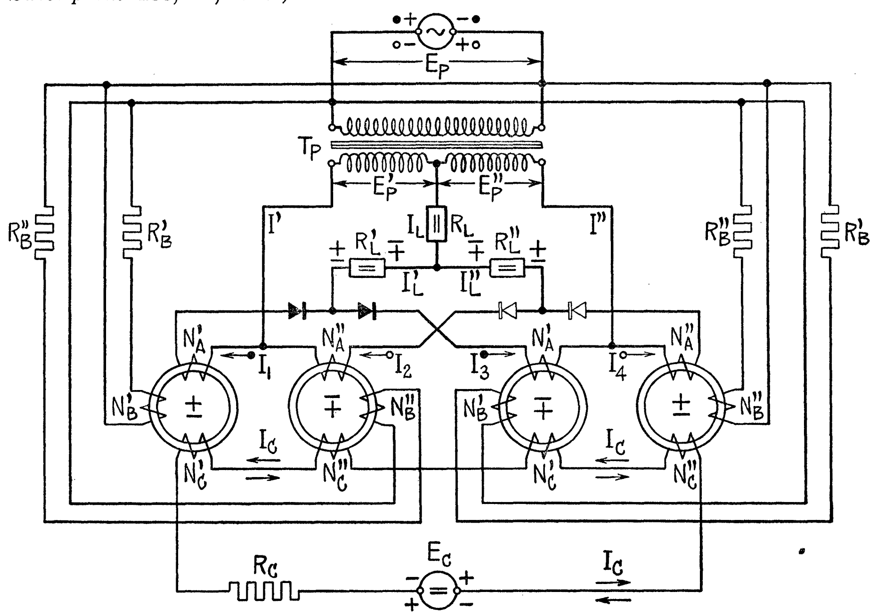


FIG. 14.2*c*. Modified d-c-controlled bridge-type self-saturating push-pull circuit with a d-c load R_L and two additional d-c load resistors R_L' and R_L''. (*G. Barth, Swiss patent* 235,142, 1943.)

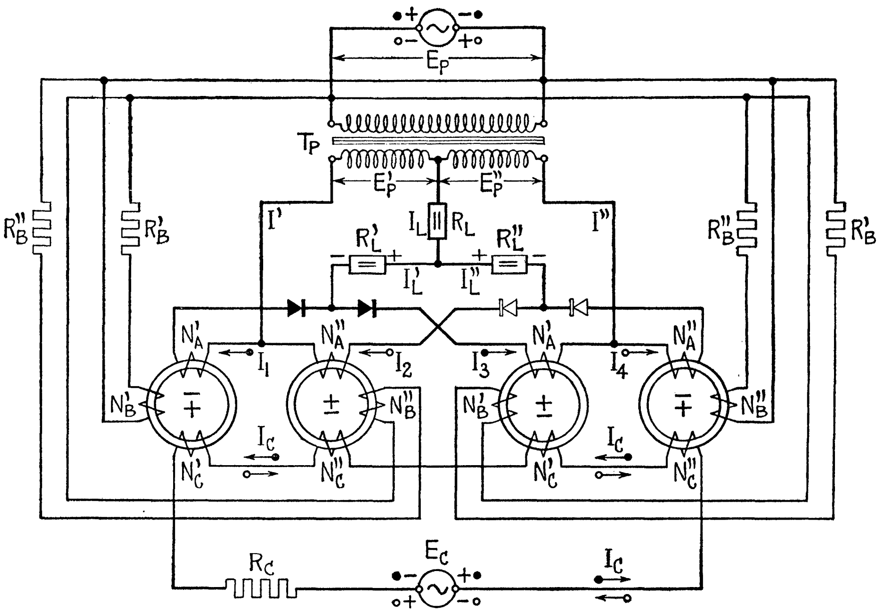


FIG. 14.2*b*. Modified a-c-controlled bridge-type self-saturating push-pull circuit with a d-c load R_L and two additional d-c load resistors R'_L and R''_L. (*G. Barth, Swiss patent* 235,142, 1943.)

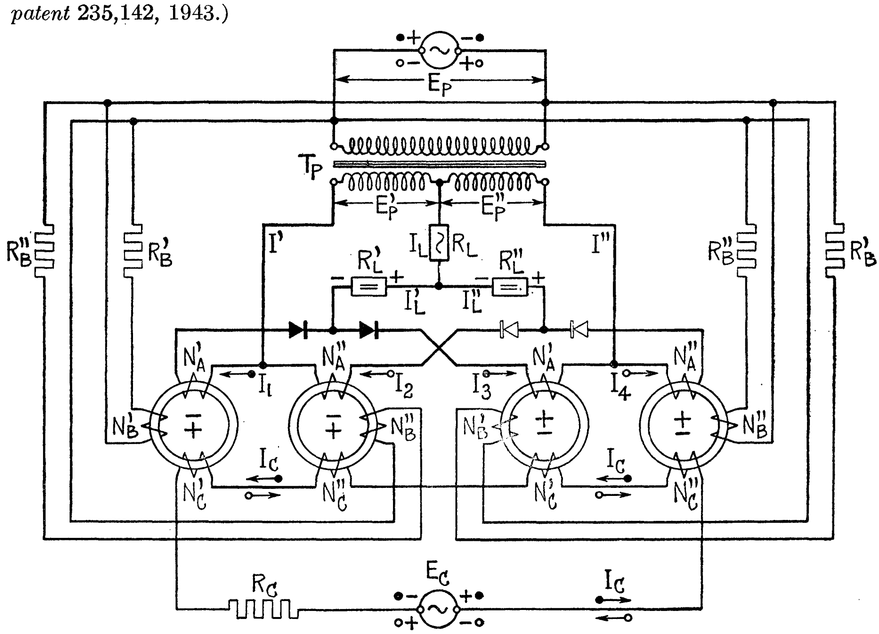


FIG. 14.2*d*. Modified a-c-controlled bridge-type self-saturating push-pull circuit with an a-c load R_L and two additional d-c load resistors R'_L and R''_L. (*G. Barth, Swiss patent* 235,142, 1943.)

arrangements of Fig. 14.2*b* and *d* having series-connected a-c control windings N'_C, N''_C, the fundamental-frequency voltages induced in each of the elements N'_C, N''_C will cancel each other only for no-signal conditions; otherwise fundamental-frequency voltages having considerable magnitudes may be induced in the control circuit. Thus, the actual turns ratio $N'_C/N'_A = N''_C/N''_A$ must have a sufficiently small value (about 0.01 to 0.1). Furthermore, it will be necessary to connect a sufficiently

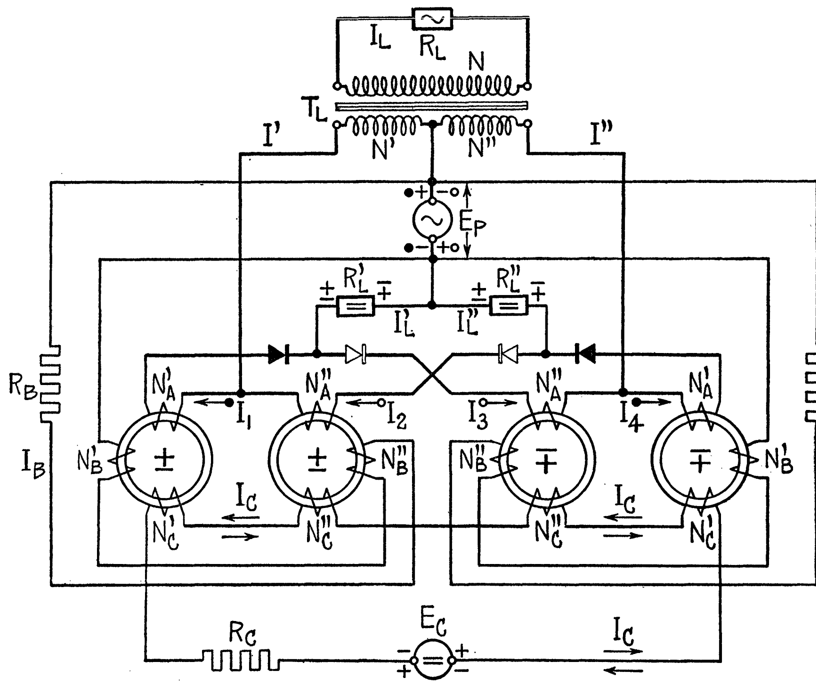


FIG. 14.3*a*. Modified d-c-controlled differential-transformer type of self-saturating push-pull circuit with an a-c load R_L and two additional d-c load resistors R'_L and R''_L. (*G. Barth, Swiss patent* 235,142, 1943.)

high resistance R_C in series with N'_C and N''_C so that the voltages induced from the load windings N'_A, N''_A into N'_C and N''_C will be unable to produce excessively large currents in the control-circuit loop.

14.4. Modified Self-saturating Circuits with Differential-type Load Transformer and Several Load Elements. Figure 14.3 shows other examples of modified self-saturating circuits with d-c control and with a-c control. In these arrangements, the a-c load R_L is connected across the secondary winding of the load transformer T_L with differential-type primary windings, so that a pure alternating current

$$I_L = \text{const} \times (I' - I'')$$

flows through R_L. The two half-wave components I'_L and I''_L flowing through two separate d-c load resistors R'_L, R''_L may be used for other purposes.

In these examples, the two pairs of saturable-reactor elements are biased by means of two a-c bias circuits with bias windings N_B and series resistors R_B. With d-c control, the fundamental-frequency voltages which are induced from the load winding elements N'_A, N''_A into the

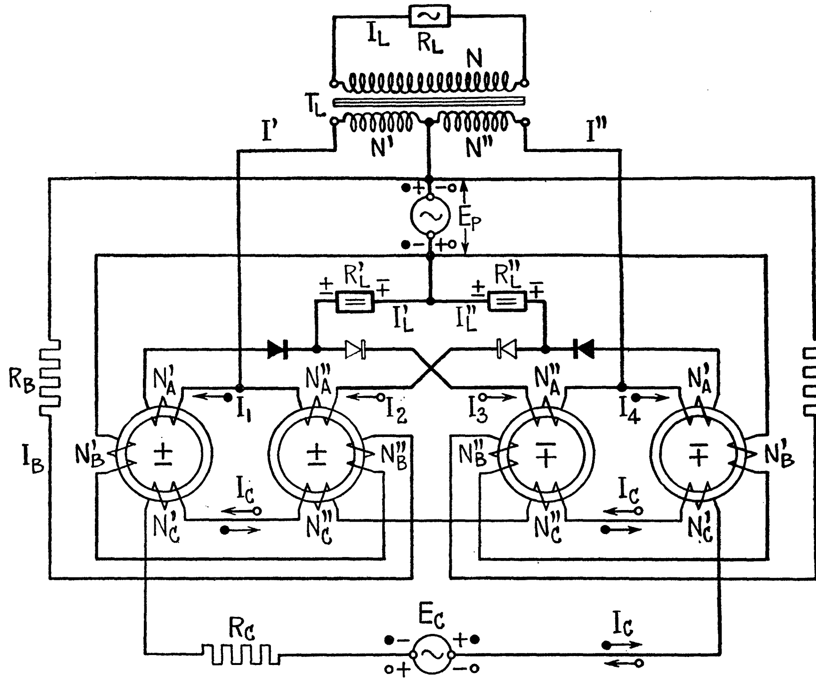


FIG. 14.3*b*. Modified a-c-controlled differential-transformer type of self-saturating push-pull circuit with an a-c load R_L and two additional d-c load resistors R'_L and R''_L. (*G. Barth, Swiss patent* 235,142, 1943.)

associated control winding elements N'_C, N''_C are always canceling each other completely, not only with $E_C = 0$ but even when the maximum value of E_C is applied. However, with a-c control, the turns ratio $N'_C/N'_A = N''_C/N''_A$ must be properly rated, and a sufficiently high series resistor R_C must be provided, so that no excessively large fundamental-frequency currents will be induced in the control-circuit loop.

14.5. Further Development of Duodirectional Self-saturating Circuits with Four Saturable-reactor Elements. Contributions concerning further development and special applications of such circuits have been made by S. E. Hedström,[4] H. Forssell,[5] H. W. Lord,[6] W. L. O. Graves,[7] and H. M. Ogle.[8]

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2. F. S. Malick, "Control of a 65-watt 2-phase Servomotor with a Magnetic Amplifier," *Electrical Engineering*, Vol. 71, No. 5, pp. 466–468, May, 1952.
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CHAPTER 15

Magnetic-amplifier Circuits of the Self-balancing Potentiometer Type

15.1. Introduction. The preceding general comments on feedback in magnetic amplifiers (Sec. 6.2) make it evident that there are several essential differences between the fundamental mode of operation of the various feedback types of magnetic-amplifier circuits and that of the conventional feedback arrangements as used in electronic amplifiers. Generally, with d-c control the feedback circuits, as described in Chaps. 6 to 14, do not have any influence upon the effective input resistance of the control circuit, because the d-c feedback component flows through special feedback windings ("external" feedback) or in the load windings carrying half-wave currents ("internal" feedback). Therefore, the effective impedance of the control circuit is solely determined by the actual total resistance of this circuit, comprising the resistance of the control windings and a series resistor which is conventionally provided to increase the speed of response of the magnetic amplifier.

Feedback arrangements which do not influence the effective input impedance of the control circuit are characterized by the fact that the winding components producing the feedback effects are *inductively* coupled with the control winding components. These arrangements may be termed "pseudo-feedback" circuits, because they do not have the inherent property of influencing the actual magnitude of the input-circuit resistance, as, for example, the electronic amplifier of the cathode-follower type does.

In 1944, the author[1] showed that "real feedback" can be applied to magnetic amplifiers, particularly for instrumentation purposes, by feeding back a voltage or current, proportional to the actual load current, *resistance-coupled* into the control-circuit loop so that a very large amount of degeneration is obtained. In this arrangement (Fig. 5.13*a*), employing a magnetic amplifier of the balance-detector type (Fig. 5.14) in combination with a phase-sensitive electronic amplifier, the actual input voltage across the control windings of the magnetic amplifier is only about 2 per cent of the input voltage to be measured (using 98 per cent degeneration), so that the effective input impedance of the control circuit is about fifty times higher than the actual resistance of the control windings (*e.g.*, effective control-circuit resistance: $6 \times 50 = 300$ ohms). Similar magnetic-amplifier circuits utilizing negative voltage feedback, resistance-coupled into the control-circuit loop and preferably having a "floating" output, have been devised by A. S. FitzGerald.[2-4]

This chapter describes special types of magnetic-amplifier circuits representing various advantageous modifications of the afore-mentioned arrangements, employing no electronic-amplifier components, and having inherent self-balancing potentiometer characteristics.[5]

15.2. Basic Magnetic-amplifier Circuits of the Self-balancing Potentiometer Type. The fundamental principle of these arrangements consists in applying a special compound-feedback circuit: Positive (external or internal) feedback produces an effectively infinite gain that is highly degenerated by negative feedback resistance-coupled into the control-circuit loop. In this way, extreme stability and the fulfillment of balance conditions in the control circuit are obtained. With this system, for either voltage or current input, the balanced control circuit demands practically no energy from the input source.

