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ERRATA SHEET

DESIGN METHODS FOR HIGH FREQUENCY TRANSFORMERS

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8-8	Col. 1, par. 1, line 2	Temperatures indicate	Temperatures up to 200°C. indicate
9-1	Col. 1, par. 2, lines 8 and 9, three extra words	as accompanied by	-----(deleted)
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14-40	Col. 2, Equation 64	$p^2 =$	$g^2 =$
14-41	Col. 2, Fig. 14-31b, resistor	-----	R_p
14-54	Col. 1, par. 4, line 8	side skirt selectivity	skirt selectivity
14-55	Col. 1, par. 1, line 8	sacrified	sacrificed
14-56	Col. 1, Example, line 11	$L_1=L_2=7.79$μh	$L_1=L_2=$ 0.79μh
14-56	Col. 1, item 8, line 8	we find Δf_p	we find Δf_b
14-60	Col. 1, Footnote 1, Relocate to pg. 14-59	-----	----- (deleted)
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14-64	Col. 2, Equation 2	Z = [illegible]	Z = [illegible]
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14-67	Col. 1, par. 1, line 6	βA	[illegible]
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14-68	Col. 1, Equation 98	$A_0 = A/(1 + \beta A)$ (98)	$A_0 = A/(1 - \beta A)$
A-1	Introduction to Appendix	A-1	xv
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DESIGN METHODS FOR HIGH FREQUENCY TRANSFORMERS

CODE FOR COIL TREATMENTS AS ILLUSTRATED IN FIGS. 7-1, 7-5, 7-8 on pages 7-2, 7-10, 7-12 respectively

T1	Wax
T2	Synthetic Baking Varnish
T2(v)	Synthetic Baking Varnish (Vacuum Impregnated)
T2 & T3	Synthetic Baking Varnish 2 coats ea. of 2 types
T3	Synthetic Baking Varnish
T3(v)	Synthetic Baking Varnish (Vacuum Impregnated)
T4	Synthetic Baking Varnish
T4(v)	Synthetic Baking Varnish (Vacuum Impregnated)
T6	Coil Impregnant (100% Resin) Composition Unknown
T7	Casting Material (Styrene Base)
T8	Casting Material (Epoxy Base)
T8(v)	Casting Material (Epoxy Base) Vacuum Impregnated
T9	Styrene Base Air Drying Lacquer
T9(2B)	Styrene Base Air Drying Lacquer (2 brush coats)
T12	Epoxy Resin
T14	Casting Material (Epoxy base-modified)
T15	Casting Material (Epoxy base) Mica filled
T16	Silicone Rubber
T17	Cement (Composition Unknown)
T18	Cement (Composition Unknown)
T19	Cement (Composition Unknown)
T20	Cement (Composition Unknown)
T22	Epoxy Resin-modified
T23	Epoxy Resin-modified
T24	Polyester Resin-modified
T22	Polyester Resin
T23	Epoxy Resin-modified

CODE FOR WIRE INSULATION AS ILLUSTRATED IN FIG. 1-2, page 1-6

W-1	Polyvinyl Acetal Film
W-2	Polyvinyl Acetal Film
W-3	Composition unknown - listed by manufacturer as being suited for operation at temperatures up to about [illegible]°C.
W-4	Polyvinyl Acetal Film with outer coating of thermoplastic material which will bond to itself under heat or solvent action.
W-5	Enamel
W-6	Nylon Enamel Film
W-7	Composition unknown - solderable type film insulation.
W-8	Nylon Enamel Film
W-10	Ceramic film insulation with polytetrafluoroethylene.
W-11	Polytetrafluoroethylene Film
W-12	Silicone Enamel Film - Recommended by manufacturer for operation up to [illegible]°C.
W-13	Combination of polyvinyl acetal film with nylon enamel film.
W-14	Single Silk Enamel
W-15	Single Nylon Enamel
W-16	Single Celanese Enamel
W-17	Single Orlon Enamel
W-18	Single Cotton Enamel
W-19	Double Silk Enamel
W-20	Silicone Enamel Film

Footnote:
More than one type number with identical classifications indicates product of additional suppliers.

CODE FOR COIL FORM MATERIALS AS ILLUSTRATED IN FIG. 5-3, page 5-4

CF-1	Paper, Waxed
CF-2	Paper
CF-3	Styrene Copolymer
CF-4	Paper, Phenolic Treated
CF-5	Steatite
CF-6	Cordierite
CF-7	Forsterite
CF-8	XXX Molded Phenolic
CF-9	XX Phenolic, Rolled
CF-10	Paper, Phenolic Treated
CF-11	Nylon
CF-12	Polytetrafluoroethylene
CF-13	Type [illegible]LF-M Plastic
CF-14	Type Z-1 Plastic
CF-15	Type G-5 Plastic
CF-16	Vulcanized Fibre
CF-17	XX Phenolic, Molded
CF-18	XXX Phenolic, Rolled

DESIGN METHODS FOR HIGH FREQUENCY TRANSFORMERS

OBJECT
DEVELOPMENT OF SIMPLIFIED PRINCIPLES
OF DESIGN AND STUDY OF MATERIALS OF
CONSTRUCTION FOR HIGH FREQUENCY TRANSFORMERS

SIGNAL CORPS ENGINEERING LABORATORIES
FORT MONMOUTH, N.J.
CO-SPONSORED BY
THE ELECTRONIC COMPONENTS LABORATORY
OF THR WRIGHT AIR DEVELOPMENT CENTER
WRIGHT-PATTERSON AIR FORCE BASE, OHIO

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Squier Signal Laboratory
Technical Requirements
Dated 8 January 1953, For
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POOR ORIGINAL

DESIGN MANUAL FOR I. F. TRANSFORMERS

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PREFACE

Each year, of many new engineers and technical assistants entering the electronics field, some will be involved with the application and design of r-f transformers and coils. This manual on "Design Methods for High Frequency Transformers" is directed in particular to these newcomers and to those already engaged in electronics who are confronted with the many complex problems relating to r-f inductors. It comprises an attempt to bring under one cover, to as great a degree as possible, some simple explanations of the basic fundamentals of coil design and construction.

Inductive components are unique among the many families of electronic parts. Whereas resistors, capacitors, switches, etc., are available as standard stock parts having established characteristics, the audio, power, pulse, r-f and i-f transformers and other specialized coils used in electronic equipment invariably must be designed for a specific application.

Since World War II considerable attention has been given to the development and refinement of the techniques of design of audio, power and pulse transformers. As a result, a wealth of straight-forward, practical design information is available. In contrast, the r-f coil design field has had no concentrated effort aimed specifically at relating the highly analytical text-book approaches to the practical problem of building a coil or transformer. As a result, the design of radio frequency coils is still practiced more as an art than as a science. Many formal text books are excellent in their scientific treatment of the subject but the designer needs, in addition, the experience of those long established in the art in order to translate into a practical design the scientific principles presented.

Literature contains many analytical articles on various specialized phases of this art but these exist as unrelated efforts and are difficult to quickly locate and utilize when a problem is presented. It is time consuming and confusing for each engineer to individually conduct the literary research necessary to establish the threshold to this specialized branch of electronics design. Years of apprenticeship are often required to establish a mature level of practical and academic "know-how".

Realizing this, the Wright Air Development Center, Wright Patterson Air Force Base, proposed the preparation of a treatise on high frequency transformers. This was implemented through the Signal Corps Engineering Laboratories on Contract No. DA-36-039-SC-52679 which has resulted in this manual.

The experience and "know-how" previously only available through association or years of experience are presented herein by chart and example in condensed form. A complete discussion of each element of a coil or transformer such as wire, shield, magnetic materials, etc., is included along with a section devoted to Theory and Design. This section outlines new approaches which are recommended to the engineer who has had little, if any, experience in the design of r-f coils.

Certain products are more commonly known in industry by "trade names" rather than by their technical or chemical designation. These terms have, in some instances, been used in this manual. Many vendors and their products are mentioned directly for purposes of illustration. There has been no attempt to completely survey the field. This does not imply an endorsement or preference of a particular product by either the government or the contractor.

POOR ORIGINAL

Many people have contributed in one way or another to this manual. It would be impossible to acknowledge each of these contributions. Acknowledgments to specialists who have contributed to specific chapters have been included at the end of those chapters. Special recognition is given to Mr. George C. Sziklai of the R.C.A. Laboratories, Princeton, New Jersey, for his advice and comments relating to the section on Theory and Design. Recognition is also given to Messrs. S. Danko and D. Elders of the Signal Corps Engineering Laboratories and Messrs. G. Tarrants and D. Crockett of the Wright Air Development Center. Acknowledgment is also made particularly to Philip J. Reich for his valuable editorial comment and to other members of the Automatic Manufacturing Corporation and F. W. Sickles Division of General Instrument Corporation Engineering Departments for their technical suggestions. Thanks are also due to the Boonton Molding Company, the Institute of Radio Engineers, Inc., and the McGraw-Hill Book Company, Inc., for permission to reproduce various charts and other information as noted throughout the manual.

Even though the manuscript has been reviewed with painstaking care it is recognized that some errors may appear. If any should be found the authors sincerely hope you will call them to their attention so that subsequent printings may be corrected.

AUTOMATIC MANUFACTURING CORPORATION

Part I

MATERIALS OF CONSTRUCTION

INTRODUCTION

The engineer confronted with the design of military r-f inductive components must be familiar with the rigorous environmental conditions imposed by military service and with the very practical and unique designs often employed in civilian components which are responsible for economical mass production. With this knowledge of military requirements and commercial practices, he can develop practical, economical and satisfactory designs suitable for military service.

The various sections of Part I contain discussions on the materials of construction peculiar to r-f inductive components. Subjects covered include conductors, shields, magnetic core materials, electronic hardware, ceramics, plastics, impregnating materials (waxes, varnishes, etc.), tapes and film insulations, and finishes. It is the purpose of Part I to fully acquaint the user with all of the critical elements that are used in r-f transformer construction.