The basic magnetic-amplifier circuits of the self-balancing potentiometer type contain two separate and equally rated saturable-reactor elements, with high-permeability core material. These simple circuits, however, can be applied only when the voltage or current to be measured will not change polarity, because direction of the output current (polarity of the variable compensating voltage or current) is always the same. Two typical examples of such arrangements will be described in the following paragraphs.

a. Voltage-balance Circuit of the External-feedback Type. Figure 15.1 shows the basic circuit of a self-balancing magnetic amplifier for voltage measurements. The two saturable-reactor elements have series-opposing-connected a-c load windings N_L, series-aiding-connected control windings N_C, series-aiding-connected external-feedback windings N_F, and series-aiding-connected bias windings N_B, so that the voltages induced in each of the d-c twin windings N_C, N_F, N_B by the fundamental wave and by the odd harmonics of power-supply voltage E_P are opposed. The d-c load resistor R_L with an ammeter A_L, a series resistor R_K, and the feedback windings N_F (resistance R_F) with a regeneration-control shunting resistor R_P are supplied by two separate full-wave dry-disk rectifiers with direct currents I_L, both representing the average value of the rectified alternating current I_L flowing through the load windings N_L. Bias windings N_B carrying the bias current I_B and reducing the quiescent-current value of I_L are supplied by a third full-wave dry-disk rectifier with a series impedance Z (nickel-iron-alloy core) and a d-c series resistor R_B. The bias current produces an additional d-c magnetization which is opposing to the d-c magnetization produced by the feedback windings N_F and independent of the control current I_C flowing through the control windings N_C.

Actual control current I_C is proportional to the difference between voltage E_X to be measured and voltage drop E_K across the resistor R_K; it is inversely proportional to the total resistance of the control circuit

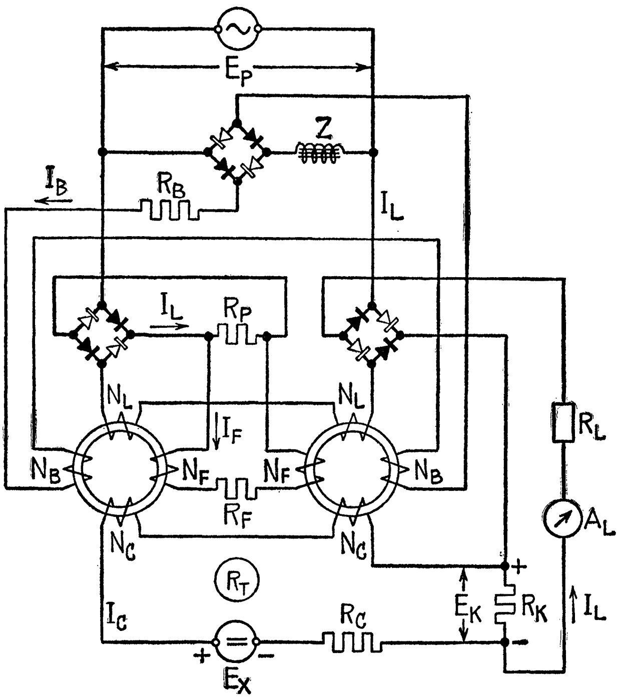

FIG. 15.1. Basic magnetic-amplifier circuit of the self-balancing potentiometer type utilizing external feedback and a control circuit which operates under voltage-balance conditions. (*W. A. Geyger.*[5])

$R_T = R_G + R_C + R_K$, where R_G is the internal resistance of the input source and R_C is the total resistance of the control windings N_C. If I_C is very small in comparison with I_L, actual control current will be

$$I_C = \frac{E_X - E_K}{R_T} = \frac{E_X - I_L R_K}{R_G + R_C + R_K} \tag{15.1}$$

With $E_X = 0$ the very small quiescent-current value of I_L causes a voltage drop $E_K = I_L R_K$ across resistor R_K, and a very small "negative" control current $I_C = -I_L R_K / R_T$ is flowing through the control windings

N_C. Note that this circuit assumes a conduction path through the voltage being measured, so that this negative control current producing negative (degenerative) feedback can flow through the galvanically coupled control-circuit-loop resistance R_T.

The actual or measurable load current I_L is the resultant of the "exciting-current component" I_{L0}, which is practically independent of the control current I_C and of the "load component" I_{LC} flowing in response to changes in control current I_C. Assuming ideal rectifier performance, the positive (regenerative) feedback ampere-turns aiding those provided by the control circuit will be $I_F N_F = I_L N_F R_P/(R_P + R_F)$, and the negative (degenerative) feedback ampere-turns of the control windings N_C will be $I_C N_C = I_L N_C R_K/R_T$.

According to the current-transformer properties of saturable-reactor elements with high-permeability core material (Sec. 3.3), total ampere-turns of load windings N_L are equal to total ampere-turns of control windings N_C and feedback windings N_F:

$$I_L N_L = I_C N_C + I_F N_F = \frac{E_X - I_L R_K}{R_T} N_C + I_L N_F \frac{R_P}{R_P + R_F} \tag{15.2}$$

When a voltage E_X is applied to the control circuit, the load current I_L increases, so that a voltage drop $E_K = I_L R_K$ takes place across resistor R_K. Because the polarity of E_K opposes an increase in load current I_L, the resistor R_K exercises a controlling effect on the load current I_L.

Referring to Eq. (15.2), the relationship between E_X and I_L will be given by the equation

$$\frac{E_X}{I_L} = R_K + R_T \frac{N_L - N_F R_P/(R_P + R_F)}{N_C} \tag{15.3}$$

Thus using positive (regenerative) external feedback with

$$\frac{N_F R_P}{(R_P + R_F)} = N_L$$

we get

$$\frac{E_X}{I_L} = R_K \qquad \text{or} \qquad \frac{I_L}{E_X} = \frac{1}{R_K} \tag{15.4}$$

where $1/R_K$ represents the "transconductance" of the circuit.

Referring to Eq. (15.4), it is evident that, with $N_F R_P/(R_P + R_F) = N_L$, the voltage drop $I_L R_K$ across resistor R_K will be equal to the voltage E_X to be measured, and the control current $I_C = (E_X - I_L R_K)/R_T$ will be zero.

The afore-mentioned condition that $N_F R_P/(R_P + R_F) = N_L$ involves the assumption of an ideal feedback-rectifier characteristic, *i.e.*, a direct current $I_L = I_F(R_P + R_F)/R_P$ is the average value of the rectified alternating current I_L flowing through the load windings N_L. In practice, however, departure of the feedback rectifier from the ideal characteristic causes a certain reduction of current ratio I_F/I_L so that actual feedback ampere-turns I_FN_F are also reduced. By proper adjustment of the shunting resistor R_P acting as regeneration control, it is possible to compensate this effect of feedback-rectifier imperfections.

If the control windings N_C are temporarily made ineffective, *e.g.*, by disconnecting the input source R_G ($R_T = \infty$), and if the regeneration-control resistor R_P is increased, a point of instability will be reached where the well-known trigger action due to excessive positive feedback occurs. At this point the circuit is critically regenerated and is essentially a magnetic amplifier of infinite gain. If the control windings N_C are then reconnected to the degenerative coupling resistor R_K, all tendency toward instability and trigger action will be restrained by superimposed degeneration, and the magnetic amplifier will operate in the usual way under extremely stable conditions. It is interesting to note that the regeneration-control resistor R_P may be further increased and the magnetic amplifier operated in the seemingly impossible region of greater than infinite (negative) internal gain.

With normal working conditions corresponding to adjustment of R_P to the point of "critical regeneration," the control current

$$I_C = \frac{E_X - I_LR_K}{R_T}$$

is zero, and the control windings N_C exert predominantly a transient type of control upon the two saturable-reactor elements, while the regenerative external-feedback windings N_F and the a-c load windings N_L do the actual work of establishing the output current I_L. If the control current I_C is observed by means of a microammeter while an input voltage E_X is applied, no change is noticeable. Therefore, this voltage-balance condition of zero control current I_C may be used in practice as the criterion for adjustment of the regeneration-control resistor R_P: If R_P is adjusted below or above the point of critical regeneration while the control circuit is subjected to variations of input voltage E_X, a control current I_C will appear, but the direction of current flow will be dependent upon the direction of actual misadjustment.

For practical applications of a magnetic amplifier as shown in Fig. 15.1, the unique feature of practically zero control current I_C for the entire working range of output current I_L means a materially more linear relationship between voltage E_X and current I_L, as represented by Eq. (15.4). The self-balanced control circuit demands practically no energy from the input source because, even with the optimum input-circuit matching conditions, *i.e.*, with

$$R_C + R_K = R_G = \frac{R_T}{2} \tag{15.5}$$

the effective input resistance E_X/I_C is theoretically infinite and is practically about 100 to 1,000 times higher than $R_C + R_K$. Of course, this fact is particularly valuable for applications where a very high input-circuit-resistance value of the magnetic amplifier is of paramount importance.

Another valuable property of the circuit of Fig. 15.1 is the fact that over certain ranges large variations in power-supply voltage E_P and load resistance R_L have a comparatively small effect upon the ratio E_X/I_L. Indeed, consideration of the results of various measurements on such arrangements[5] leads to the conclusion that the actual behavior of a self-balancing potentiometer type magnetic amplifier approximates very closely corresponding characteristics of an "ideal amplifier" having a constant transconductance $I_L/E_X = 1/R_K$, just as does an electronic amplifier of the compound-feedback type (R. W. Gilbert[6]).

Of course, in accordance with common practice in the field of d-c amplifiers, some variations of the degeneration network are possible. For example, a voltage divider may be provided in this network, so that the "voltage ratio" between load voltage $E_L = I_L R_L$ and voltage E_X is given by the actual resistance ratio of the voltage divider.

b. Current-balance Circuit of the Internal-feedback Type. Figure 15.2 shows the basic circuit of a self-balancing magnetic amplifier for current measurements. This arrangement uses self-saturation, an a-c bias circuit (similar to Fig. 7.4*b*), and additional external-feedback ("boosting") windings N_D to provide critical regeneration by proper adjustment of the regeneration-control resistor R_D. In this case,

$$\frac{I_L}{I_X} = \frac{R_N + R_M}{R_N} \tag{15.6}$$

represents the "current ratio" of the circuit.

Referring to Eq. (15.6), it is evident that with critical regeneration the compensating current $I_K = I_L R_N/(R_N + R_M)$ will be equal to the current I_X to be measured, and the control current $I_C = I_X - I_K$ will be zero. In this case, the control windings N_C exert predominantly a transient type of control upon the saturable-reactor elements, while the load windings N_A of the self-saturating circuit and the external-feedback (boosting) windings N_D do the actual work of establishing the output current I_L. If the control current I_C is observed by means of a microammeter while an input current I_X is applied, no change is noticeable. Therefore, this current-balance condition of zero control current I_C may be used in practice as the criterion for adjustment of the regeneration-control resistor R_D: If R_D is adjusted below or above the point of critical regeneration while the control circuit is subjected to variations of input current I_X, a control current I_C will appear, but the direction of current flow will be dependent upon the direction of actual misadjustment.

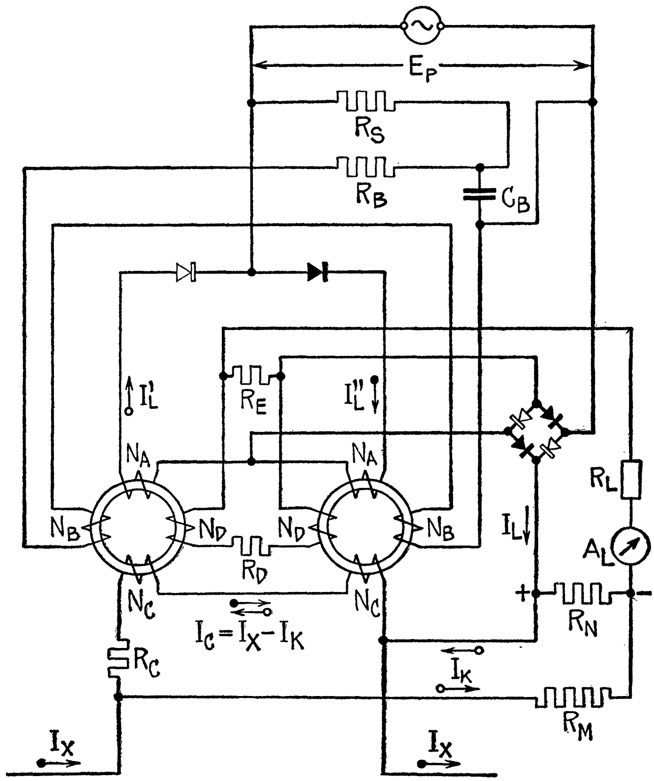

FIG. 15.2. Basic magnetic-amplifier circuit of the self-balancing potentiometer type utilizing internal feedback and a control circuit which operates under current-balance conditions. (*W. A. Geyger*.[5])

For practical applications of a magnetic amplifier as shown in Fig. 15.2, the unique feature of practically zero control current I_C for the entire working range of output current I_L means a materially more linear relationship between the currents I_X and I_L, as represented by Eq. (15.6). The self-balanced control circuit demands practically no energy from the circuit under test carrying the direct current I_X, because, even with optimum input-circuit matching conditions, the effective input voltage $(I_X - I_K)R_C$ is theoretically zero and is practically about

100 to 1,000 times smaller than $I_X R_C$. Evidently, this fact is particularly important for those applications where a very small input resistance value of the magnetic amplifier is of paramount importance.

Another property of the circuit of Fig. 15.2 is the fact that over certain ranges large variations in power-supply voltage E_P and load resistance R_L have a comparatively small effect upon the ratio I_X/I_L. Considering results of various measurements on such arrangements containing saturable-reactor elements with Orthonol tape cores, it was learned that the actual behavior of the amplifier investigated approximates very closely corresponding characteristics of an "ideal amplifier" having a constant current ratio $I_L/I_X = (R_N + R_M)/R_N$.

In this case, too, some variations of the degeneration network are possible. For example, a series resistor may be provided in this network, the actual resistance value of which (R_S) represents the "transresistance" (the ratio between load voltage $E_L = I_L R_L$ and input current I_X) of this magnetic amplifier.

Furthermore, it is possible to apply the fundamental principle of self-balancing compound-feedback circuits (Figs. 15.1 and 15.2) using critical regeneration for the design of push-pull-type magnetic amplifiers, as described in Sec. 14.3. Experimental investigations[5] have shown that excellent results can be obtained with such arrangements operating with either voltage-balance or current-balance conditions.

15.3. Derivative-feedback Circuits of Self-balancing Magnetic Amplifiers. An efficient method to reduce the response time of magnetic amplifiers consists in providing "derivative feedback" (Sec. 9.4). This method is based upon the fact that actual magnitude of the feedback effect in this case is a function of time, and a condenser or a reactor acting as a time-variable impedance is provided in the "derivative-feedback circuit." Using this method, it is possible to reduce the response time to about one-fifth its original value.

Evidently, with "critical regeneration" the control current is zero, and the control windings exert predominantly a transient type of control upon the two saturable-reactor elements, while the regenerative-feedback windings and the load windings do the actual work of establishing the output current.

It is possible to alter materially the transient-response characteristics of a self-balancing magnetic amplifier through the use of derivative feedback. This may be accomplished by providing a time-variable impedance (a condenser or a reactor) in the degeneration network in such a way that a delayed degenerative-feedback effect is obtained.

Figure 15.3 shows several typical examples for using a time-variable impedance in the degeneration network of a self-balancing magnetic amplifier working under voltage-balance conditions, as illustrated in

Fig. 15.1. A condenser (Fig. 15.3*a*) or a linear reactor (Fig. 15.3*b*) is used in connection with several resistors R_1, R_2, R_K to obtain a delayed-feedback action of the compensating voltage E_K. Figure 15.3*c* illustrates another possibility for providing delayed negative feedback by means of a properly rated mutual inductance M, which consists of a simple air-gap-core transformer. Load current I_L flows through the primary windings, and transient voltage E_T representing the derivative of I_L is opposing the actual voltage drop I_LR_K, so that delayed-feedback action of compensating voltage $E_K = I_LR_K - E_T$ will be obtained.

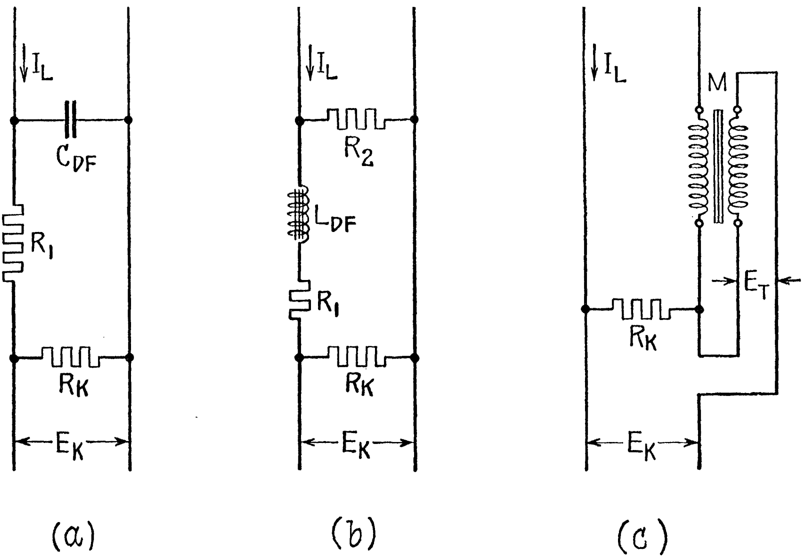

FIG. 15.3. Three typical examples of derivative-feedback circuits, as used in self-balancing magnetic amplifiers: (*a*) circuit with a condenser C_{DF} acting as a time-variable impedance; (*b*) circuit with a linear reactor L_{DF} acting as a time-variable impedance; (*c*) circuit with a mutual inductance M for providing delayed negative feedback.

Similar degeneration networks containing time-variable impedances or a mutual inductance can be used in connection with self-balancing magnetic amplifiers working under current-balance conditions, as illustrated in Fig. 15.2. Of course, equivalent arrangements may be applied in various other types of magnetic-amplifier circuits.

15.4. Practical Applications of Magnetic Amplifiers of the Self-balancing Potentiometer Type. In order to describe the various applications of these magnetic amplifiers in the fields of electrical measurements, telemetering, and automatic control and for the operation of small d-c or a-c reversible motors working in connection with servomechanisms, it is important to stress the following facts:

1. Self-balancing magnetic amplifiers operating under voltage-balance or current-balance conditions of the control circuit demand practically no energy from the d-c input source producing the voltage or current to be measured.
2. The effective input resistance of a self-balancing magnetic amplifier for voltage measurements is practically about 100 to 1,000 times higher than the actual input-circuit-loop resistance.
3. The effective input resistance of a self-balancing magnetic amplifier for current measurements is practically about 100 to 1,000 times smaller than the actual resistance of the control windings.
4. The unique feature of a considerably reduced control current for the entire working range of the self-balancing magnetic amplifier means a materially more linear relationship between the input and output quantities.
5. Over certain ranges large variations in power-supply voltage, load resistance, and control-circuit resistance have a comparatively small effect upon the "transconductance," "voltage ratio," "current ratio," or "transresistance" of self-balancing magnetic amplifiers operating under extremely stable conditions with a high over-all power gain.
6. For certain applications special compensating effects can be applied by designing the degeneration network in such a way that resistance values of this network are a function of voltage, current, temperature, or time (derivative-feedback effects, *e.g.*, delayed degenerative feedback).
7. Results of various measurements[5] lead to the conclusion that the actual behavior of self-balancing magnetic amplifiers approximates very closely corresponding characteristics of an "ideal amplifier," just as an electronic amplifier of the compound-feedback type (R. W. Gilbert[6]) does.

a. Measurement of Electrical Quantities. Magnetic amplifiers of the self-balancing potentiometer type are particularly suitable for extending the sensitivity of direct-indicating electrical (multirange) instruments and ink recorders making a permanent record of very small direct voltages and currents, such as those created, *e.g.*, by thermocouples, photoelectric cells, or radiation pyrometers. Such amplifiers can also be used in connection with phase-sensitive rectifiers for the measurement of various a-c quantities, *e.g.*, in a-c potentiometer circuits and a-c bridge networks. Voltage dividers and shunting arrangements may be provided to extend the voltage and current ranges of such amplifiers.

b. Measurement of Nonelectrical Quantities. In accordance with common practice, nonelectrical quantities will be measured by means of "transducers" providing the desired transition into an electrical value. There are numerous kinds of transducers representing variable resistors, inductances, mutual inductances, condensers, etc.[7]

c. Special Applications. Referring to Fig. 15.1, it is evident that the self-balancing magnetic amplifier demands practically no energy from a thermocouple (R_G), because the effective input resistance (E_X/I_C) is practically about 100 to 1,000 times higher than the actual resistances of the control-circuit loop. Therefore, comparatively large changes in resistance of the thermocouple or other resistors contained in the control circuit will have no appreciable effect upon the accuracy of measurement.

Photoelectric cells of the selenium barrier-layer dry-disk type find wide application in the measurement and control of many industrial processes, including both mechanical and chemical. When these cells are used in circuits of low external resistance, the variation of the photoelectric current with respect to the illumination is practically linear. Referring to Fig. 15.2, it is evident that the self-balancing magnetic amplifier demands no energy from a photocell producing the input current I_X, because even with optimum input-circuit matching conditions, the effective external resistance is practically about 100 to 1,000 times smaller than the actual resistance value of the control windings of the amplifier. Therefore, the voltage across the photoelectric cell will be practically zero, and the relationship between current and illumination is then substantially linear.

Although such photoelectric cells generally develop ample current for direct operation of moving-coil-type instruments and sensitive relays without the use of amplifiers, it will be necessary in special cases to provide a simple and reliable amplifier, preferably d-c, for operation of moving-coil-type ink recorders. Evidently, it is possible to extend the sensitivity of this type of ink recorder. This will be useful in many other cases, especially for the purpose of telemetering.

Power measurements can be made at high frequencies by dissipating the power in a bolometer, which is a resistive element with a large temperature coefficient of resistance. The magnitude of this power can be determined from the measured change in resistance of a bolometer bridge operating in connection with a self-balancing magnetic amplifier and a moving-coil-type ink recorder.

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CHAPTER 16

Second-harmonic-type Magnetic-amplifier Circuits

16.1. Introduction. The fundamental-frequency magnetic-amplifier circuits dealt with in Chaps. 3 to 15 are characterized by the fact that the frequency of the output current is that of the fundamental wave of power-supply voltage (E_P). This output current may be either an alternating current or a unidirectional current obtained by means of full-wave or half-wave rectification. These arrangements, in their various applications, act as power amplifiers, flux-gate magnetometers, current- or voltage-measuring devices, or, in special cases, also as magnetic modulators (Sec. 5.4*c*).

It has already been shown (Secs. 4.2 and 4.3) that the actual operating conditions of the fundamental-frequency circuits may be altered materially by the effects of double-frequency currents, because of the superposition of a-c and d-c magnetization. In these arrangements, however, the second-harmonic components are generally not utilized for supplying the load power to the output circuit.

Besides the fundamental-frequency circuits, there are other arrangements which utilize the second-harmonic component of the output, so that the frequency of the load voltage is twice the fundamental-wave frequency of power-supply voltage (E_P). With this operating principle, the saturable-reactor device acts more as a magnetic modulator or voltage amplifier than as a power amplifier.

This chapter describes typical examples of second-harmonic-type magnetic-amplifier circuits, which have proved to be especially useful for their ability to convert very small d-c signals into a-c voltages with a high degree of stability. In such arrangements, the saturable-reactor elements may have a "closed" magnetic circuit (magnetic modulators) or an "open" magnetic circuit (flux-gate magnetometers).

16.2. The Fundamental Principle of Second-harmonic-type Saturable-reactor Devices. In 1902, J. Epstein[1] showed that superposition of a-c and d-c magnetization in transformerlike structures offers the possibility of producing an alternating current having a frequency which is twice the frequency of the a-c generator supplying the circuit. Such an arrangement represents a second-harmonic-type frequency transformer, sometimes called "static frequency doubler," or "second-harmonic generator." Some modifications of Epstein's basic method of frequency transformation have been described by J. M. A. Joly[2] (1911) and G. Vallauri[3] (1912). In 1914, L. Dreyfus[4] gave an analysis of the

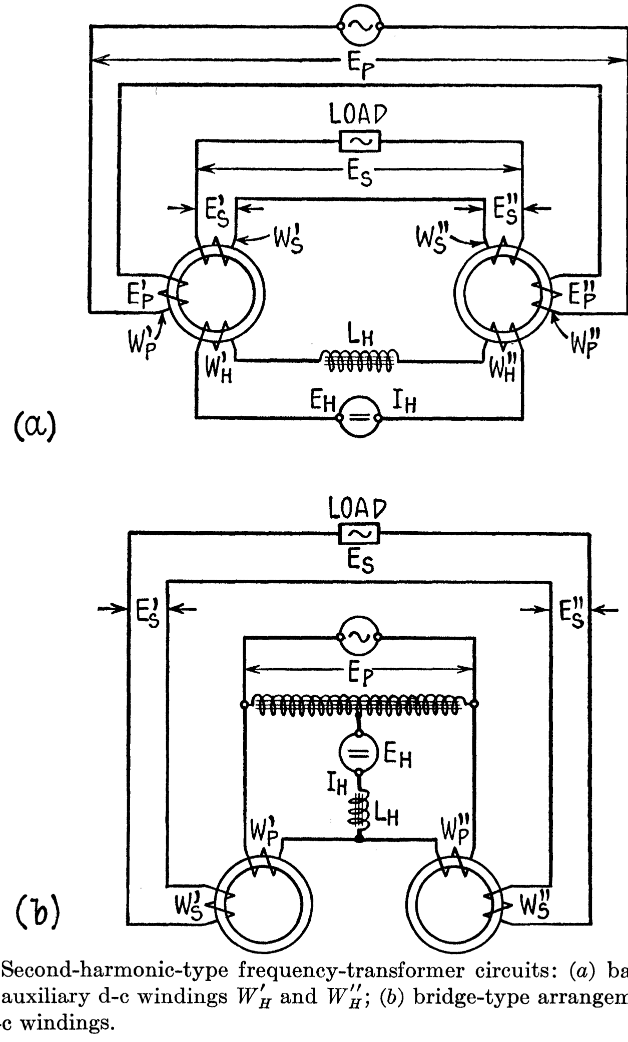


FIG. 16.1. Second-harmonic-type frequency-transformer circuits: (a) basic arrangement with auxiliary d-c windings W'_H and W''_H; (b) bridge-type arrangement without separate d-c windings.

static frequency doubler, describing the fundamental mode of operation of second-harmonic-type saturable-reactor devices, as used in magnetic modulators and flux-gate magnetometers.

Figure 16.1 shows two examples of second-harmonic-type frequency transformers. Two equally rated saturable-reactor elements acting as transformers, and preferably having high-permeability core material, are used in both cases. The series-aiding-connected primary windings W'_P, W''_P are supplied from an a-c generator (voltage E_P, frequency f_P); and the series-opposing-connected secondary windings W'_S, W''_S apply to the load an alternating voltage $E_S = E'_S - E''_S$ having the frequency $2f_P$.

In the arrangement of Fig. 16.1*a*, a substantially constant direct current I_H derived from a d-c generator (*e.g.*, battery voltage E_H) flows through the auxiliary windings W'_H, W''_H and a choke coil which prevents second-harmonic current components from flowing through the d-c circuit. In the bridge-type arrangement of Fig. 16.1*b* having no additional d-c windings, the direct voltage E_H is applied to the primary windings W'_P, W''_P and to the center-tapped autotransformer. In this case, too, a choke coil L_H must be provided so that the second-harmonic-frequency current, proportional to the secondary voltage $E_S = E'_S - E''_S$, will flow only through the load and will not be short-circuited by the d-c generator (E_H).