Supporting data, which is graphical in many cases, is included to assist in the selection of the proper materials for a given application. The practical aspects, including suggestions for preparing parts specifications for procurement along with established commercial tolerances, are fully covered. It is recommended that Part I of this manual be carefully studied to provide a working background for the Design Theory presented in Part II.

It must be remembered that the r-f coil design art is a fast changing one and that design practices and materials used to-day may be superseded by newer developments tomorrow. It is suggested that the coil designer keep abreast of all new developments through the medium of literature such as technical articles, vendors' catalogues and data books. This documented source of information should be supplemented by frequent contact with suppliers and manufacturers and with other development groups where exchanges of technical information will provide an up-to-date design background.

ATTACHMENT NO. 7

Section 1

CONDUCTORS

GENERAL

As a general rule, coils are wound of insulated copper conductors commonly known as magnet wire. Because of its ductility, copper may be drawn through dies into the form of rods and/or filaments of a size in conformance with specification JAN-W-583 (similar to that provided by the National Electrical Manufacturers' Association (NEMA). After being drawn, the wire is annealed to give it elongation properties suitable for winding into coils.

Size is most often expressed in American Wire Gauge (AWG) numbers. These numbers are so arranged that a larger number denotes a smaller wire with each gauge number approximating the successive steps in the wire drawing and every sixth smaller number representing a wire with a doubled diameter. In the electronics industry, the range of sizes usually falls between No. 14 with a diameter of 0.0641 inch and No. 44 with a diameter of 0.0020 inch. Special applications may involve wire as small as No. 50 with a diameter of only 0.0010 inch. (A complete copper wire table appears in the Appendix of this manual.)

In a few highly specialized cases, conductors of aluminum, silver, or resistance metals are employed. Limited use, particularly in the higher frequencies, is found for conductors which are in the form of ribbon. Silver plated copper is also used in many high frequency applications because of its lower resistance. Electro-deposited metals —commonly copper or silver—are also becoming important as coil conductors, particularly in printed circuit applications.

Bare copper wire is rarely used in electronics because of the danger of shorted turns and also because of the fact that unprotected copper very quickly acquires an oxide coating which makes it difficult to solder. Where an uninsulated wire is specified, the choice is invariably copper which has been run through a hot tin bath, thereby producing what is called *tinned copper wire*.

FILM INSULATED WIRES

Insulations applied to bare copper wire are of two basic types. The most common are insulations of the "film" type such as enamel, Formex[1] (polyvinylformal), nylon, and other specialized insulations. Insulations of this general type are characterized by high dielectric strength and will be found to possess various degrees of abrasion and solvent resistance.

ENAMEL

The most common film insulation is plain enamel which consists of an oleoresinous varnish. The film is applied in multiple coats by running the wire at controlled speeds through a varnish of low viscosity followed by baking in a continuous oven. Enamel is commonly applied in vertical coating machines without the use of dies, although some manufacturers do use dies when enameling the larger sizes. Electrically, this is one of the better film insulations, possessing good dielectric strength, hardness, adhesion to the copper, and film flexibility. In addition, enamel films are resistant to most acids and alkalies and have remarkable moisture resistance. When thoroughly cured, they are but slightly affected by varnish solvents of the petroleum types or by neutral mineral oil. Lack of abrasion resistance is the most serious defect since it greatly limits the applications in which enameled wire may be used without an additional protective coating—usually a textile serving. When served wires are used, it is the enamel which provides the moisture resistance and the dielectric strength, while the textile serving protects the enamel film and spaces adjacent turns of the winding.

VINYL ACETAL

One of the most popular film insulations in current use is the polyvinylformal film sold under the

[1] Manufactured by General Electric Company.

trade names of "Formex" or "Formvar"[1] – terms which will be used interchangeably throughout the balance of this discussion. The varnish which forms this film is based on the synthetic organic resin, vinyl acetate, and also contains a phenolic resin which serves as a heat stabilizing and hardening agent. This varnish is applied directly to the copper from a solvent solution, usually in horizontal coating machines. Special dies limit the amount of varnish which remains on the wire, and the addition of multiple coats insures an extremely uniform build-up of the insulating film.

Formex wire is made in four grades–single, heavy, triple, and quadruple. Compared to enamel wire of the oleoresinous type, Formex has much greater resistance to abrasion, exceptional film flexibility, and far better solvent resistance. In the opinion of many engineers, its electrical charatertistics are not quite so good, particularly at temperatures in the vicinity of 75 C, but the slight loss in Q when coils are wound with Formex coated wires is more than offset by the improvement in abrasion and solvent resistance and by the lowered tendency to crack when bent around small diameters.

One property of polyvinyl acetal films is commonly known as *solvent crazing* and is of special significance in the case of coils which are to be varnish impregnated. Solvent crazing takes place when Formex coated wires in which the insulating film is under strain–usually as the result of bending–are placed in a solvent which wets the surface of the film. Under these conditions, what seem to be minute cracks appear in the Formex film. Actually, there is some question as to whether these marks are cracks in the conventional sense, since they do not penetrate through the film to the copper conductor. Tests, however, do indicate that the dielectric strength of solvent-crazed Formex is substantially lowered, and it is therefore recommend that coils wound with Formex be annealed prior to the application of any varnish or similar treatment material.

Annealing consists simply of heating the coils before application of any treatment material for a period of time varying from five minutes at 105 C to one hour at 80 C. Once cracks due to crazing have occurred, it is somewhat more difficult to heal them, and a cycle of one half hour at 150 C is generally accepted as being required. Exhaustive tests seem to indicate that no attention need be given to solvent crazing in those instances where the varnish treatment receives a baking cycle of at least two hours at 125 C.

In moisture resistance, acid and alkali resistance, and in dielectric strength, Formvar compares favorably with enamel, but its improved abrasion resistance accounts for its great popularity throughout the electrical industry.

It is this same high abrasion resistance, coupled with its good adherence to copper, that has brought about one of the major problems facing the electronics industry today–the removal of Formex film from fine wire. In the larger sizes – which in the electronics industry means No. 30 or larger – this is less of a problem, since, if the wire is passed quickly through a small gas or alcohol flame, the insulation may then be easily removed by rotating wire or glass filament brushes, emery paper, or others means. The larger sizes, particularly No. 25 and larger, may be cleaned by dipping the wires in a solder pot filled with 50/50 solder and operating at a temperature of not less than 500 C. This method has the added advantage of providing a freshly tinned surface on the cleaned copper, making subsequent soldering operations much easier. The real difficulty in removing Formvar comes in the smaller sizes such as No. 38, No. 39, and No. 40, all of which are commonly used in high frequency transformers. Many methods have been evolved, ranging from actual chemical attack to the use of glass filament and wire brushes. Opposition to the use of chemicals is great because of the fear that some ionizable material will be left on the wire surface following the cleaning process. Should this occur, it would constitute an invitation to corrosion and electrolysis. (See Section 8 for more detailed discussion of electrolytic corrosion.) Of those methods of removing Formex which are currently in effect, the one which seems safest and best but is by no means foolproof, involves the use of rotating brushes, preferably of the glass filament type. The wide acceptance of Formex and Formvar by the electronics industry is largely due to their toughness, resistance to moisture and solvents, and the fact that, properly impregnated, they can be used continuously at temperatures as high as 125 C. These and other recognized properties of vinyl acetal insulated wire account for its choice as the standard of comparison for the temperature coefficient studies performed in support of this manual.

[1]Supplied by Belden Manufacturing Company, Chicago, Illinois; Hudson Wire Company, Winsted Division, Winsted, Connecticut; Phelps-Dodge Copper Products Corporation, Fort Wayne, Indiana; Warren Wire Company, Pownal, Vermont; Wheeler Insulated Wire Company, Waterbury, Connecticut; and many others.

"SOLDERABLE" INSULATIONS

Because of the serious difficulties encountered in removing Formex from the copper and also because plain enamel itself is somewhat difficult to remove, a need developed within the industry for a so-called "solderable" wire. In answer, a number of formulations have appeared on the market, ranging from applications of cellulose acetate lacquer to extruded nylon coatings and nylon varnish films. Unfortunately, solderable wire insulations are generally lacking in abrasion resistance and in temperature stability, and their use is recommended only for single layer windings or for applications where performance is secondary to cost. As may be expected, those coatings made up of cellulose lacquer formulations are low in solvent resistance, and particular care must be taken in treating these wires to avoid dissolving the film insulation. A vast amount of research is under way on solderable film insulation, and the design engineer will do well to keep in close contact with the magnet wire manufacturers as indications are that satisfactory, easily solderable insulations will soon appear on the market. A list of some of the currently available solderable magnet wires and their manufacturers appears in Fig. 1-1.

Fig. 1-1 TYPICAL SOLDERABLE MAGNET WIRES

(Film Insulated Type)

Trade Name	Manufacturer*
Celenamel	Belden Manufacturing Company
Dipsol	Wheeler Insulated Wire Company
EZ Sol	Hudson Wire Company, Winsted Division
Nylon Enamel	Rea Magnet Wire Company
Nylonel	Warren Wire Company
Sodereze	Phelps-Dodge Copper Products Corporation
Nylon Varnish	Essex Wire Corporation

*Addresses may be found at the end of this section.

HIGH TEMPERATURE INSULATIONS

The growing demand among users of electronic equipment for coils capable of withstanding much higher operating temperatures has resulted in the appearance of a number of new film insulations. This problem has been attacked in two ways: by the use of inorganic ceramic coatings and by the use of organic materials such as Teflon and Silicone.