The operation of these arrangements is based upon the differential magnetization of the two saturable-transformer elements by the d-c ampere-turns acting in conjunction with the a-c ampere-turns of the primary windings W'_P, W''_P. When no direct current is applied to the circuit ($I_H = 0$), then the equal and opposite voltages E'_S, E''_S induced in the secondary windings W'_S, W''_S are identical in waveshape and balance each other in their effect upon the load circuit (Fig. 16.2). This is true in spite of changes in magnitude and frequency of power-supply voltage E_P, since the equal and opposite secondary voltages E'_S and E''_S are always symmetrical, although they may not be sinusoidal and may contain comparatively large odd-harmonic components.

However, when a direct current I_H producing a d-c flux Φ is applied, the voltage waves E'_S, E''_S become asymmetrical, and a second-harmonic voltage $E_S = E'_S - E''_S$ appears on the load (Fig. 16.2). It is to be noted that the phase of E_S reverses when the direction of the direct current (the d-c flux $\pm\Phi$) is reversed. Thus, the second-harmonic-type magnetic-amplifier circuits are inherently polarized, so that duodirectional transfer characteristics (Fig. 5.1*c*) will be obtained without providing additional d-c bias ampere-turns. The direct current I_H produces a unidirectional magnetomotive force, which aids that of the first primary winding W'_P and opposes that of the second primary winding W''_P (or vice versa) during the half-cycle periods of power-supply voltage E_P. As a result, highly asymmetrical voltage waves E'_S, E''_S are induced in the corresponding secondary windings W'_S, W''_S, and a large second-harmonic voltage E_S will be induced in the load circuit.

In any case, asymmetrical voltage waves, as shown in Fig. 16.2, are indicative of the presence of even harmonics, principally the second. Since the second-harmonic components of the voltage waves E'_S, E''_S repeat themselves every half cycle of the fundamental wave, the fundamental and the odd-harmonic components of E'_S and E''_S are caused to balance each other, but the second-harmonic components are caused to

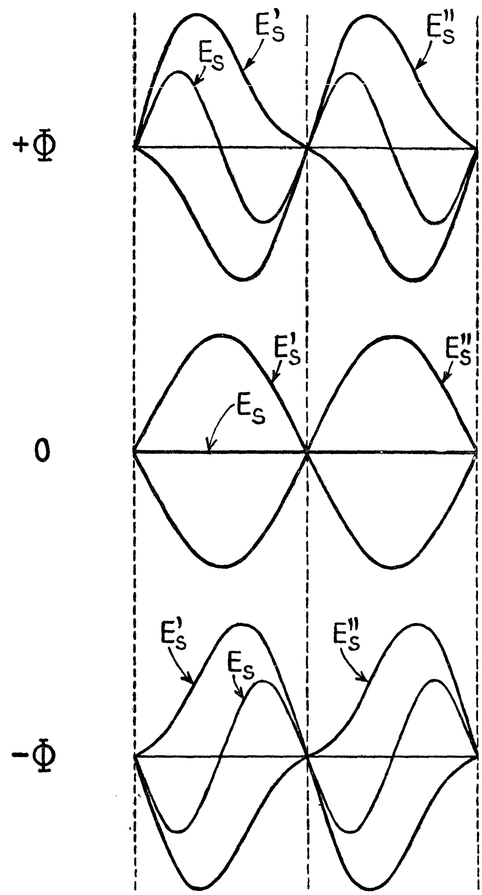


FIG. 16.2. Waveshapes of the secondary voltages E'_S, E''_S, and $E_S = E'_S - E''_S$, without and with the effect of a d-c flux $\pm\Phi$.

act cumulatively and produce a reversible double-frequency output voltage. This voltage $E_S = E'_S - E''_S$ is, within certain limits, proportional to the actual magnitude of the d-c flux Φ or direct current I_H, respectively. The second-harmonic-type magnetic modulators and flux-gate magnetometers are based upon the detection or measurement of this reversible voltage E_S, preferably by means of (1) a moving-coil galvanometer operating in connection with a phase-sensitive rectifier,

(2) a separately excited wattmeter, or (3) a two-phase induction-type reversible motor, which is separately excited by a substantially constant current having twice the frequency of the power-supply voltage E_P.

16.3. Second-harmonic-type Magnetic Modulators. Conversion of low-voltage d-c signals into a-c voltages with a high degree of stability can be accomplished through the use of a second-harmonic-type magnetic modulator containing no rectifiers in the saturable-reactor bridge circuit.

Figure 16.3 shows two examples of such arrangements. In both cases, the magnetic-modulator circuit is used for controlling an electronic a-c

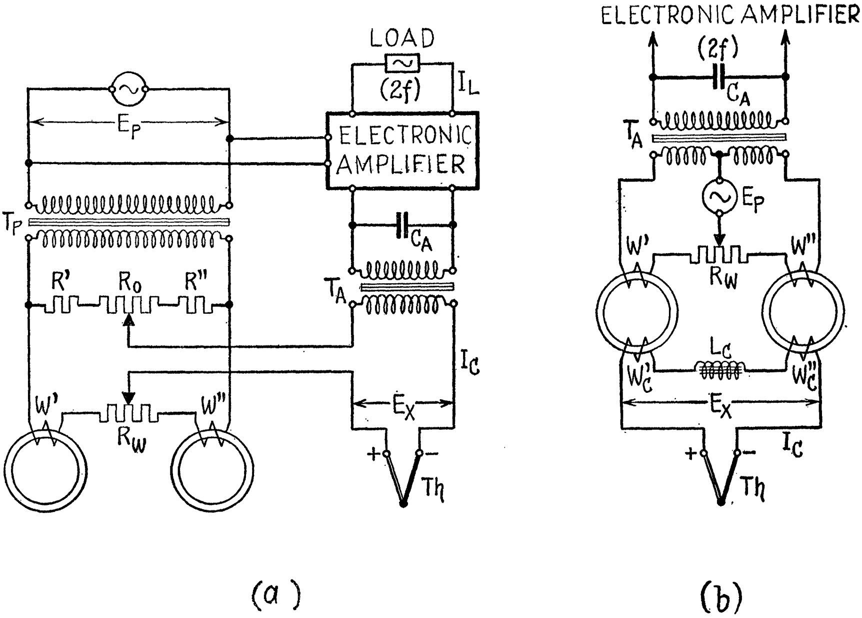


FIG. 16.3. Second-harmonic-type magnetic-modulator circuits: (*a*) bridge-type circuit; (*b*) differential-transformer circuit.

amplifier, unaffected by drift in d-c level and introducing only errors due to change of gain, which can easily be eliminated by application of negative feedback. In these arrangements, especially suitable for operation of high-speed electronic recording potentiometers, the direct millivoltage created by a thermocouple *Th* or a radiation pyrometer is changed to an alternating voltage of proportional magnitude in a saturable-reactor bridge circuit acting as a d-c/a-c converter. This alternating voltage may readily be amplified by an ordinary a-c amplifier. Since an a-c amplifier is inherently more stable than a d-c amplifier, it is comparatively easy to maintain a very constant adjustment of the system if such a magnetic modulator is used.

The bridge-type circuit of Fig. 16.3*a* contains:

1. Two equally rated saturable-reactor elements with high-permeability core material and a-c windings W', W'' acting simultaneously as d-c control windings and carrying the direct control current I_C
2. Two fixed resistors R', R'' and two adjustable potentiometer-type resistors R_0 and R_W
3. A small power-supply transformer T_P (*e.g.*, $^{115}/_{10}$ volts)
4. The conventional type of a-c amplifier
5. An input transformer T_A with a condenser C_A tuned to the frequency of the second harmonic
6. The load carrying the reversible alternating current I_L (phase 0 or 180°), which is a function of the small d-c voltage E_X to be measured

Figure 16.3*b* illustrates the possibility of providing a differential-type magnetic-modulator circuit which contains:

1. Two equally rated saturable-reactor elements with high-permeability core material, a-c windings W', W'', and separate d-c control windings W'_C, W_C with a series-connected choke coil L_C
2. A differential transformer T_A with a condenser C_A tuned to the frequency of the second harmonic
3. A potentiometer-type resistor R_W for zero adjustment

In this case, too, the magnetic-modulator circuit is followed by an electronic a-c amplifier controlling separately excited phase-sensitive devices.

Figure 16.4 shows the complete circuit of a d-c balance detector operating in connection with d-c bridge or potentiometer circuits. According to the author's investigations, this arrangement may replace a mirror galvanometer in many cases. The main advantage of such a device is the fact that the response time of the pointer-type microammeter M in the output circuit is much lower than that of a mirror galvanometer. Consequently, the time necessary for balancing the bridge or potentiometer circuit can be considerably reduced in this way.

The arrangement of Fig. 16.4 contains:

1. A second-harmonic-type magnetic modulator with two saturable-transformer elements having the series-connected primary windings W'_P, W''_P, a common secondary winding W_S, a common d-c control winding W_C, and a choke coil L_C in the d-c control circuit (control current I_C)
2. A bridge-type phase-sensitive rectifier circuit comprising two triode elements, a cathode resistor R_K with shunt condenser C_K, two grid resistors R'_G, R''_G, two plate resistors R', R'', a potentiometer-type

resistor R_0 (zero adjustment), and the moving-coil microammeter (current range preferably +25 to 0 to −25 μa)

3. An auxiliary plate-supply circuit with power-supply transformer T_A, two half-wave-rectifier sets, and a fixed resistor R_A (plate-supply voltage E_A)

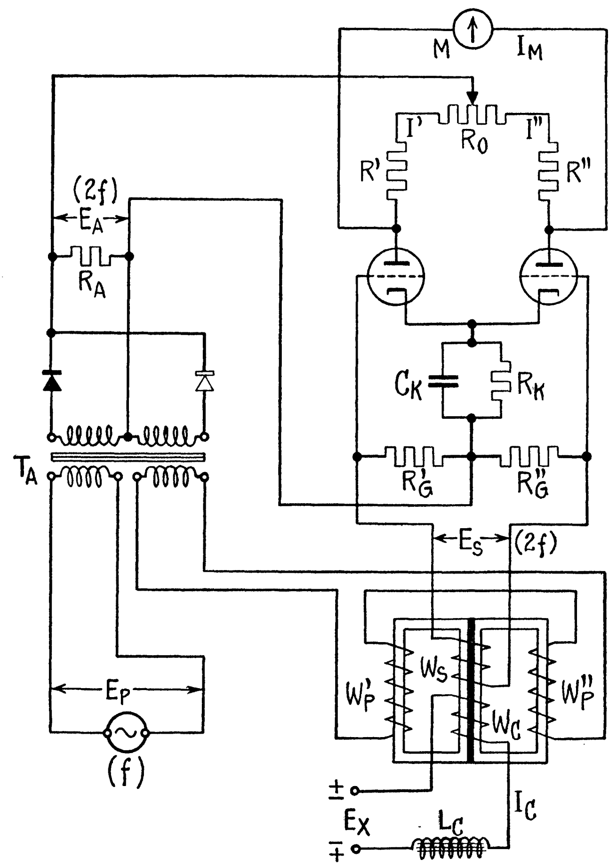


FIG. 16.4. Complete circuit of a d-c balance detector operating in connection with d-c bridge or potentiometer arrangements.

It is to be noted that the phase-sensitive push-pull tube circuit is separately excited by the plate-supply voltage E_A, which contains a d-c component and a superimposed a-c component having twice the frequency of the power-supply voltage E_P. Thus, the secondary voltage $E_S = f(E_X)$ produces a strong response in the frequency-selective push-pull tube circuit.

In many cases, excellent results have been obtained with this simple arrangement. Its sensitivity, of course, may be considerably increased by connecting a conventional electronic a-c amplifier between the secondary winding W_S of the magnetic modulator and the grid circuit of the phase-sensitive rectifier device.

Reference to the literature shows that considerable effort has been expended in the study and development of second-harmonic-type magnetic modulators. In 1935, C. W. La Pierre[5] described several types of such devices and various practical applications. J. F. Coales[6] has stressed that "all magnetic amplifiers are, in fact, modulators, the input being direct current and the output before rectification alternating current." He has classified the different kinds of "amplifiers employing a modulated magnetic field" and also presented a tabulation of the various typical properties of such arrangements.

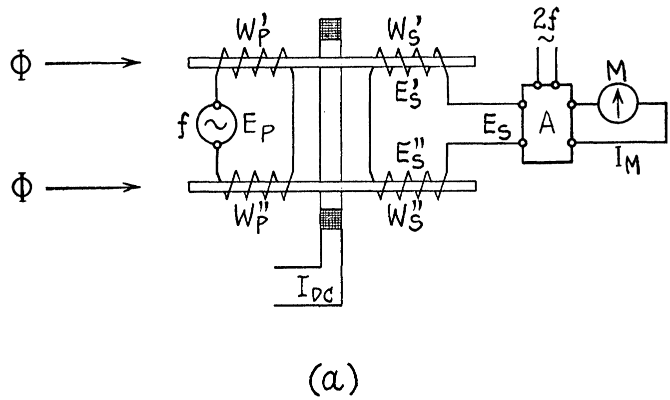
According to Coales's study, the zero of the different fundamental-frequency types of magnetic modulators is stable in the order of from 10^{-6} down to 10^{-15} watt, calculated at the input. The input stability level of the second-harmonic-type magnetic modulators is listed as from 10^{-10} down to 10^{-17} watt.

Further contributions concerning the performance of second-harmonic-type magnetic modulators and their applications have been made by H. S. Sack, R. T. Beyer, G. H. Miller, and J. W. Trischka,[7] F. C. Williams and S. W. Noble,[8] J. M. Manley,[9] and G. Wennerberg.[10]

16.4. Second-harmonic-type Flux-gate Magnetometers. By modifying the frequency-transformer and magnetic-modulator arrangements already described in this chapter, it is possible to obtain devices for measuring external magnetic fields. In 1928, H. Aschenbrenner and G. Goubeau[11] applied a second-harmonic-type flux-gate magnetometer for recording the changes in intensity of the earth's field. Such devices are excellent direction finders and can be used as a compass. Since delicate and sluggish moving systems are eliminated, the "Permalloy compass," which may be made self-orienting or may also be used for automatic steering (aerial navigation), is superior to the conventional magnetic compass.

Second-harmonic-type flux-gate magnetometers are particularly suitable for geophysical measurements and special military applications ("air-borne magnetometer"). There are also numerous other applications in testing of nonmagnetic materials (*e.g.*, detection of iron particles in aluminum or copper rods), telemeter arrangements ("Magnesyn" system),[12] automatic compensation of the earth's field in test rooms, etc.

Figure 16.5 shows two examples of second-harmonic-type flux-gate magnetometers. Two equally rated saturable-transformer elements with high-permeability (Permalloy) cores, in the form of a rod ("open" magnetic circuit), are used in both cases. It is to be noted that the longitudinal dimension of these cores is great in proportion to the transverse dimension. The primary windings W'_P, W''_P are supplied from an a-c generator (voltage E_P, frequency f); and the secondary windings W'_S, W''_S apply to the electronic amplifier A an alternating voltage $E_S = E'_S - E''_S$ having the frequency $2f$. The output current I_M, derived from a phase-sensitive rectifier circuit combined with A, is



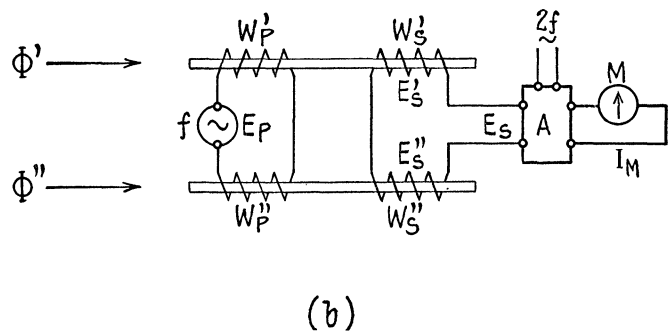


FIG. 16.5. Second-harmonic-type flux-gate magnetometer circuits: (a) basic circuit: $I_M = f(\Phi)$; (b) gradiometer circuit: $I_M = f(\Phi' - \Phi'')$.

measured by means of a moving-coil instrument M, which may be a recorder.

The operation of the arrangement with series-aiding-connected primary windings W'_P, W''_P (Fig. 16.5a) is similar to that of the frequency-transformer circuit of Fig. 16.1a. With no-signal conditions ($\Phi = 0$), the symmetrical and opposite secondary voltages E'_S, E''_S are identical in waveshape and balance each other in their effect upon the a-c amplifier A (Fig. 16.2). However, with a d-c flux Φ, the voltage waves E'_S, E''_S become asymmetrical, and a second-harmonic voltage $E_S = E'_S - E''_S$ appears on the input of the amplifier A. Since the phase of E_S reverses when the direction of the d-c flux $\pm\Phi$ is reversed, this device is able to measure not only the intensity of an external magnetic field but also its direction. Thus, the output current I_M is a function of the d-c flux $\pm\Phi$.

Figure 16.5*a* illustrates also the possibility of providing a compensating winding carrying an auxiliary direct current I_{dc}, which may be constant or variable. Application of this additional arrangement offers the following methods for measuring the actual intensity of the external magnetic field Φ:

1. A large amount of the d-c flux Φ is compensated by means of an opposing *constant* d-c flux, produced by a constant direct current I_{dc}, which may be derived from a battery or from a self-balancing potentiometer controlled by a standard cell. In this case, the flux-gate magnetometer measures the actual difference between the d-c flux Φ and the constant d-c flux of current I_{dc}. Obviously, this differential method is particularly suitable for measuring very small changes in intensity of the external magnetic field. Of course, the compensating winding may also be replaced by a permanent magnet having a fixed position with regard to the magnetic-core elements.
2. The d-c flux Φ to be measured is *completely* compensated by the opposing d-c flux of variable direct current I_{dc}. In this case, the procedure consists in adjusting the magnitude of I_{dc} so that the output current I_M is zero. With this "full-balance condition," the d-c flux produced by I_{dc} is exactly equal to the d-c flux Φ to be measured: $I_{dc} = \text{const} \times \Phi$. Thus, variations of Φ may be measured in this way after proper calibration of the device. Of course, such an arrangement may be automatically balanced by means of conventional self-balancing instruments, as used in the technique of self-balancing d-c potentiometers.
3. The compensating winding is supplied from the d-c output of the phase-sensitive rectifier circuit operating in connection with the electronic a-c amplifier A, so that $I_{dc} = I_M$. In this case, the d-c flux Φ to be measured is *nearly* (*e.g.*, in an amount of 99.5 per cent of Φ) balanced by the opposing d-c flux of current I_{dc}; and the output current $I_M = I_{dc}$, established by the very small flux difference (*e.g.*, 0.5 per cent of Φ) is proportional to the actual magnitude of Φ. Such an arrangement represents a self-balancing flux-gate-magnetometer circuit employing a very large amount of negative feedback, so that the accuracy of measurement will not be appreciably affected by changes in gain of the amplifier A and the associated phase-sensitive rectifier circuit.

The arrangement with series-opposing-connected primary windings W'_P, W''_P (Fig. 16.5*b*) operates in such a way that the second-harmonic voltage $E_S = E'_S - E''_S$ is proportional to the actual *difference* between the d-c flux values Φ' and Φ'': $I_M = \text{const} \times (\Phi' - \Phi'')$. Therefore, this differential-type flux-gate magnetometer, sometimes called "gradiometer," measures the actual difference between the flux Φ' in the first core element (W'_P, W'_S) and the flux Φ'' in the second core element (W''_P, W''_S). When the external field is homogeneous ($\Phi' = \Phi''$), then the

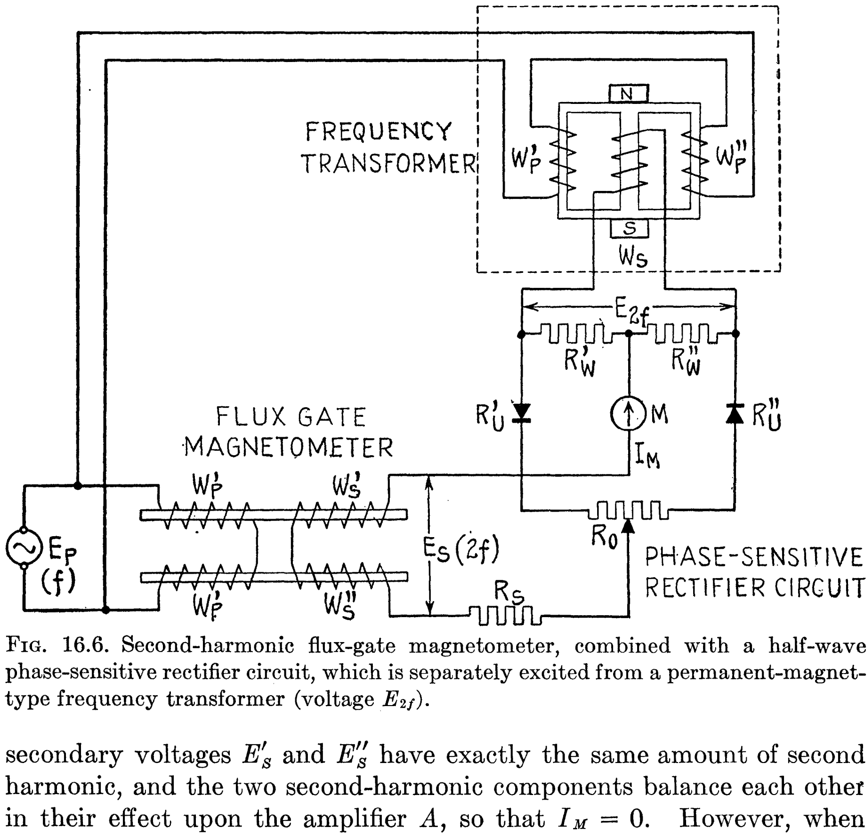


FIG. 16.6. Second-harmonic flux-gate magnetometer, combined with a half-wave phase-sensitive rectifier circuit, which is separately excited from a permanent-magnet-type frequency transformer (voltage E_{2f}).

secondary voltages E'_S and E''_S have exactly the same amount of second harmonic, and the two second-harmonic components balance each other in their effect upon the amplifier A, so that $I_M = 0$. However, when there is a certain amount of inhomogeneity ($\Phi' \gtrless \Phi''$), then a second-harmonic voltage E_S will appear at the input of A, and also an output current I_M which is a function of $\Phi' - \Phi''$.

Figure 16.6 shows a second-harmonic flux-gate magnetometer, combined with a half-wave phase-sensitive rectifier circuit, which is separately excited from a permanent-magnet-type frequency transformer. The author has successfully used this simple arrangement for demonstrations, to illustrate the operation of second-harmonic-type frequency-transformer and flux-gate-magnetometer circuits and to show the properties of phase-sensitive rectifiers. This arrangement contains:

1. The flux-gate magnetometer itself, having two Permalloy cores, in the form of rods, with primary windings W'_P, W''_P and secondary windings W'_S, W''_S
2. The phase-sensitive rectifier circuit, comprising two half-wave-rectifier units R'_U, R''_U; three fixed resistors R'_W, R''_W, R_S; a potentiometer resistor R_0 (zero adjustment); and a microammeter ($+100$ to 0 to -100 μa)
3. The frequency transformer has a three-legged Permalloy core with series-connected primary windings W'_P, W''_P, a common secondary winding W_S (secondary voltage E_{2f}), and a permanent horseshoe magnet providing the necessary d-c flux

This arrangement has been used for demonstrating the fundamental principles of the "Permalloy compass" (I_M is zero, when the two cores are in west-east direction) and the flux-gate magnetometer (I_M is a function of an external magnetic field, produced by permanent magnets of different sizes, which are placed at variable distances from the two core elements).

Contributions concerning the performance of second-harmonic-type flux-gate magnetometers and their applications have been made by E. P. Felch, W. J. Means, T. Slonczewski, L. G. Parratt, L. H. Rumbaugh, and A. J. Tickner,[13] F. Förster,[14] and R. Feldtkeller,[14] and others.[15]

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CHAPTER 17

Technical Properties of Magnetic Amplifiers

17.1. General Comments on the Technical Properties of Magnetic Amplifiers. The adaptability of magnetic-amplifier circuits to various problems in the techniques of measurement and control may be judged by considering their typical properties, which are summarized as follows:

1. Magnetic amplifiers using no vacuum tubes avoid the serious disadvantages caused by the facts that the life of vacuum tubes is limited and that shocks or vibrations are dangerous to such tubes. Therefore, magnetic amplifiers are sturdier and more reliable than electronic amplifiers.
2. There are no filaments to burn out, and since there is no warm-up time, a magnetic amplifier is instantly operable on switching on.
3. Magnetic amplifiers, having no moving parts and not requiring servicing and replacement, are of rugged construction and can be hermetically sealed.
4. The fact that complete isolation is possible between input and output of magnetic-amplifier circuits represents an important advantage for many applications, particularly in the field of instrumentation problems.
5. Magnetic amplifiers offer the possibility of mixing several (d-c or a-c) control currents at different voltage levels; and they can be used to convert d-c quantities to corresponding a-c or d-c quantities at higher level.
6. The core size of saturable-reactor elements, as used in magnetic-amplifier circuits, is inversely proportional to the power-supply frequency (f_P = 50 to 2,000 cps).
7. It is possible to obtain power gains of 10,000 to 100,000 in one stage. Higher power gains can be obtained by using multistage arrangements.
8. The transient performance of a magnetic amplifier depends upon the actual values of over-all power gain and ampere-turn gain. Generally, the response time is about 2 to 100 cycles of the power-supply frequency (f_P). In special cases, *e.g.*, with properly rated

half-wave self-saturating circuits, inherent one-cycle response can be obtained.

9. The control circuit of magnetic amplifiers can be matched to input impedance values in the range from 0.1 up to about 100,000 ohms, and the load circuit can be matched to output impedance values of about 5 to 5,000 ohms at a power-supply frequency of 50 or 60 cps. With higher power-supply frequencies, application of excessively high number of turns of the different windings must be avoided with regard to the distributed capacitance of such arrangements, which may cause disturbing effects.
10. Magnetic amplifiers of the balance-detector type having duo-directional transfer characteristics and a very small drift rate can be made by providing symmetrical push-pull circuits. Their zero stability is equivalent to an input power of about 10^{-8} watt. This value may be reduced to about 10^{-11} watt, if differential-feedback arrangements and special drift-compensating means are used.[1]
11. Low-impedance current-operated magnetic amplifiers and high-impedance voltage-operated electronic amplifiers are complementary devices and, therefore, should not be considered as rivals. They may be combined in many cases to give the best solution for a particular problem in the techniques of measurement and control.
12. It is to be noted that successful research work has been extended toward the development of audio-frequency magnetic amplifiers operating from an a-c power supply with a carrier frequency of about 10,000 to 20,000 cps (F. G. Logan,[2] 1948, and E. H. Frost-Smith,[3] 1949).

17.2. Transient Response of Magnetic Amplifiers. The use of magnetic amplification is associated with a certain time lag between the application of the input signal and the attainment of the full output load current. This sluggishness of response is not an important drawback in some applications, where the additional lags arising from external parts of the electrical control system (due, for example, to the thermal inertia of a thermocouple) are appreciably greater than those inherent in the magnetic amplifier itself. But in other applications, especially those in the field of high-performance servomechanisms and some other types of automatic control and instrument work, a very short response time is of paramount importance.

Reference to the literature[4] shows that considerable effort has been expended in theoretical and experimental investigations on the transient performance of various types of magnetic-amplifier circuits. The object of these investigations is to determine the major factors which control the transient response of saturable-reactor devices with resistive or inductive loads. It has been shown that application of rectangular-hysteresis-loop core materials,[5,6] in combination with a suitable core construction, results in a very close approximation between the usual theoretical assumptions and actual experimental conditions. Therefore, the time constant of such magnetic amplifiers can be described by simple linear equations with sufficient accuracy to be used in the design of these amplifiers.

a. Transient Response of the Series-connected Saturable Reactor Operating under Nonfeedback Conditions. H. F. Storm's analysis[7] deals with saturable reactors whose core material can be approximated by a rectangular magnetization curve (Fig. 2.1*b*). The transient response of the basic nonfeedback circuit with resistive load is analyzed with respect to changes of d-c control voltage, power-supply voltage, and load resistance. When operating in the proportional region of the transfer characteristic, the response of the load current to any of the previous changes is an exponential function, governed by a single time constant, which varies with the load and inversely with control-circuit resistance and power-supply frequency.