Ceramic coatings by themselves have not been entirely satisfactory, especially on fine wire. When combined with materials such as Teflon, magnet wires capable of continuous operation at temperatures in excess of 200 C have been successfully produced. Teflon (known chemically as polytetrafluoroethylene) is characterized by exceptionally high chemical resistance and by an ability to operate over wide temperature ranges. Its electrical characteristics are good, particularly at higher frequencies, and its moisture resistance is exceptionally high. When Teflon is used by itself in coating magnet wire, the resulting wire is so smooth and slippery that its use in winding coils of the universal type often presents rather serious problems. When applied directly over ceramic insulations, the surface is less smooth, making winding somewhat easier.

Enamel coatings based on silicones are presently becoming available. Recommended by their manufacturers[1] for use at temperatures up to 180 C, these wires are so new that it is difficult at this time to assess their true value to the industry. Information available at this writing would seem to indicate that a very satisfactory high temperature film insulation will shortly be on the market in the form of silicone enamels.

[1]Hitemp Wires, Inc., 26 Windsor Avenue, Mineola, New York; Hudson Wire Company, Winsted Division.

SPECIAL PURPOSE INSULATIONS

Numerous other film insulations are available and are included in the temperature coefficient results shown in Fig. 1-2. Many of these insulations were developed to fill a particular need as, for example, those wires which are actually Formvar covered with a thermoplastic material[4] which, under heat or solvent action, softens sufficiently to permit the turns of a winding to bond one to another. Wires insulated with alternate films of vinyl acetal and nylon[5] also typify attempts of the wire industry to combine the good points of two of their insulations and, at the same time, supply their customers with a more satisfactory product.

EFFECT OF WIRE INSULATIONS UPON TEMPERATURE COEFFICIENT OF UNIVERSAL COILS

#39 WIRES
CERAMIC FORM 1/2" O.D. X 3/8" I.D. X 2" LONG
CAM 1/16
INDUCTANCE 1.275 MH ± 3 %
NO IMPREGNATION

NOTE: PRIOR TO TEST ALL WINDINGS GIVEN 20 ALTERNATE 15 MINUTE EXPOSURES TO -12 AND +85C.
ALL VALUES ARE POSITIVE AND REPRESENT AVERAGES OF 6 COILS.

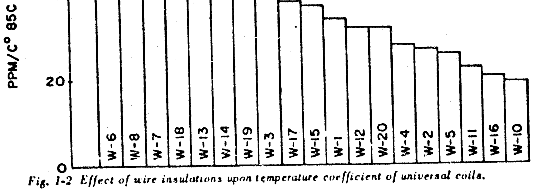

Fig. 1-2 Effect of wire insulations upon temperature coefficient of universal coils.

[4] Formbond — Acme Wire Company, New Haven, Connecticut
Bondez — Essex Wire Company, Fort Wayne, Indiana
Bondere — Phelps-Dodge Copper Products Corporation

[5] Nylclad — Belden Manufacturing Company
Nyform — Warren Wire Company

LITZ WIRE STUDY
Q AND O.D.
CAM 3/32" FORM 1/2" O.D. X 3/8" I.D. X 2" LONG
INDUCTANCE 1.725MH ± 3 % TEST FREQ. 455 KC
NO IMPREGNATION
NOTE: ALL WIRE SINGLE SILK ENAMEL EXCEPT 5/44 BWG (APPROXIMATELY EQUAL TO 5/40 AWG.) WHICH WAS SERVED WITH SINGLE RAYON.
*37 EQUIVALENT IN CIRCULAR MILS TO 5/44 A.W.G.
ALL VALUES ARE AVERAGES OF 5 COILS.

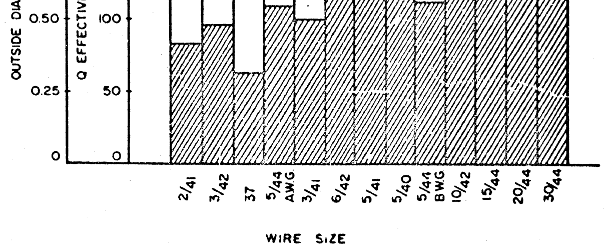

Fig. 1-3 Effect of wire upon OD and Q of universal coils.

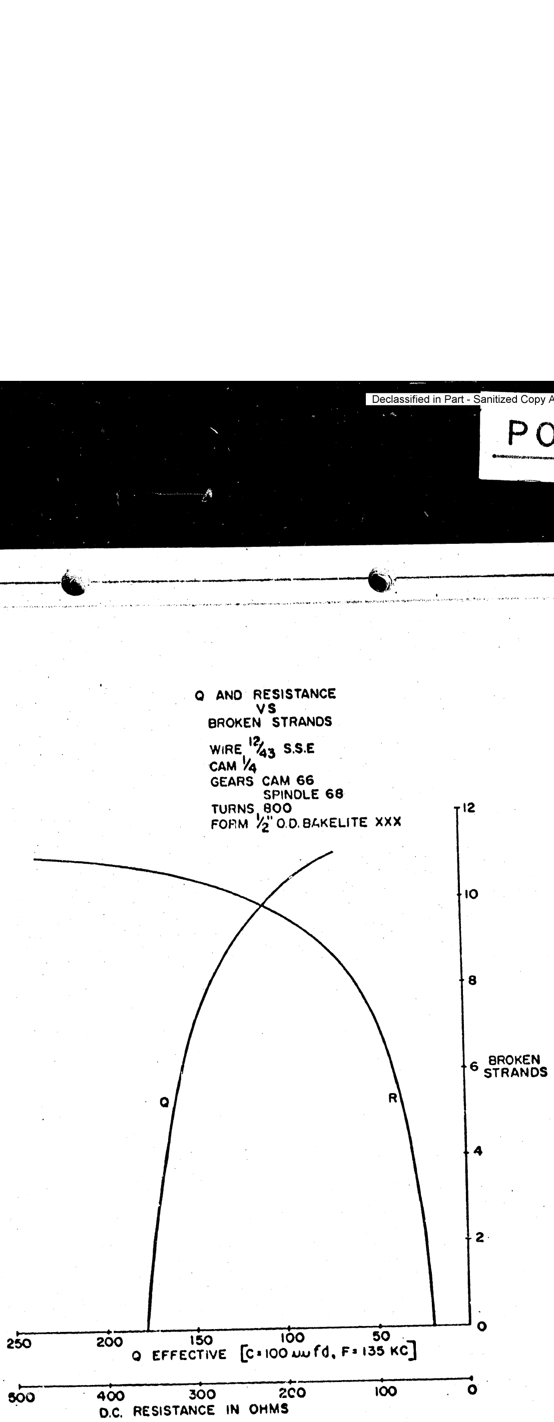

Fig. 1-4 Effect of broken strands upon Q and resistance of universal coils

LITZ WIRES

At frequencies up to 2000 kc, Litz wire is widely used in coils where high Q is of primary importance. Litz wire consists of a number of strands of very small wire, each strand insulated from the other. The insulation most commonly used is enamel, but Formex Litz is available on special order. Most commonly, the strands of insulated wire are enclosed within a textile wrap, but Litz wire without a textile serving and even without a means of bonding between strands has been used. In general, better results are obtained with the use of textile served Litz. However, its use greatly increases the size of the winding and is, therefore, impractical for miniature and subminiature applications. (See Fig. 1-3)

Fig. 1-5

EFFECT ON Q OF NUMBER OF TWISTS PER FOOT IN LITZ WIRE

Twists/ft. (5/44 SSE)	54/106 Gears Cam 1/4" Form OD = 1/2" 1100 turns		67/44 Gears Cam 3/32" Form OD = 1/2" 300 turns		51/48 Gears Cam 1/16" Form OD = 0.175" 250 turns	
	Q	L in mh	Q	L in mh	Q	L in mh
Commercial Grade	89	19.75	113	1.72	76	0.47
Parallel	84	19.75	108	1.74	75	0.48
18 twist/ft	87	19.70	109	1.72	74	0.47
65 twist/ft	85	19.45	110	1.74	76	0.46

NOTE: OD's of separate strands measured over the enamel
(Limits 0.00200 - 0.00230)

Commercial Grade	Parallel	18 T/ft	65 T/ft
0.0022	0.0021	0.0020	0.0020
0.0021	0.0021	0.0021	0.0020
0.0022	0.0021	0.0020	0.0021
0.0022	0.0020	0.0020	0.0021
0.0021	0.0021	0.0020	0.0020

Throughout the years that Litz has been used, considerable disagreement has been noted among users as to the relationship the various strands should bear with respect to one another. It is possible to buy so-called Litz wire in which the strands are simply bunched, and any twist that they may assume is the result of having the textile wrap placed about them. Other manufacturers make their Litz wire with a definite number of twists per foot, usually somewhere between 8 and 36.

Tables appearing throughout this discussion show the results of tests conducted on various types and sizes of Litz and solid wires and are intended to give an idea of the effect of these various wires upon the electrical characteristics of universal coils.

TEXTILE COVERED WIRES

General: The thickness of film insulation which can be placed on a wire is definitely limited, and because many applications require an appreciable spacing between adjacent turns, a textile serving is frequently placed on the wire as a means of obtaining this spacing. The textile may be applied to bare copper wire or to film insulated wire with the latter being far more common since the textile serving itself is of little actual insulation value. Of the various textile servings which are applied to wire, silk is probably the best known and enjoys the widest use in the electronics field. Other textiles used for wire serving are nylon, orlon, and celanese.

In each case, the textile is applied to the wire by a wrapping action. The effect is to enclose the conductor within a continuous spiral of textile ribbon. These ribbons are made up of a certain number of "ends" or strands, each of a given "denier" or size. (The term *denier* is borrowed from the silk industry and is a measure of fineness of textile fibers with a smaller number indicating a smaller fiber). Silk is available in much finer deniers than are the synthetic fibers, 20 denier silk is relatively common, while 80 denier is the finest orlon available at this writing. The use of more ends of fine denier fibers results in better coverage and a more flexible wrapping, therefore automatically giving silk an advantage over synthetic fibers in difficult winding applications. In the case of double wrapped wires, the second layer is applied in a direction opposite to the first—the idea being to eliminate as completely as possible open spaces in the wrap.

Many problems are connected with textile served wires, not the least of which is a satisfactory method of measuring the outside diameter (OD) of the served wire. Many methods have been suggested, but the one most commonly used is dependent upon hand micrometers, closed to a point where the wire can be dragged through the opening with a recognizable amount of resistance. The very means of stating this method of measurement indicates the amount of personal touch involved. Surprisingly enough, there is excellent correlation among the readings of experienced operators once the necessary skill has been acquired.

Another serious problem connected with the use of textile—served wire concerns the variations in a particular wire as supplied by different manufacturers. These differences are largely the result of employing different angles of lead and different amounts of tension in introducing the textile ribbon onto the surface of the conductor. A long angle of lead—that is with the textile more parallel to the conductor—results in a "soft" wire, often showing a tendency for the textile to open up when the wire is bent. Because this type of wire is soft, it is difficult to use on a narrow winding of large diameter since the resulting mechanical structure is springy and subject to collapse.

When the textile serving is applied at nearly right angles to the conductor, the resulting wire is much less flexible and, in general, presents a harder surface when compared to the more loosely wrapped wire. Carried to an extreme, this type of wrap can work-harden the copper to a point where the wire becomes too stiff and too brittle to wind without breaking. Unfortunately, no standards—military or civilian—include any reference to the way textile servings shall be applied to the wire other than to give minimum and maximum builds and to include references to skips and barberpoling. It follows, therefore, that the product of one manufacturer may be definitely superior to that of another when used in a specific application. So great may this difference be that it actually may be necessary to change the set-up of a winding machine when changing from the wire of one supplier to that of another.

NYLON SERVING

The use of nylon-served wire may occasionally introduce some unusual situations in winding. Nylon tends to be slippery and in addition is elastic to the point where tension applied to the winding causes nylon to stretch in a fashion similar to a rubber band. While this is taking place, the copper is being elongated (Specification JAN-W-583 requires a minimum elongation of 7.5 to 35 per cent depending upon insulation and size), and when the winding is completed and the tension released, the nylon tends to spring back, whereas the copper has taken a permanent set. The result is a winding which tends to "explode"—a term more descriptive of the result than of the act—particularly when the winding is of a high and narrow type. When this action occurs, the wire will stick out through the textile wrap in a series of loops. This phenomenon does not occur in the case of silk—served wires, since the silk fibers lack the elasticity of the nylon.

ORLON SERVING

Orlon, one of the newer synthetic fibers, is slowly coming into use as a substitute for nylon. In winding characteristics, orlon-served wire closely resembles silk except as noted below and is slightly better than nylon in its electrical characteristics. A major trouble with orlon at this time is a lack of tensile strength in the fiber which often allows the textile to break when going through the tension devices and other guides leading the wire onto the winding form. Once a method is developed for overcoming this weakness, it is likely that interest will develop rapidly in orlon-served wires.

CELANESE SERVING

Celanese like orlon, is low in cost compared to other served wires. Also like orlon, celanese yarn is low in tensile strength and is therefore difficult to wind on conventional winding equipment. In many instances, the high percentage of rejects at winding traceable to breaks in the celanese yarn will far more than offset the lower initial cost of the wire. An added feature of this type of wire is that the nature of the serving makes it possible to solder, without removing the textile, provided, of course, that the serving was applied over bare or solderable type wire.

Another point to be considered in the case of celanese served wire is its low resistance to solvent attack. The yarn used in wrapping this type of magnet wire is a form of cellulose acetate rayon which, therefore, is readily attacked by nearly all common solvents. This property of celanese-covered wire requires particular care in the selection of impregnations and other treatments subsequent to winding as well as in the selection of the cements used to start and terminate the coils. The presence of acetone in either instance is an immediate invitation to the disintegration of the textile—a condition sometimes deliberately introduced as a means of producing a self-supporting winding.

One feature of wires covered with celanese yarn is emphasized in Fig. 1-2 where it is shown that no other served wire will produce universal coils with so high a degree of temperature stability as will celanese-served wire. The reason for this greater stability is not immediately apparent, but repeated tests in every instance have shown similar results.

In view of the obvious disadvantages as well as advantages to be found in the use of celanese-served wire, it is recommended that specification of this type of serving should come only after careful weighing of the relative merits of celanese and and other available textile servings.

UNIFORMITY OF COVERAGE

The textile serving should be continuous over the surface of the conductor. The applicable NEMA standard (NW21-1953 Section 5.2.2 "Coverage of Silk") states that "the silk-covered wire shall be wound around the mandrel having a diameter equal to ten times the diameter of the bare wire under sufficient tension to insure an even compact layer. After being so wound, the silk covering shall not open sufficiently to expose the bare wire on the film or the film-coated wire when examined with normal vision." Normal vision is defined in a footnote as "20-20 vision after correction with eyeglasses if necessary". In actual practice it is difficult to purchase wires completely free from "skips" or "barberpoling"—skips being occasional open spots in the wrap, while barberpoling indicates a serving applied in an open spiral with the conductor clearly visible between the turns. In general, barberpoling is recognized as a basis for rejection of served wires, although instances are on record where this type of wire has been specified for reasons of space and/or cost.

MOISTURE RESISTANCE OF TEXTILE COVERED WIRES

The use of textile-served wire complicates the procedures necessary to protect a winding against moisture, since regardless of the type of treatment used, the textile fibers serve as a wick through which moisture may travel to the interior of the coil. Reference to Fig. 1-6 will show that in all instances, textile-served wires exhibited less resistance to humidity than did wires insulated only with film coatings.

COST

Cost-wise, textile serving is an expensive procedure. In Fig. 1-7 are shown comparative costs of the various types of wires based on prices in effect during October 1953. At first glance, it may appear that the difference in cost between plain enameled wire and single silk enameled wire is excessive, but it must be considered that an average serving machine required 23.9 hours* to serve one pound of No. 39 single silk enameled wire.

The period since World War II has seen a significant decline in the demand for textile-served wires. This statement is not meant to imply that textile-served wire no longer occupies a prominent place in electronics, but rather that increased emphasis on cost, a definite swing toward miniaturized coil components, improvements in winding techniques and equipment, and in film insulation have, during these years, added to the attractiveness of the non-textile served wires.

SELECTION OF WIRE

Selection of the proper wire for a particular coil must be based on several factors, including size of the end product, its operating frequency, Q, type of winding, operating temperature, humidity requirements, temperature stability requirements, impregnation, and cost. In nearly every case, some

*This figure courtesy The Wheeler Insulated Wire Company.

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compromise is necessary. This is especially true in the design of miniature and subminiature coils where space limitations may demand the use of a film-insulated wire regardless of all other factors. An idea of the comparative size of equal inductance windings made from various served and unserved wires may be gained from Figs. 1-3 and 1-12.

upon the insulation. Universal windings, on the other hand, require a wire possessing good abrasion resistance together with the ability to stand up under the pressures resulting from the winding process and the coil structure. These pressures are of considerable magnitude at the points of crossover since the nature of a universal winding requires the wire to cross at regular intervals while

Fig. 1-6 COMPARATIVE MOISTURE RESISTANCE OF TEXTILE-SERVED AND FILM-INSULATED WIRES

Wire	Initial Q	Q Measured 1/2 hour after Humidity[2]	Per cent of Q Remaining
3/41 SSE	39	30	70
3/41 SSE	39	29	67
3/41 SSE	39	30	70
3/41 SSE	39	30	70
6/42 HF	35	31	89
6/42 HF	36	31	86
6/42 HF	35	31	89
6/42 HF	34	31	91

NOTES:

[1]All coils treated with one coat of synthetic baking varnish followed by one coat of silicone baking varnish.

[2]96 per cent relative humidity and 40 C for 200 hours.

Where space is not particularly limited and where emphasis is on Q or voltage breakdown, a textile-served wire is indicated. If Q is of the greatest importance, silk is the logical choice with orlon, celanese, and nylon following in respective order. When voltage breakdown is the chief concern—for example, in bifilar windings—the order would probably change to celanese, nylon, orlon, and silk simply because of the relative thickness of the servings.

Choice of a particular type of winding may directly influence the selection of wire because of proximity of turns and/or mechanical stresses resulting from the winding process. In a space-wound solenoid, any wire—even bare wire—may safely be used. Close-wound solenoids may use any insulated wire whose covering is electrically satisfactory, since once in place, there is no mechanical strain under winding tension. To be satisfactory under these conditions, the insulation must afford maximum mechanical protection and exhibit a minimum of cold flow to prevent shorts at the crossover points.

Best suited for universal windings are wires with a textile serving applied over either enamel or Formvar. If space does not permit the use of a textile-served wire, the designer's next best choice is heavy or triple Formvar or one of the nylon-Formvar combinations. Wires of the solderable type are, however, generally undesirable because of their tendency to short at the crossover points within the windings. Plain enamel wire is also generally unsatisfactory for universal windings because of its inability to withstand the scraping action involved in the winding process.

If it is known that the transformer must operate

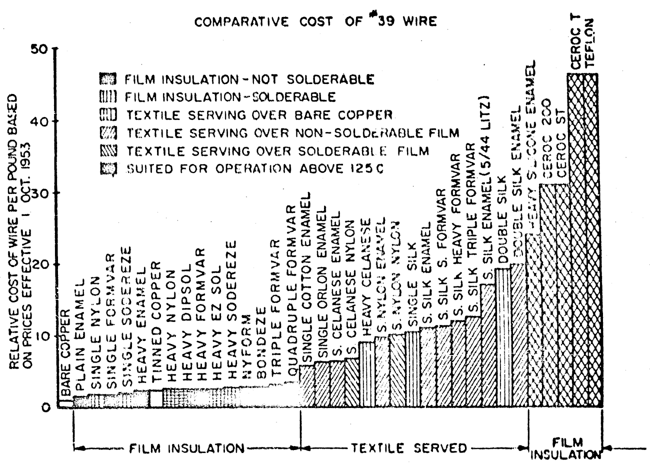

Fig. 1-7 Comparative cost of various types of No. 39 magnet wire.

under conditions of high humidity, the use of a textile-served wire is not recommended, regardless of its subsequent impregnation. Film insulations in general, particularly those of Formex or enamel, are definitely superior to any textile-served wire when subjected to either static or cycling humidity exposure. No treatment has yet been found which will effectively seal the fibers of the served wire and prevent the movement of moisture along these fibers toward the interior of the coil.