The response of the load current I_L to sudden changes of control voltage E_C can be described by a time constant T, which depends upon the actual turns ratio N_C/N_L, the actual resistance ratio R_L/R_C, and the power-supply frequency f_P:

$$T = \frac{1}{4f_P}\frac{N_C^2}{N_L^2}\frac{R_L}{R_C} \qquad \text{seconds} \tag{17.1}$$

where N_L is the number of turns of the a-c load windings, N_C is the number of turns of the d-c control winding, R_L is the total resistance of the load circuit, and R_C is the total resistance of the control-circuit loop.

The over-all power gain K_P is defined as the ratio of the output power $P_L = I_L^2R_L$ divided by the input power $P_C = I_C^2R_C$:

$$K_P = \frac{P_L}{P_C} = \frac{I_L^2}{I_C^2}\frac{R_L}{R_C} = \frac{N_C^2}{N_L^2}\frac{R_L}{R_C}k_f^2 \tag{17.2}$$

where $k_f = I_{L(\text{rms})}/I_{L(\text{avg})}$ is the form factor of the load current. Considering Eqs. (17.1) and (17.2), the time constant is

$$\begin{aligned} T &= \frac{1}{4f_P}\frac{K_P}{k_f^2} \qquad \text{seconds} \\ &= \frac{1}{4}\frac{K_P}{k_f^2} \qquad \text{cycles} \end{aligned} \tag{17.3}$$

Thus, the ratio of over-all power gain over time constant becomes

$$\frac{K_P}{T} = 4f_Pk_f^2 = 4k_f^2 \qquad \text{per cycle} \tag{17.4}$$

With nonfeedback conditions, the time constant T, as defined by Eq. (17.3), is the same for increasing or decreasing control voltage E_C. The geometrical dimensions of the magnetic core do not enter into this expression. However, in practice, the influence of the nonideal magnetization curve will cause certain differences between measured and calculated values of time constant T. Generally, for preliminary design work, Eq. (17.3) gives a useful estimate of the response time; it also applies when Mumetal cores are used (C. S. Hudson,[8] A. G. Milnes,[9] H. M. Gale and P. D. Atkinson[10]).

b. Transient Response of Full-wave Feedback-type Magnetic-amplifier Circuits with Series-connected D-C Control Windings. S. E. Hedström and L. F. Borg[11] have shown that the time constant of such circuits can be expressed in terms of power-supply frequency f_P and actual over-all volts-per-turn gain K_P/A:

$$\begin{aligned} T &= \frac{1}{4f_P}\frac{K_P}{A} \qquad \text{seconds} \\ &= \frac{1}{4}\frac{K_P}{A} \qquad \text{cycles} \end{aligned} \tag{17.5}$$

where $A = AT_L/AT_C$ is the actual ampere-turn gain, as defined by Eq. (6.11), and K_P is the actual over-all power gain, as defined by Eq. (6.2). Thus, the ratio of over-all power gain to time constant becomes

$$\frac{K_P}{T} = 4f_Pk_f^2A = 4k_f^2A \qquad \text{per cycle} \tag{17.6}$$

This expression shows that the ratio K_P/T increases linearly with the actual ampere-turn gain A.

Storm[12] stressed that, with positive feedback, the time constants for increasing and decreasing control voltages (E_C) are no longer identical and that, for time constants of less than one cycle, Eq. (17.5) becomes inaccurate and indicates values of T that are too small.

c. Transient Response of A-C Controlled Full-wave Magnetic-amplifier Circuits. F. G. Logan[13] pointed out that, in a-c-controlled self-saturating circuits, "control is achieved in full during each cycle, and when the controlling current is changed, full response to this change is reached not longer than one cycle later. As the controlling potential is alternating, its current is capable of substantially instantaneous change." However, "in order that the control winding may have current passed through it to produce this bucking effect with reference to the flux due to the load windings, the voltage applied to the control winding must be sufficient to overcome the induced voltage in the control winding due to the flux of the load windings." When an alternating voltage is suddenly applied to the control winding by means of a switch, "the change accomplished by the control is very rapid, not requiring more than one cycle with a resistive load after the switch contacts close."[13]

d. Transient Response of Special Types of Magnetic-amplifier Circuits. Various half-wave and full-wave types of d-c-controlled self-saturating circuits having inherent one-cycle response have been devised by R. A. Ramey.[14,15] These circuits employ a combination of an a-c bias transformer acting as a "magnetizing element" with half-wave rectifiers in the d-c control circuit (Figs. 10.6 and 11.10).

A. E. Schmid[16] has shown that a response time of one cycle can be obtained in half-wave push-pull circuits, which may be d-c or a-c controlled. Several two-stage half-wave magnetic servo amplifiers with inherent one-cycle response, devised in the U.S. Naval Ordnance Laboratory,[17] have proved that excellent results can be obtained with such arrangements, especially when using those improved types of two-phase induction motors which have inherent damping properties.

17.3. Experimental Methods for Analyzing the Transient Response of Magnetic Amplifiers. In considering the accuracy of theoretical methods for determining the response time of conventional types of magnetic amplifiers having a response time of about 6 to 30 cycles, it must be remembered that in most practical design problems an approximate value is all that is required. However, in regard to the transient performance of special magnetic-amplifier circuits having a very short response time of only about one cycle,[18] experimental methods have assumed increasing importance in their development.

Obviously, the cathode-ray oscilloscope, representing the most versatile instrument with which to observe the waveforms of electric and magnetic quantities varying with time, is extremely suitable for such purposes. The circuit diagram of a simple transient-response analyzer, devised by the author,[19] is shown in Fig. 17.1. According to conventional technique, the voltage drop $I_L R$ produced by the output (load) current I_L across an ohmic resistor R is applied to the one channel (V_1) of a dual-beam oscilloscope, while the input (control) voltage E_C is applied to the second channel (V_2). The magnetic amplifier to be investigated and a small step-down transformer T are supplied from a phase shifter, whereas a synchronous motor SM operating a special switch device with contacts A and B is directly supplied from the three-phase power supply. This switch device is synchronously operated in such a way that it makes during a period of 6 cycles and breaks during the succeeding period of 6 cycles, alternately. Therefore, with a 60-cycle power supply, the shaft of the switch device moving the cam, which operates the contacts A and B, rotates with a synchronous speed of 300 rpm corresponding to 12 cycles per revolution.

The control voltage $E_C = I_M R_N$, supplied from a battery (d-c control) or from the step-down transformer T (a-c) control, can be adjusted by

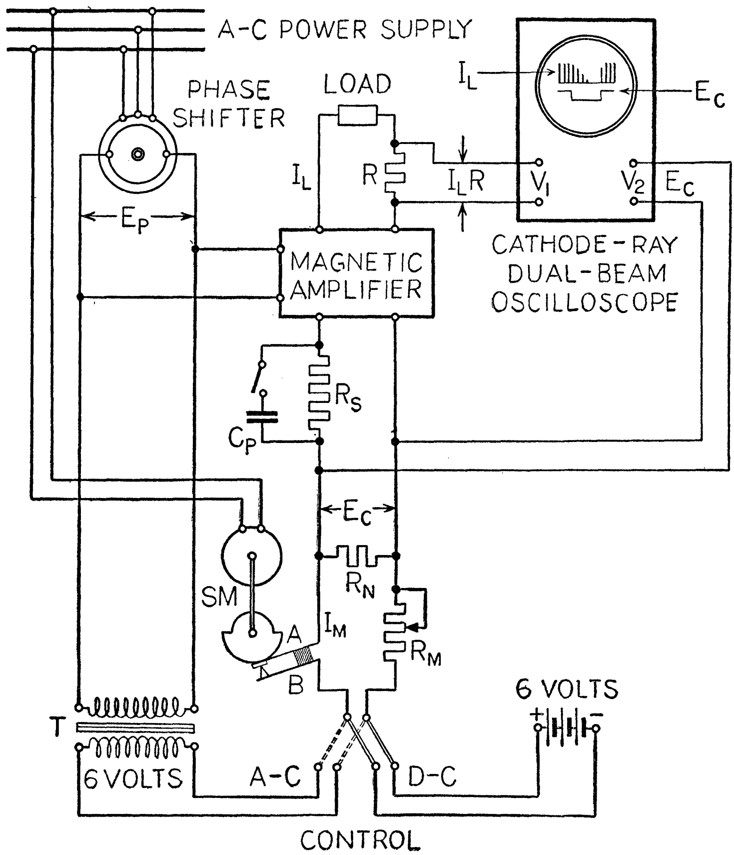
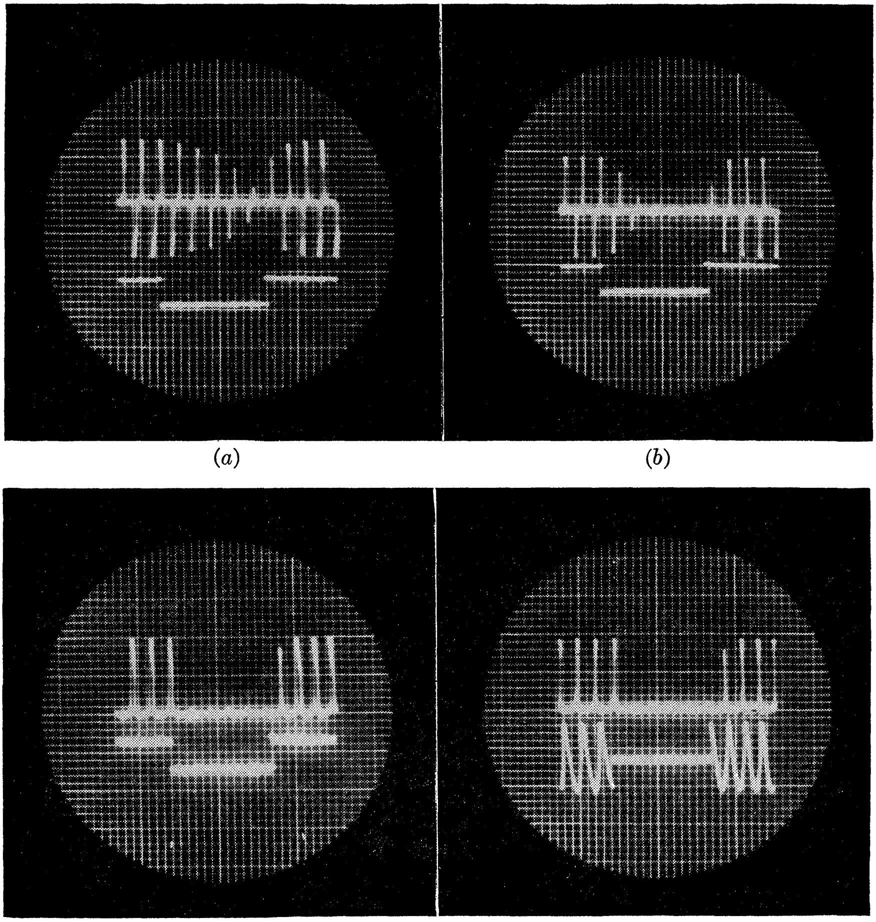


FIG. 17.1. Circuit diagram of the transient-response analyzer, which exhibits the complete transient phenomena of the magnetic amplifier permanently as a "standing" picture on the long-persistence screen of a cathode-ray dual-beam oscilloscope. (*W. A. Geyger.*[19])

means of a variable series resistor R_M, according to the actual requirements. Figure 17.1 also illustrates the possibility of varying the operating conditions of the control-circuit loop by means of a "forcing resistor" R_S and a shunt condenser C_P. In this way, the effect of "lead networks" on the transient performance of magnetic amplifiers may be investigated.

When the sweeps of the dual-beam oscilloscope are synchronized so that 12 cycles of the (a-c or d-c) output current I_L appear on the screen,



(a) (b)

(c) (d)

FIG. 17.2. Oscillograms presenting the transient performance of various push-pull self-saturating magnetic-amplifier circuits operating under the following conditions: (a) Conventional type of circuit with four series-connected d-c control winding units and an a-c output with resistive load. Series resistance in the control circuit $R_S = 500$ ohms; $C_P = 0$. (b) Conditions like (a), but $R_S = 2{,}000$ ohms, $C_P = 0$. (c) Half-wave circuit having inherent one-cycle response, with d-c control. (d) Conditions like (c), but with a-c control.

then the operation of the switch device will also be synchronized; and a sequence of 6-cycle making and 6-cycle breaking will exhibit the complete transient phenomena of the magnetic amplifier permanently as a "standing" picture on the long-persistence screen.

Introduction of the phase shifter offers the possibility of varying the actual time interval between the instant of making (or breaking) of the contacts A and B and the instant when the power-supply voltage E_P

of the magnetic amplifier goes through zero. Of course, the phase shifter may be eliminated if the angular position of the cam on its shaft can be continuously changed by mechanical means, as in conventional Joubert-disk-type contact arrangements.

When operating with higher power-supply frequencies, the synchronously operated switch device representing a mechanical modulator may be replaced by any suitable type of electrical modulator. Electronic devices producing a synchronous auxiliary current with rectangular waveshape are very suitable for this purpose.

Figure 17.2, illustrating the practical application of the transient-response analyzer shown in Fig. 17.1, presents the transient performance of various push-pull self-saturating magnetic-amplifier circuits operating under the indicated conditions. It is to be noted that the time of exposure was about 2 sec in all cases. Thus, the photographic reproduction of the illuminated calibrated scale, which greatly facilitates such studies, is very good.

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CHAPTER 18

Typical Applications of Magnetic Amplifiers

18.1. Introduction. Reference to the literature shows that magnetic-amplifier circuits are already applied in various fields of electrical engineering, to replace vacuum-tube amplifiers and also to obtain special functions in the techniques of measurement and control.

This chapter illustrates some typical applications of magnetic amplifiers and describes also several kinds of auxiliary circuits for magnetic-amplifier applications: magnetic voltage stabilizers, static frequency transformers, and special control circuits for magnetic amplifiers.

18.2. Application of Magnetic Servo Amplifiers. In principle, the use of magnetic amplification in servo systems is associated with a certain additional time lag of the magnetic amplifier itself. However, it should be borne in mind that the lags arising from the external parts of the servomechanism are often appreciably greater than those inherent in the magnetic amplifier. For example, when working with a power-supply frequency of 400 cps, the usual response time of about 6 to 20 cycles (15 to 50 msec) represents a transient performance which will not cause any difficulty in many cases. Of course, if the magnetic servo amplifier is designed to operate from a power supply of 60 cps, a response time of 6 to 20 cycles (100 to 333 msec) may not be acceptable. Thus, in this case, a very short response time of only about one cycle is really desirable. It has already been stressed in Sec. 17.2*c* and *d* that this short response time may be achieved in different ways by using special push-pull circuits with inherent one-cycle response.