When 85 C is the maximum operating temperature of a transformer, the designer has complete freedom of choice in selection of wire insulation. Raising the operating temperature to 125 C begins to limit the choice since all insulations that are thermo-plastic in nature are unsatisfactory in this temperature range. The use of enamel is not recommended since it is at about this temperature that the film begins to disintegrate from the action of heat. Formvar, while not rated by its manufacturers as being satisfactory for use above 105 C, has been found to operate successfully at 125 C when protected by an adequate impregnation, such as the dual varnish treatment recommended for maximum moisture protection and described in Section 7 of this manual. For units intended to operate above 125 C, a designer is, for the most part, limited to wires insulated with Teflon, ceramic materials, or combinations of the two. It is entirely possible that the new silicone enamels will prove satisfactory in this range, but too little is known of them at this time to warrant a definite recommendation.

Reference to Fig. 1-2 will give an indication of the degree of temperature stability which may be expected from coils wound with various types of insulated wires. An indication of the solvent resistance of various insulations is presented in the table appearing as Fig. 1-9 which may be used as a guide in the selection of compatible impregnation

Fig. 1-8 RECOMMENDED MAGNET WIRE INSULATIONS FOR THREE MAJOR TEMPERATURE CLASSIFICATIONS[1]

85 C	125 C	200 C
Enamel	Silicone enamel	Ceramic
All textile servings	Vinyl acetal[2]	Teflon
All solderable films		Ceramic plus Teflon
Vinyl acetal		Silicone enamel[1]

NOTES: [1]Any wire listed in a higher group may, of course, be used safely in a lower group.

[2]Not recommended by their manufacturers for operation in this classification. However, tests conducted during the preparation of this manual indicate that with proper impregnation these wires may be used as shown above.

Fig. 1-9 SOLVENT RESISTANCE OF COMMON FILM INSULATED WIRES*

SOLVENT	ENAMEL	FORMVAR	NYLON
Naptha	Poor	Very good	Very good
Kerosene	Poor	Very good	Very good
Alcohol	Fails	Good	Very good
Xylol	Fails	Good	Very good
Acetone	Fails	Good	Very good
5% H_2SO_4	Very good	Very good	Fair
Gasoline	Fails	Very good	Very good
Benzene	Poor	Very good	Very good
Toluol	Fails	Good	Very good
Ethyl Acetate	Fails	Fair	Very good
Boiling Ethanol/Toluol	Fails	Fails	Good
Cresylic Acid	Fails	Poor	Fails (dissolves)
Ammonia	Poor	Good	Very good
Carbon Tetrachloride	Fails	Very good	Very good
1% KOH	Very good	Very good	Very good

*This table compiled at Automatic Manufacturing Corporation from information supplied by Phelps-Dodge Copper Products Corporation and Wheeler Insulated Wire Company.

materials. In this connection, it should be stressed that under certain circumstances it is perfectly possible for a wire insulation to soften in the presence of the solvent and still be acceptable for use if not subjected to stress while softened and if subsequent treatments insure complete removal of the solvent. Because of the difficulties often encountered in identifying various wire insulations, a series of simple identification tests have been worked out and incorporated in the tables appearing as Figs. 1-10 and 1-11.

The importance of magnet wire in high frequency transformer design is great. Fortunately for the design engineer, the major wire manufacturers have excellent product information available and will be found willing to lend their "know-how" in new and special cases. Close contacts with the representatives of these various companies will be most valuable.

Fig. 1-10 IDENTIFICATION TESTS FOR FILM INSULATIONS*

TEST	FILM INSULATION		
	ENAMEL	FORMVAR	NYLON
Dip in Acetone	Film softens in very few minutes	No effect	No effect
Dip in 600 to 700 F Solder Pot	No effect	No effect	Wire tins
Dip 950 to 1050 F Solder Pot	Enamel may crumble but wire will not tin	Wire tins	Wire tins
Apply small flame	Burns with black smoke Leaves black surface	Burns with black smoke Leaves black surface	Melts and burns leaving clean copper
Dip in Cresylic Acid	Film softens and is easily removed	Film softens slightly	Dissolves
Dip in boiling mixture of 30% toluol and 70% denatured alcohol	Film softens and is easily removed	Film softens and is easily removed	No effect

*This table compiled at Automatic Manufacturing Corporation from information furnished by representatives of Belden Wire Company, Phelps Dodge Copper Products Corporation, Wheeler Insulated Wire Company, and Winsted Division of Hudson Wire Company.

Fig. 1-11 SIMPLE IDENTIFICATION TESTS FOR TEXTILE SERVINGS*

TEST	TEXTILE			
	SILK	NYLON	ORLON	CELANESE
Dip in Acetone	No effect	No effect	No effect	Dissolves
Dip in Cresylic Acid	No effect	Dissolves	No effect	———
Strip from conductor and bring near small flame.	Burns. Ash dark and easily crumbled. Odor resembles burning feathers.	Melts. May burn. Forms hard resin ball of greyish-tan color. Odor resembles burning flesh.	Melts. May burn in flashes. Leaves hard, slightly gummy ball, dark in color.	———
Dip in 60% H_2SO_4	———	Dissolves	No effect	———

*This table compiled at Automatic Manufacturing Corporation from information furnished by representatives of Belden Wire Co., Phelps Dodge Copper Products Corp., Wheeler Insulated Wire Co. and Winsted Division of Hudson Wire Co.

Fig. 1-12 COMPARATIVE SIZE OF COILS[1] WOUND WITH FILM-INSULATED AND TEXTILE-SERVED WIRES

L in mh	Coil OD in inches	
	No. 39 HF	No. 39 SSE
53.0	1.000	———
50.5	———	1.430
20.4	———	1.130
18.0	0.790	———
11.0	———	0.980
10.0	0.722	———
4.7	———	0.827
4.4	0.649	———
1.2	———	0.672
1.1	0.577	———

[1]Coil data: Form - OD; 1/2 inch, Wire size - No. 39, Gears - Formex 59/58, SSE 100/66

BIBLIOGRAPHY

Young, James F.
Materials and Processes, Eighth Printing
John Wiley & Sons, Inc., New York, 1949

CATALOGS AND TECHNICAL INFORMATION OF:

Anaconda Wire and Cable Company
25 Broadway
New York, New York

Belden Manufacturing Company
Chicago 80, Illinois

The Electric Auto-Lite Company
Point Huron, Michigan

General Electric Company
Construction Material Department
Bridgeport, Connecticut

Hitemp Wires, Inc.
26 Windsor Avenue
Minneola, Long Island, New York

Hudson Wire Company
Winsted Division
Winsted, Connecticut

Phelps-Dodge Copper Products Corporation
Fort Wayne, Indiana

Rea Magnet Wire Company
Fort Wayne, Indiana

Sprague Electric Company
North Adams, Massachusetts

Warren Wire Company
Pownal, Vermont

SPECIFICATIONS

JAN-W-583(3)
"Wire, Magnet"

QQ-W-341a(2)
"Wire; Copper, Soft or Annealed"

Various specifications of the American Standards Association sponsored by National Electrical Manufacturers Association of which examples are:

C9.1-1955
"Enamel-Coated Round Copper Magnet Wire"

C9.3-1953
"Silk-Covered Round Copper Magnet Wire"

C9.4-1953
"Nylon-Fibre-Covered Round Copper Magnet Wire"

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Fort Wayne, Indiana

Section 2

SHIELDS

REASONS FOR SHIELDING

To operate successfully, modern electronic equipment must be so constructed that coupling between the various circuits is limited to the amount intended by the designer of the equipment. Essentially, this requirement can be met by confining within a limited space the electromagnetic and electrostatic fields which surround any inductance through which a current is flowing. Because both of these fields tend to link (couple) readily with other similar fields, it follows that coupling may be either *inductive* as a result of the electromagnetic field, or *capacitive* as a result of the electrostatic field. It is usually necessary to prevent both types of coupling, and the means most often employed is that of shielding the inductive components.

Shielding, as practiced in electronics, usually consists of enclosing an inductive component within a metallic container called a *shield can*. These containers are usually made of a metal having a relatively high conductivity. Aluminum is the most common shield material with copper and zinc being used for those cases where it is necessary or desirable to solder directly to the can. Sometimes iron or steel is used although it is not a common practice at radio frequencies.

ELECTROMAGNETIC SHIELDING

Electromagnetic fields may be confined in two ways: (1) by the use of conducting shields of non-magnetic material or (2) by the use of high-permeability, low-reluctance magnetic shields.

In the case of conventional shield cans made from low resistance, non-magnetic metals, the shielding (reduction in inductive coupling) is largely the result of eddy currents induced in the metal can. The energy used to form these currents is drawn from the field of the inductance to which the shield bears somewhat the relationship of an untuned secondary, thus creating a loss in the enclosed winding which shows up as an increase in the effective resistance of the coil and a subsequent lowering of its Q.

Since the shielding of a magnetic field is an eddy current phenomenon, it is apparent that unless these currents can flow freely wherever they please, the shielding will not be effective. This means that shield cans must be made from low-resistance materials, free from breaks or high-resistance joints. In other words, if shielding is to be effective, there must be a continuous, low-resistance path through which eddy currents can flow with complete freedom. Were it not for this fact, shield cans made up of metal foil interspaced between layers of paper could provide adequate, low-cost shielding. That this is not the case can be easily demonstrated by using copper foil as a liner in steel shield cans – a shielding procedure which will be found completely ineffective at radio frequencies until the overlapping portion of the copper foil is soldered throughout its length.