Referring to instability in the operation of servomechanisms, which gives rise to "oscillating" or "hunting," some antihunt method must be employed for canceling the tendency to oscillate. Common remedies for hunting are damping the motor itself by introducing dynamic braking, application of a resistance-capacity differentiating network or tachometer to stabilize by means of "derivative feedback," and magnetic or viscous "flywheel" dampers.[1] The purpose of such devices is to provide damping at a rate appropriate to the requirements of stability.

Another important factor governing the satisfactory operation of a closed-loop servomechanism is the proper adjustment of the "sensitivity" or "gain" of the magnetic amplifier. The combination of the error-signal voltage per degree and the actual amplifier gain governs the "tightness" of the system. If the gain is too great, the servomechanism will oscillate in spite of application of an effective antihunt method. It is often desirable to adjust the gain of the magnetic servo amplifier in such a way that not more than one overshoot, or no overshoot at all, will be obtained.

Referring to the damping properties of two-phase induction motors, which are extensively used in high-performance servomechanisms, it is important to stress that some of these motors have good ratios of viscous friction to inertia or special damping means, and others have not. Separately excited induction-meter-type reversible motors, as sometimes used in self-balancing potentiometer recorders,[2] have a permanent magnet producing eddy-current damping. Some types of low-inertia squirrel-cage induction motors and drag-cup-type motors have eddy-current damping devices built into the motor. There are, however, other constructions of such motors which have virtually no damping properties, so that additional damping methods must be applied to prevent hunting.

The use of vacuum-tube amplification offers the possibility of introducing effective dynamic braking of the motor by means of the d-c component flowing through the plate circuit, which contains the amplifier field winding of the two-phase induction motor. However, if such a motor is to be controlled by means of a magnetic servo amplifier, considerable practical difficulty may be caused by the fact that the alternating load current of the conventional magnetic amplifiers of the balance-detector type does not have a d-c component which may be used for producing dynamic braking of the motor.

Referring to the half-wave self-saturating push-pull type of magnetic-amplifier a-c motor-control circuits, as described in Chap. 13, it is to be noted that the torque-producing a-c component and the associated d-c component of the unidirectional reversible-motor-control current will always increase or decrease at the same rate. Thus, there is no damping effect at all with no-signal conditions. This is generally not desirable in operations with those constructions of two-phase induction motors which have very poor damping properties, because there is no dynamic-braking effect in the vicinity of the null point. However, these half-wave push-pull-type circuits have proved to be very useful with a reversible a-c motor having inherent damping properties or special damping means.[3]

With regard to the application of derivative feedback in the field of magnetic servo amplifiers, it is important to note that some difficulty may be caused by the fact that the output (load) current of magnetic amplifiers—sometimes even with no-signal conditions—contains a very large a-c component superimposed on the d-c component which is used for supplying the derivative-feedback circuit.

The special magnetic-servo-amplifier circuits with inherent dynamic-braking characteristics (Secs. 8.4, 8.5, and 12.4) have proved to be particularly useful for the control of those types of two-phase induction motors having such a low ratio of effective viscous friction to inertia

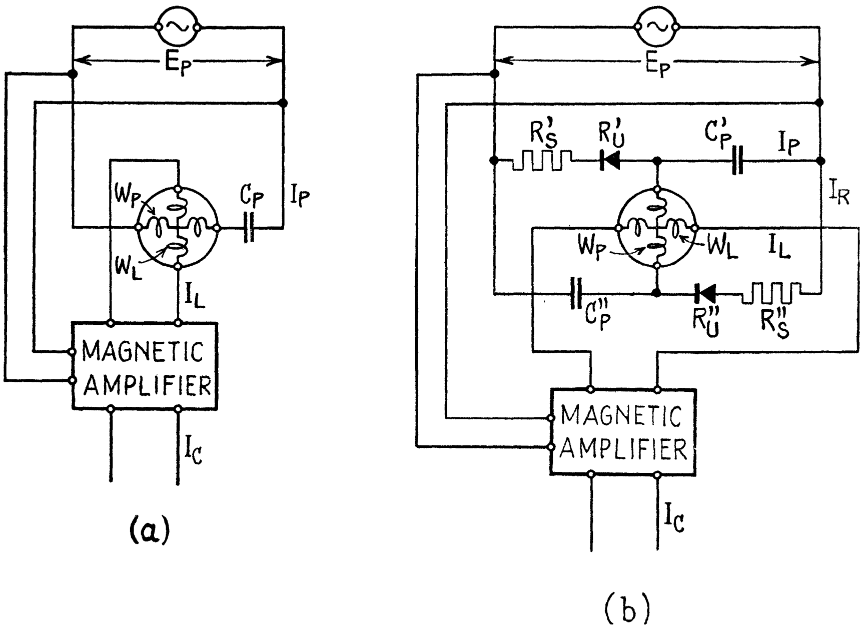

FIG. 18.1. Magnetic-servo-amplifier circuits with two-phase induction-type reversible motor W_PW_L: (*a*) conventional circuit with a single series condenser C_P; (*b*) bridge-type circuit producing constant dynamic braking of the two-phase motor.

that it is difficult to stabilize at high gain in a closed loop. The basic idea consists in feeding the amplifier field winding of the a-c reversible motor with a variable unidirectional current having a fundamental-frequency (60- or 400-cycle) component and a d-c component producing an effective dynamic braking of the motor, preferably in such a way that the d-c component increases when the a-c component decreases, and vice versa.

Experience has shown that, when using magnetic-servo-amplifier circuits without inherent dynamic-braking characteristics in connection with those types of a-c motors having good ratios of viscous friction to inertia, additional damping means may be necessary in some cases.

Application of a constant d-c component in the line field winding of the two-phase motor has proved to give satisfactory results, provided that the auxiliary d-c damping circuit is properly designed.

Figure 18.1 shows the conventional arrangement and a bridge-type circuit which produces constant dynamic braking of the two-phase motor. The bridge circuit contains two condensers C'_P, C''_P, two unidirectional resistors R'_U, R''_U with series resistors R'_S, R''_S, and the line field winding W_P of the motor, whereas in the conventional circuit this winding W_P is connected with a simple series condenser C_P to the power-supply

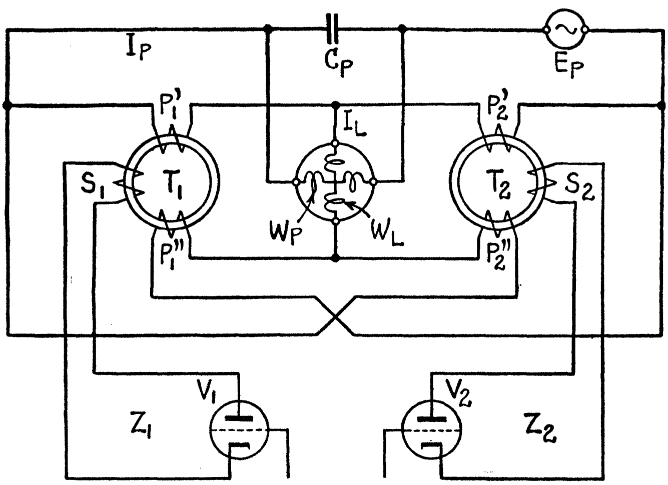

FIG. 18.2. Motor-control network with impedance-controlled saturable transformers T_1, T_2 and triodes V_1, V_2 (RCA type 8005).

voltage E_P. In both cases, the amplifier field winding W_L is supplied from the output (I_L) of the magnetic servo amplifier, which is controlled by the reversible d-c or a-c control current I_C.

18.3. Application of Electronic Components in Magnetic-amplifier Circuits. It has already been mentioned that a properly designed combination of magnetic- and electronic-amplifier components may give the best solution of a particular problem in many cases, the magnetic-amplifier circuit operating either as an input stage (Figs. 5.13, 5.14, and 16.4) or as an output stage (Figs. 5.11 and 5.12).

Figure 18.2 illustrates another example of a motor-control network[4] with finite impedances, which is particularly suitable for controlling large-sized motors. Under the operating conditions of this circuit, the torque-speed characteristic curves differ over the useful range from those obtained when the two-phase induction motor is energized from a supply source with zero impedance.

The fundamental principle of this arrangement consists in controlling the a-c motor by varying the actual impedances of the primary windings $P_1'P_1''$, $P_2'P_2''$ of T_1 and T_2. One winding (W_L) of the motor is fed from the output of the symmetrical bridge circuit, which may be unbalanced in either direction, depending on the control voltages applied to the triodes V_1 and V_2 (*e.g.*, RCA type 8005). The other winding (W_P) of the motor, with shunt condenser C_P, is connected in series with the bridge circuit and the a-c power supply (E_P). The phasing condenser C_P must be properly rated so that the current in W_P is approximately 90° out of phase with the current in W_L, as required for the operation of a two-phase induction-type reversible motor.

It is to be noted that the impedance-controlled transformers T_1, T_2 are wound with double primary windings $P_1'P_1''$, $P_2'P_2''$ and have a high step-up ratio (*e.g.*, transformer turns ratio about 30:1) to the secondary windings S_1, S_2. Thus, the actual impedance of the low-voltage, high-current motor-control windings W_L is matched to the plate-circuit impedances Z_1, Z_2 of the high-voltage, low-current tubes V_1, V_2, which are connected across S_1 and S_2. When V_1 and V_2 are not conducting, they reflect a high impedance into the primary windings of the two transformers T_1 and T_2. However, when V_1 and V_2 are conducting, the effective impedances of the primary windings are correspondingly reduced according to the actual magnitudes of the two plate currents, which are composed of a succession of half sinusoids. In practice, the effective impedances of the primary windings $P_1'P_1''$ and $P_2'P_2''$ may be varied independently by applying variable (d-c or a-c) control voltages to the grid circuits of the triodes V_1 and V_2, preferably in a push-pull relationship.

Figure 18.2 shows that the alternating voltages which are induced from the primary windings into the secondary windings S_1, S_2 are applied to the plate circuits of V_1 and V_2. Therefore, plate current will flow only during those half cycles when the plates of V_1 and V_2 are positive, the actual magnitudes of the plate currents being a function of the instantaneous grid voltages of V_1 and V_2. In this way, the motor-control network produces motor torque, controlled in amount and direction by means of two push-pull-operated triode elements.

18.4. Auxiliary Circuits for Magnetic-amplifier Applications. Generally, application of magnetic amplifiers is associated with application of those auxiliary circuits which, while necessary in a practical design, do not represent essential parts of the magnetic-amplifier circuit itself. In practice, there are constant-voltage devices, frequency transformers, voltage-, frequency-, or temperature-sensitive bridge or potentiometer circuits, etc., which are needed as well as the switches, relays, protection devices, and other conventional arrangements of commercial equipment.

In the preceding chapters tracing the fundamental magnetic-amplifier circuits and describing their essential features, it was desirable to eliminate these subordinate circuits. However, in the following sections of this chapter, a few typical examples of such auxiliary devices will be briefly described, with reference to the literature.

a. Magnetic Voltage Stabilizers. Since constant magnitude of supply voltage is vital to satisfactory performance of various components in electrically operated equipment, numerous methods have been developed to compensate for the line-voltage fluctuations.[5] The magnetic voltage stabilizer, a cousin of the magnetic amplifier, using a combination of nonlinear and linear magnetic circuits, consists of special transformerlike structures and condensers. It has no moving parts or tubes and can perform the added function of a step-up or step-down transformer, eliminating the use of additional transformer components.

Obviously, the performance of some types of magnetic-amplifier circuits may be improved by providing a magnetic voltage stabilizer acting as power-supply transformer of the arrangement. There are various kinds of such devices, according to the different fields of practical applications.

Figures 5.9 and 7.7 show a simple magnetic voltage stabilizer consisting of a three-legged transformer T_E with a saturated and an unsaturated (air-gap) magnetic circuit and a properly rated shunt condenser C_S. The primary winding N_P on the center leg is directly connected to the power-supply voltage E_P. The series-opposing-connected secondary windings N'_S and N''_S on the outer legs supply the magnetic-amplifier circuit, representing an inductive a-c load of the constant-voltage transformer.

Referring to the drift rate of push-pull circuits, it is to be noted that, with perfectly symmetrical circuit components, there will be no drift at all. In practice, however, a certain amount of drift will be caused by the fact that the two parts of the push-pull circuit are not perfectly matched with regard to their voltage-current characteristics. Experience has shown[6] that the "voltage drift" of very sensitive push-pull input-stage circuits, preferably utilizing differential feedback, can be considerably reduced by application of a small constant-voltage transformer, the actual variations of the secondary (load) voltage of this transformer being about one-twentieth of line-voltage fluctuations of the order of ± 10 per cent.

b. Static Frequency Transformers. Since the minimum time delay in the operation of a magnetic amplifier will be of the order of one cycle of the power-supply frequency f_P, an obvious method for reducing the response time is increasing the supply frequency. Furthermore, with higher values of supply frequency, the saturable-reactor components of the magnetic amplifier are smaller than those used on a 60-cycle power supply.

In 1939, A. U. Lamm[7] suggested the use of a static frequency transformer in connection with magnetic amplifiers to increase the speed of response, particularly that of the input circuit of multistage arrangements. Static frequency transformers, sometimes called "frequency multipliers" or "harmonic generators," contain saturable reactors, linear reactors, filter circuits, and, in some cases, also rectifiers. The basic principle of such arrangements consists in producing higher harmonics of the power-supply frequency by saturation of magnetic cores and eliminating the fundamental-frequency component by means of bridge and/or filter circuits. In this way, a-c power of 60 cps can be converted into power of higher frequency, *e.g.*, 120, 180, 300, or 540 cps.

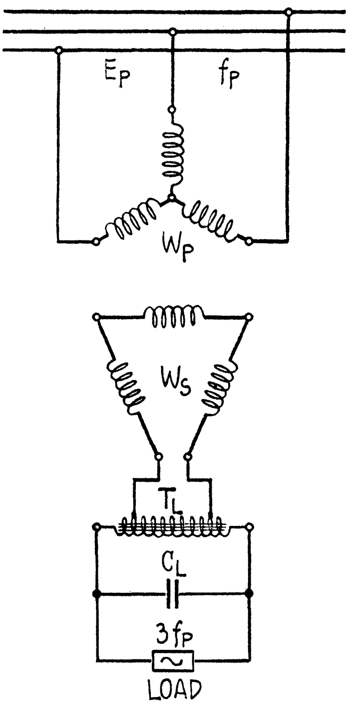

FIG. 18.3. Method of producing third-harmonic-frequency power by means of three transformers with Y-connected primary windings W_P and open-Δ-connected secondary windings W_S.