Eddy currents which are set up in shield cans will be found to be in opposition to the fields of the enclosed windings and therefore will act to reduce the effective coil inductance. It is for this reason that it is always necessary to specify the conditions under which inductance readings have been taken. The effect to which "in shield" and "in air" readings may vary is illustrated by measurements made on a conventional 455 kc intermediate frequency transformer. Measured in air, this transformer had a primary inductance of 2.026 mh, and a secondary inductance of 2.042 mh, but when enclosed within its shield L_p became 1.986 mh and L_s became 2.008 mh – an average loss of approximately 2 per cent in inductance. Mutual inductance between the two windings was also affected and to an even greater extent since it measured 128 microhenrys in air and only 105 microhenrys when enclosed in the shield can. It proba-

Part I MATERIALS OF CONSTRUCTION

bly should be noted at this point that variations in inductance and Q can be introduced not only by shielding but also by placing a coil in close proximity to such metallic objects as mounting brackets, a chassis, core screws, or other similar masses of metal.

Magnetic shielding may also be accomplished through the use of cups or sleeves of powdered iron, ferrite, or other suitably high-permeability, low-reluctance material. In such cases, external coupling is reduced because the magnetic flux is concentrated in the low-reluctance path which is placed about the coil. Unlike the shielding resulting from eddy currents, this type tends to raise both the inductance and the Q of the enclosed windings. In general, magnetic shielding is not particularly effective in the reduction of extraneous coupling, and it is customary in the design of transformers utilizing this type of shielding to enclose the complete assembly in a conventional shield can despite the presence of magnetic cores or sleeves. In such instances the outer shield can serves primarily as an electrostatic shield since an amount of flux sufficient to generate eddy currents rarely reaches the outer can but instead stays within the low-reluctance path of the magnetic material.

The discussion up to this point has been primarily concerned with electromagnetic shields. Since, however, the basic requirement for electrostatic shielding is to enclose by a conducting surface the space to be shielded, it will be seen that the use of conventional shield cans provides electrostatic shielding as well as electromagnetic shielding. The electromagnetic shielding is, of course, the result of the eddy currents which are set up in the shield and which oppose the passage of the flux lines. The continuous conducting path provided by the shield can is sufficient to prevent capacitive coupling through the electrostatic field.

In this connection, it should be noted that a solid conductive screen is not necessary for electrostatic shielding and that a grid-like structure of the general type shown in Fig. 2-1 will be satisfactory for this purpose. Because only one end of the conductors making up this device is connected to the common bus, there is no opportunity for the formation of circulating currents, and therefore there is little or no effect upon inductive coupling. Such an arrangement is known as a *Faraday Screen*, and examples may be found in many modern commercial receivers where the screens are often made by printed circuit techniques.

Faraday Screens must, of course, be grounded if they are to be effective. The principal purpose of these devices is to furnish a means of eliminating capacitive coupling while at the same time permitting inductive coupling — a condition which can result from the insertion of a properly grounded screen in such a manner as to seperate and enclose the windings of a transformer.

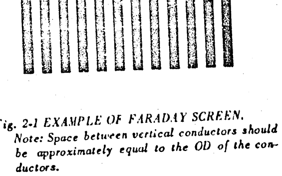


Fig. 2-1 EXAMPLE OF FARADAY SCREEN.
Note: Space between vertical conductors should be approximately equal to the OD of the conductors.

FACTORS AFFECTING SHIELDING

Several factors may be said to influence the overall effectiveness of shielding. If the shield material and its thickness remain constant, frequency will have a direct influence upon the efficiency of the shielding, since increased frequency means increased eddy currents which in turn mean better shielding. When the frequency remains constant and the metal is not changed, the effectiveness of shielding increases as the thickness of the shield increases. Actually, this latter condition is not a linear function, and experience has indicated that at common r-f frequencies little is to be gained by increasing the thickness of an aluminum shield beyond the normal 0.018 to 0.020 inch. Heavier shields may, however, be required at lower frequencies.

It must be remembered that the efficiency of a shield is directly related to the conductivity of the metal used in the fabrication of the shield. This means that copper cans are more effective than those made of aluminum, although for average applications aluminum is perfectly satisfactory as is evidenced by its almost universal acceptance in all equipment except the most precise, as, for example, standard signal generators. Reference to Fig. 2-2 will provide an indication of the com-

parative effectiveness of various materials when made into shield cans of uniform size and checked over a wide range of frequencies.

In the design of high-gain amplifiers, care should be taken to avoid direct contact between shield cans since there may be considerable coupling between stages if the shields are in contact at any point. For those cases requiring maximum isolation, it is desirable rather than to increase the thickness of the shield cans to use two or more separate shields located one inside the other with contact between the two limited to one point, if possible. It is this type of double shielding which has proved most successful in the manufacture of standard signal generators where stray couplings are of the utmost importance.

Among the effects of shielding which should be mentioned is the increase in distributed capacitance that is always noted in shielded windings. This increase in C_d is essentially an electrostatic phenomenon which occurs because a shield is at ground potential while at least a part of the winding is always substantially above ground. Since distributed capacitance is a factor which influences the self-resonance of a winding as well as the relationship between its true and apparent inductance, it therefore becomes clear that it is difficult to predict accurately the effective inductance of a shielded coil. It is, however, equally obvious that the closer the shield approaches the coil, the greater will be the difference between the true and the apparent inductance of an enclosed winding.

Not only is the size of the shield important when viewed from the standpoint of its electrostatic effects, but it must be remembered that when metal of any sort is moved closer to an energized winding, the amount of magnetic field that will enter into that metal is increased. Most engineers consider it to be an accepted fact that this entrance of energy into the walls of the shield can and the resulting eddy currents formed therein will show up as a circuit loss and that a shielded coil will lose both in inductance and in Q.

EFFECT OF SIZE AND SHAPE OF SHIELD CANS

In the course of laboratory work performed as background for this manual, a substantial amount of study was devoted to the effect upon enclosed windings of variations in the shape, size, and material of shield cans. As can be seen from the graphs and tables throughout this section, the effect of shielding is not one which is clear-cut but rather is one which is dependent upon a number of factors including frequency of operation, core material used in the inductance, and the *Q in air* of the enclosed winding. Reference to Fig. 2-3 will show that at frequencies of 30 Mc or higher, it is entirely possible for a coil to gain as much as 30 per cent in Q when enclosed within a relatively close-fitting aluminum shield. It is important to note in this regard that air-core coils do not respond to shielding in this manner at any frequency between 455 kc and 180 Mc. Only those coils having iron cores show this property, and here again it should be noted that the mere presence of an iron core is insufficient basis for this behavior. Only certain kinds of iron cores induce a response of this sort, thus indicating something of the general difficulties involved in predicting the performance of shielded coils.

As will be seen from the experimental data accompanying this section, there is good reason to accept the oft-quoted rule of design that "a shield can should never come closer to a nonmagnetically shielded inductor than a distance equal to the diameter of the coil itself", thus pointing out the importance of cup cores or other magnetic shielding in miniature and subminiature transformer design.

Because the selection of the size and shape of transformer shield cans is more often dictated by the available space in the end equipment than by those factors constituting optimum coil design, it follows that good transformer design practice should start with the shield since it is necessarily a limiting factor in the physical size of the completed unit. A study of the sizes of shield cans presently available from established manufacturers lends credence to the theory that all too often engineers design a transformer and then as their last move design a shield can to fit their new creation. One major shield manufacturing company[1] reports that it has on hand approximately 150 sets of drawing tools representing an investment in the order of \$300,000 — a figure which is easy to understand since a single set of tools may cost anywhere between \$1500 and \$3000. This figure of \$300,000 does not include piercing tools — those tools which punch the required holes and other openings in the cans. No estimate as to the number of piercing tools owned by this company was available other than that for one particular shield can (an item used in large quantities on military equipment) twelve different sets of piercing tools were in current operation at the time this survey was made. Ad-

[1] Paul & Beekman, Inc., Philadelphia, Pennsylvania.

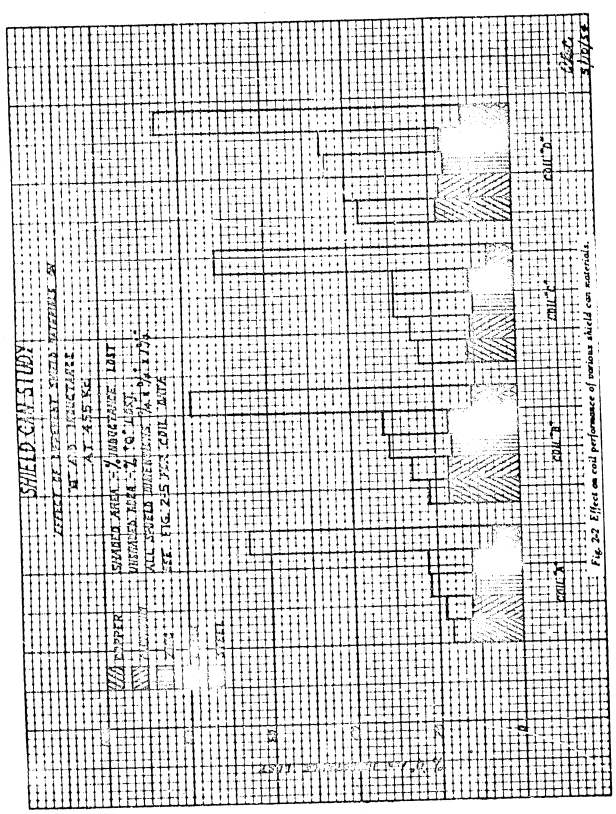
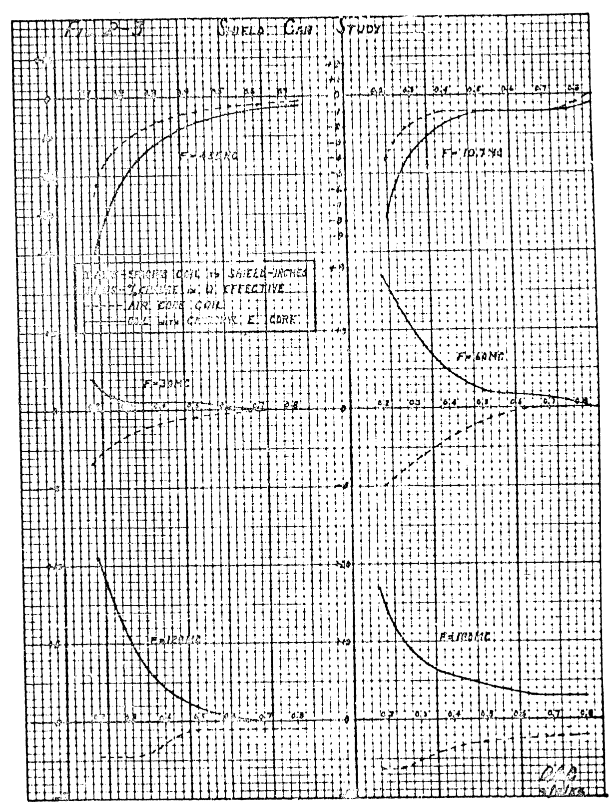