Referring to the different types of static frequency transformers, it is to be noted that *even*-harmonic frequencies are produced by superposition of a d-c magnetization over the a-c magnetization produced by the power-supply generator. The simplest device is the "frequency doubler" (Sec. 16.2). Cascading of this arrangement makes it possible to produce the frequencies $4f_P$ and $8f_P$ with moderate efficiency. On the other hand, *odd*-harmonic frequencies are produced by superposition of two voltages or currents having a highly distorted waveshape (saturated reactor) and a substantially sinusoidal waveshape (linear reactor), respectively. In this way, higher frequencies in the range from $3f_P$ up to $9f_P$ can be produced with sufficiently high efficiency, using either one or two frequency-multiplier stages. Higher odd-harmonic frequencies (up to about $47f_P$) can be produced at higher-frequency levels with special arrangements.

Figure 18.3 illustrates the conventional method of producing third-harmonic-frequency power by means of three transformers with Y-connected primary windings W_P and open-Δ-connected secondary windings

W_S. An autotransformer T_L with shunt condenser C_L may be provided in the load circuit to increase the efficiency of the circuit.

In 1927, the author[8] devised various types of frequency transformers for supplying frequency meters and a-c bridge or potentiometer networks. Figure 18.4 shows a single-stage arrangement (frequency ratio 3:1 up to 11:1) containing a saturable reactor L_S, three properly tuned circuits $R_PL_1C_1$, L_2C_2, L_3C_3, and a transformer T. This arrangement is a modification of the basic frequency-transformer circuit, which was devised in 1923 by K. Schmidt[9] for operating radiotelephony transmitters from a-c machine generators (frequency ratio up to 47:1). In 1948, the author designed a small frequency transformer (output power about 1 watt) using rectangular-hysteresis-loop core material and the circuit of Fig. 18.4 for converting input power of 400 cps into

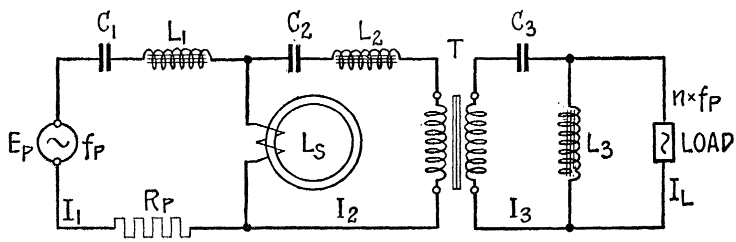

FIG. 18.4. Frequency transformer with a single saturable-reactor element L_S, three oscillating circuits $R_PL_1C_1$, L_2C_2, L_3C_3, and a transformer T.

output power of 2,000 cps (efficiency about 15 per cent). The amplitude of the 400-cycle component in the output circuit is about 5 per cent of the amplitude of the 2,000-cycle component. This device has been successfully used in several cases.

Figure 18.5 shows a two-stage type of static frequency transformer, each stage consisting of a bridge network with a saturated reactor (L_1', L_1'') and a linear (air-gap) reactor (L_2', L_2''), and properly tuned filter circuits ($C'L'$, $C''L''$). This arrangement makes it possible, for example, to convert an input power of 60 cps into an output power of $3 \times 3 \times 60 = 540$ cps, with satisfactory efficiency.

c. Special Control Circuits for Magnetic Amplifiers. Application of magnetic amplifiers for automatic control of electrical and nonelectrical quantities is associated with various problems concerning special arrangements for introducing the controlled quantity into the control circuit of the amplifier. The fundamental principle of automatic regulators consists in measuring the actual deviation of the regulated value from the rated value and converting the result of the measurement into another value which affects the power relations in the regulation system in such a way that the deviation is reduced substantially to zero. The operation of such regulation systems corresponds to that of an amplifier with negative feedback, and similar theoretical considerations can be applied in both cases.[10]

The basic method for comparing the regulated value with an electrical standard consists in providing a potentiometer or bridge network, the

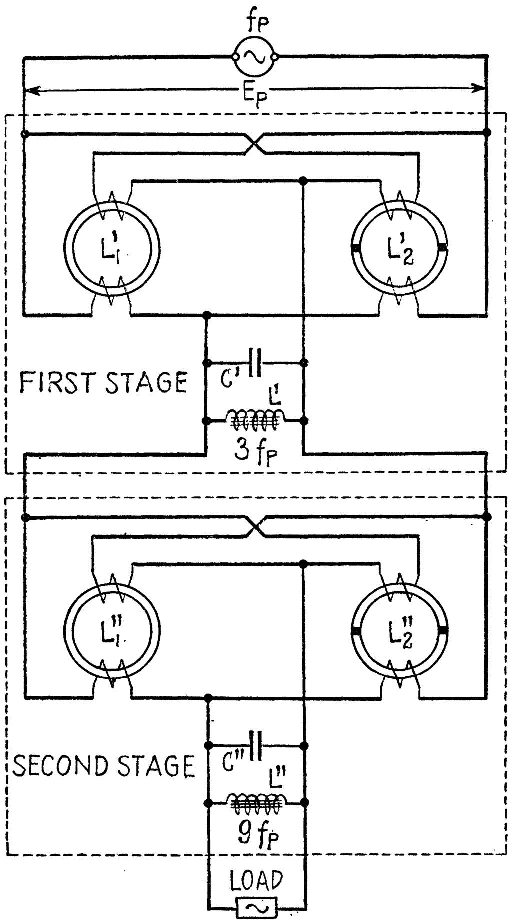

FIG. 18.5. Two-stage frequency transformer, each stage consisting of a bridge network and a filter circuit.

output of which is a function of the actual deviation of the regulated value from the rated value. There are numerous possibilities for designing such arrangements which represent voltage-, frequency-, or temperature-sensitive differential-type circuits.

Figure 18.6 shows two typical examples of a-c bridge networks operating in connection with an electronic amplifier, the output of which is used for supplying the "load," the control circuit of a magnetic amplifier. The voltage-sensitive circuit (Fig. 18.6*a*) containing two linear resistances R', R'' and two nonlinear resistances R'_E, R''_E is balanced at the rated value of the regulated voltage E_P. When the actual value of E_P deviates from the rated value, then the bridge network will be out of balance, and a corresponding current will flow through the load, with a direction depending on whether E_P has actually increased or decreased.

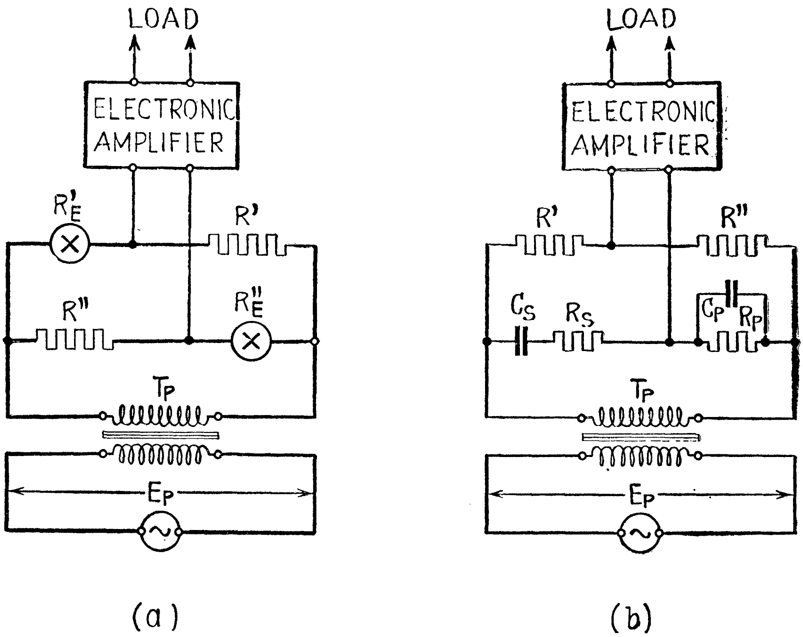

FIG. 18.6. A-c bridge networks operating in connection with an electronic amplifier, the output of which is used for supplying the "load," represented by the control circuit of a magnetic amplifier. (*a*) Voltage-sensitive circuit containing two linear resistances R', R'' and two nonlinear resistances R'_E, R''_E; (*b*) frequency-sensitive circuit containing four resistors R', R'', R_S, R_P and two condensers C_S, C_P.

Figure 18.6*b* illustrates the application of a frequency-sensitive Wien-Robinson type of bridge network[11] for similar purposes. This circuit, containing four resistors R', R'', R_S, R_P and two condensers C_S, C_P (preferably with $R_P = 2R_S$ and $C_S = 2C_P$), is balanced at the rated value of the regulated frequency f_P of power-supply voltage E_P:

$$(2\pi f_P)^2 = \frac{1}{R_S R_P C_S C_P}$$

A deviation of the actual value of f_P from the rated value will bring the bridge out of balance, and a corresponding current will flow through the load, with a direction depending on whether f_P has actually increased or decreased.

Figure 18.7 shows two typical examples of a-c differential-type networks supplying the twin control windings of a magnetic servo amplifier, without electronic-amplifier components. The voltage-sensitive circuit (Fig. 18.7*a*) compares the actual value of the nonlinear resistance R_E with the value of standard resistor R. The effective ampere-turns of the control windings, proportional to the actual difference of the full-

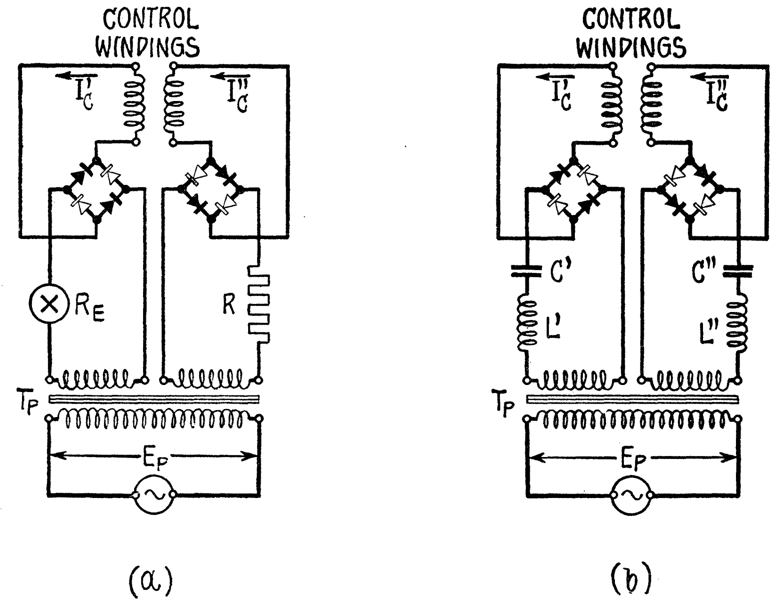

FIG. 18.7. A-c differential-type networks supplying the twin control windings of a magnetic servo amplifier, without electronic-amplifier components. (*a*) Voltage-sensitive circuit with a nonlinear resistance R_E and a standard resistor R; (*b*) frequency-sensitive network with two oscillating circuits $C'L'$ and $C''L''$, which are tuned to different frequency values, below and above the rated value of the regulated frequency, respectively.

wave rectified currents I'_C and I''_C, are a function of the actual deviation of the regulated value of voltage E_P from the rated value.

Figure 18.7*b* shows a frequency-sensitive differential network which is rated so that the oscillating circuits $C'L'$ and $C''L''$ are tuned to different frequency values, below and above the rated value of frequency f_P, respectively. In this case, the effective ampere-turns of the control windings, proportional to the actual current difference $I'_C - I''_C$, are a function of the actual deviation of the regulated frequency f_P from the rated value.

Similar arrangements may be used for providing temperature-sensitive or photoelectric control of magnetic amplifiers. Application of various types of "transducers," consisting of variable resistors, reactors, or condensers, makes it possible to introduce nonelectrical quantities into the control circuit of such amplifiers. There are a great many different ways in which such devices can be built up, as illustrated in the special literature concerning the techniques of measurement and control.

Furthermore, it is to be noted that magnetic servo amplifiers have been successfully used in the output-stage circuits of self-balancing d-c potentiometer recorders and also in "coordinate ink recorders" plotting automatically a continuous curve which represents the relationship $x = f(y)$ between any two variables (*e.g.*, dynamic hysteresis loops).[12] Similar applications are possible in the field of self-balancing a-c bridge and potentiometer circuits, which have proved to be of considerable practical value for testing of high-voltage equipment[13] and instrument transformers.[14]

18.5. Special Applications of Magnetic-amplifier Circuits. H. B. Rex[15,16] has shown that the fundamental principle of saturable-reactor devices, without or with feedback, may be applied in the field of *electrodynamic* arrangements, *e.g.*, to a-c motors and electrical instruments.

A-c motor control[15] may be achieved by means of a special type of polyphase induction motor, wherein a regulatable high-permeability magnetic shunt is utilized to control the field flux as impressed upon the armature. Another possibility consists in using a d-c-controlled saturable magnetic shunt to provide pole shading in variable-speed reversible motors of the induction or hysteresis type.

Improvements in electrical measuring instruments[16] of the dynamometer or movable-vane type may be achieved by making use of the magnifying effect of the electric current in a feedback circuit which links the core components of the instrument, to enable the measurement of electrical currents of minute values. The basic idea consists in providing an electrical measuring instrument that contains, integrally within itself, a means for amplifying the current to be measured. There are numerous possibilities for introducing this interesting principle in the field of indicating and recording instruments, which may be combined with various kinds of transducers.

Further contributions concerning special applications of magnetic-amplifier circuits have been made by J. L. Wolff,[17] F. N. McClure,[18] L. J. Johnson and H. G. Schafer,[19] T. R. Specht and R. N. Wagner,[20] H. E. Larson and T. Dunnegan,[21] F. S. Malick,[22] H. M. Ogle,[23] W. Jellinghaus,[24] and S. E. Hedström,[25] and others.[26]

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Author Index

Subject Index

D

E

M

P

Q

R

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