Fig. 2-2 Effect on coil performance of various shield can materials.



mittedly, piercing tools are far less expensive than drawing tools, but the fact remains that had more thought been given to the basic design of this series of coils, production costs could have been substantially lowered through a reduction in the cost of tooling.

METHODS OF FABRICATION OF SHIELD CANS

Shield cans may be made in a number of ways with the most important and generally satisfactory method being known as *drawing*. This operation is carried out in multiple stage presses utilizing strip stock which is blanked in the first stage and then is progressively formed into shapes and sizes more nearly approaching the final form as it passes through each successive stage in the drawing operation. In some tools, provision is made in the final stage for piercing and cutting to length, whereas in other instances these two operations are performed on separate equipment after the shield has been drawn. It is worthy of note that drawn shields are very uniform in size and in wall thickness, and that they have an end thickness equal to that of the stock from which they were drawn.

A second manufacturing process which at one time was of considerable importance in shield can production is known as *extrusion*. In this method, the can is formed from a predetermined mass of metal which is placed in a cavity having the size and shape of the can which is to be formed. A ram having the dimensions of the inside of the can then enters the cavity and by tremendous pressure actually causes the metal to flow upward into the space between the ram and the cavity wall, thus forming the shield can. This method is used today by some manufacturers for small sizes of round cans but, in general, it has been replaced by drawing. Extruded shields can easily be recognized by the thick closed ends which are always present — a somewhat undesirable condition inasmuch as it makes piercing that much more difficult. Another point in which extruded shields are inferior is found in the nonuniformity of sidewall thickness which is a characteristic of the extrusion process.

Round shields are sometimes spun,[3] but like casting — a method once used in certain instances — this method of making shield cans offers no advantage great enough to warrant its additional cost. In view of the trend toward miniaturization, it is highly improbable that spun shields will again become a factor of any importance in the electronics industry.

[3]See Section 4 for description of the spinning process.

DIMENSIONS AND TOLERANCES

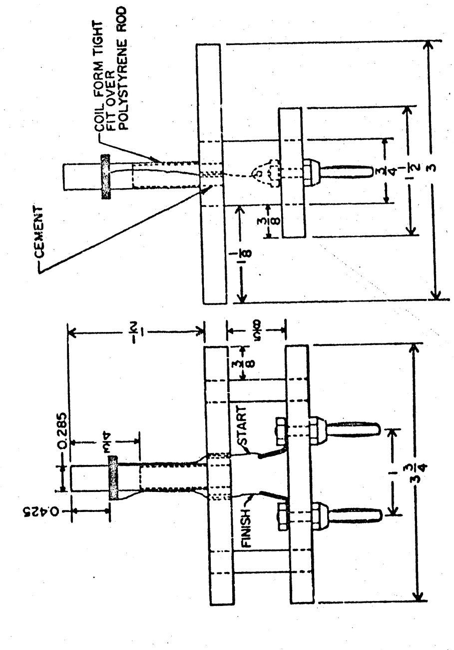
The most critical dimensions in shield design are the internal radii. For fairly obvious reasons, it is not possible to produce a drawn shield can with perfectly square corners on either the inside or the outside of the shield. It is, however, desirable to keep these radii as small as possible in order to utilize most fully the space within the shield. Since, however, tool cost and tool maintenance are both influenced by the size of the radii specified, it is generally accepted that 1/16 inch is the smallest practical radius that should be specified in shields of approximately 0.750 inch inside dimension. Larger shields, of course, demand larger radii with 1.250 inch shields requiring 7/64 inch radii for economical and satisfactory production. Both inside and outside dimensions have been used at one time or other in specifying shield sizes, but the best and most widely accepted practice seems to be to work with inside dimensions on cross sections and with outside dimensions on length. Commonly accepted tolerances are ±0.003 to 0.005 inch on the cross section; ±0.003 inch on wall thickness; and either ±0.008 or $^{+0.020}_{-0.000}$ inch on the length. Most can manufacturers will find these tolerances acceptable without additional cost. To specify closer tolerances will inevitably require special tools, special handling, and additional expense.

Up to this time, very few serious attempts have been made to standardize on shield can sizes. At various times designers have specified shields which in cross section were round, rectangular, oval, or square. Because of a desire to conserve chassis space, the recent trend has been away from round and oval shapes toward either rectangular or square shields. Probably the nearest to a "standard" size in use today is the so-called "3/4 inch" which actually measures 0.735 inch square on the inside. Other popular sizes have inside dimensions of 1.125 inches and 1.375 inches, and it is upon those three sizes that the majority of the experimental work for this section was based. As these words are written (toward the end of 1954), there is good evidence that a new subminiature size of square cross section having inside dimensions of 0.500 $^{+0.006}_{-0.000}$ inch will become popular.

METHODS OF MOUNTING

When installed in a piece of equipment, shield cans must be firmly connected to the chassis both electrically and mechanically since shielding be-

FIG. 2-4 Q-METER JIG FOR SHIELD CAN STUDY



MATERIAL: POLYSTYRENE 1/4 THICK
SCALE 1:1

Fig. 2-5 Description of various coils used in shield can studies

SHIELD CAN STUDY

COIL DATA

	COIL "A"	COIL "B"	COIL "C"	COIL "D"
Wire	No. 39 H.F.	No. 39 H.F.	No. 5/44 SSE	No. 5/44 SSE
Cam	3/32	3/32	3/32	3/32
Gears	56/74	56/74	51/67	51/67
Turns	500	500	255	255
End Spacing	0.425"	0.425"	0.425"	0.425"
OD	0.516"	0.516"	0.523"	0.523"
Coil Width	0.115"	0.115"	0.120"	0.120"
Coil Form OD	0.285"	0.285"	0.285"	0.285"
Core	None	Plast-Iron B-231	None	Carbonyl E
Impregnation	Wax	Wax	Wax	Wax
Frequency	455kc	455 kc	455 kc	455 kc
Coil "Q" (No Shield)	59	74	79	122
C	46.91 uuf	27.93 uuf	252.09 uuf	155.91 uuf

	COIL "E"	COIL "F"	COIL "G"	COIL "H"	COIL "I"
Wire	No. 36 SSE	No. 36 SSE	No. 30 SSE	No. 30 SSE	No. 30 SSE
Winding	Solenoid	Solenoid	Solenoid	Solenoid	Solenoid
Turns	10	10	15	15	16
End Spacing to Coil Center	0.511"	0.511"	0.528"	0.528"	0.512"
End Spacing to Coil	0.457"	0.467"	0.421"	0.421"	0.463"
OD	0.301"	0.301"	0.310"	0.310"	0.523"
Coil Width	0.087"	0.087"	0.214"	0.214"	0.213"
Coil Form OD	0.285"	0.285"	0.285"	0.285"	0.500"
Core	None	Carbonyl E	None	Carbonyl E	None
Impregnation	Wax	Wax	Wax	Wax	Wax
Frequency	10.7 Mc	10.7 Mc	10.7 Mc	10.7 Mc	10.7 Mc
Coil "Q" (No Shield)	64	92	95	116	114
C	184.42 uuf	103.60 uuf	141.07 uuf	68.58 uuf	54.61 uuf

SHIELD CAN STUDY

COIL DATA

	COIL "M"	COIL "N"	COIL "O"
Wire	No. 20 H.F.	No. 20 H.F.	No. 20 H.F.
Winding	Solenoid	Solenoid	Solenoid
Turns	7½	7½	7½
End Spacing to Coil Center	0.503"	0.503"	0.503"
End Spacing to Coil Edge	0.358"	0.358"	0.358"
OD	0.371"	0.371"	0.371"
Coil Width	0.290"	0.290"	0.290"
Coil Form OD	0.285"	0.285"	0.285"
Core	None	Carbonyl E	Carbonyl C
Impregnation	Wax	Wax	Wax
Frequency	30 Mc	30 Mc	30 Mc
Coil "Q" (No Shield)	143	99	40.8
C	67.05 uuf	35.30 uuf	32.98 uuf

	COIL "P"	COIL "Q"	COIL "R"	COIL "S"
Wire	No. 20 H.F.	No. 20 H.F.	No. 20 H.F.	No. 20 H.F.
Winding	Solenoid	Solenoid	Solenoid	Solenoid
Turns	3½	3½	3½	3½
End Spacing to Coil Center	0.499"	0.499"	0.499"	0.499"
End Spacing to Coil Edge	0.427"	0.427"	0.427"	0.427"
OD	0.362"	0.362"	0.362"	0.362"
Coil Width	0.143"	0.143"	0.143"	0.143"
Coil Form OD	0.285"	0.285"	0.285"	0.285"
Core	None	Carbonyl E	Carbonyl C	IRN8
Impregnation	Wax	Wax	Wax	Wax
Frequency	60 Mc	60 Mc	60 Mc	60 Mc
Coil "Q" (No Shield)	158	76.0	27	158
C	34.49 uuf	22.85 uuf	21.25 uuf	28.62 uuf

SHIELD CAN STUDY

COIL DATA

	COIL "T"	COIL "U"	COIL "V"	COIL "W"
Wire	No. 20 H.F.	No. 20 H.F.	No. 20 H.F.	No. 20 H.F.
Winding	Solenoid	Solenoid	Solenoid	Solenoid
Turns	1½	1½	1½	1½
End Spacing to Coil Center	0.501"	0.501"	0.501"	0.501"
End Spacing to Coil Edge	0.465"	0.465"	0.465"	0.465"
OD	0.358"	0.358"	0.358"	0.358"
Coil Width	0.072"	0.072"	0.072"	0.072"
Coil Form OD	0.285"	0.285"	0.285"	0.285"
Core	None	Carbonyl E	Carbonyl C	HRN8
Impregnation	Wax	Wax	Wax	Wax
Frequency	120 Mc	120 Mc	120 Mc	120 Mc
Coil "Q" (No Shield)	171	76.5	30	150
C	12.79 uuf	10.52 uuf	10.41 uuf	11.90 uuf

	COIL "X"	COIL "Y"	COIL "Z"	COIL "1"
Wire	No. 17 H.E.	No. 17 H.E.	No. 17 H.E.	No. 17 H.E.
Winding	Solenoid	Solenoid	Solenoid	Solenoid
Turns	1½	1½	1½	1½
End Spacing to Coil Center	0.496	0.496	0.496	0.496
End Spacing to Coil Edge	0.437	0.437	0.437	0.437
OD	0.392"	0.392"	0.392"	0.392"
Coil Width	0.095	0.095	0.095	0.095
Coil Form OD	0.285"	0.285"	0.285"	0.285"
Core	None	Carbonyl E	Carbonyl C	HRN8
Impregnation	Wax	Wax	Wax	Wax
Frequency	180 Mc	180 Mc	180 Mc	180 Mc
Coil "Q" (No Shield)	238	50	17.5	158
C	10.25 uuf	7.91 uuf	7.53 uuf	8.76 uuf

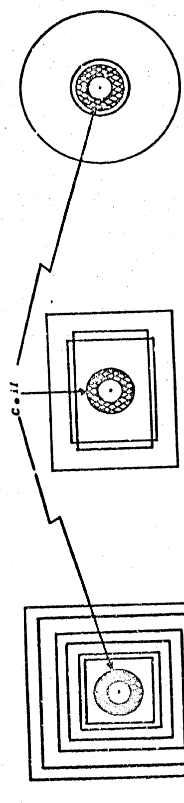


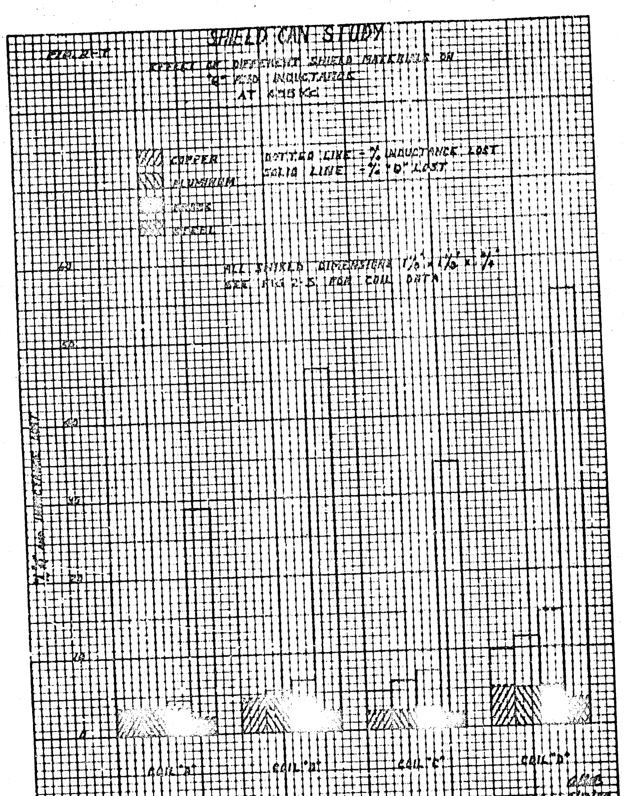
Fig. 2-6 PERCENTAGE "Q" LOST DUE TO SHIELD SIZE

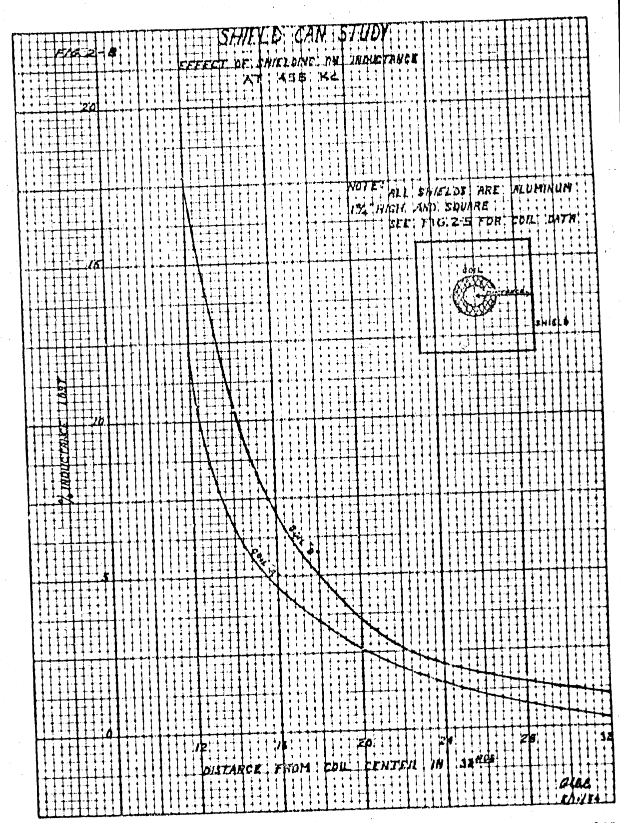
	1 x 1"	7/8 x 7/8"	1-1/8x1-1/8"	1-3/8x1-3/8"	1¼x1¼"	2 x2"	13/16x1-3/8"	13/16x 1-3/16"	1-3/8 x 13/16"	5/8" OD	1¼" OD
COIL "A"	19.1%	10.4%	4.4%	2.6%	1.7%	2.6%	8.0%	5.9%	1.7%	55.7%	2.6%
COIL "B"	25.8%	13.5%	5.4%	2.6%	1.3%	0.6%	10.7%	6.7%	2.0%	63.5%	1.3%
COIL "C"	25.3%	13.9%	7.5%	3.7%	2.5%	1.2%	11.4%	8.1%	2.5%	63.9%	2.5%
COIL "D"	41.0%	25.4%	12.2%	6.5%	3.2%	2.3%	20.5%	15.5%	4.8%	75.6%	4.8%

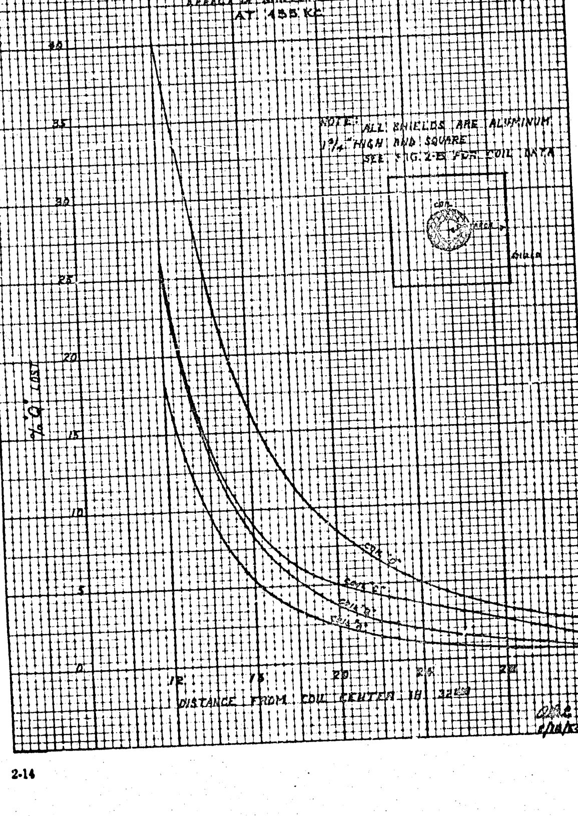
ALL SHIELDS ARE ALUMINUM 1-3/4 inches HIGH AND 0.017-0.024 inches THICK

SEE FIG. 2-5 FOR COIL DATA

SCALE 1:1


SHIELD CAN STUDY
EFFECT OF DIFFERENT SHIELD MATERIALS ON "Q" AND INDUCTANCE AT 455 KC
COPPER
ALUMINUM
STEEL
DOTTED LINE = % INDUCTANCE LOST
SOLID LINE = % "Q" LOST
ALL SHIELD DIMENSIONS 1½ x 1½ x ¾
SEE FIG 2-5 FOR COIL DATA
COIL "A"
COIL "B"
COIL "C"
COIL "D"


SHIELD CAN STUDY
FIG 2-8
EFFECT OF SHIELDING ON INDUCTANCE AT 455 KC
NOTE: ALL SHIELDS ARE ALUMINUM 1¾" HIGH AND SQUARE
SEE FIG. 2-5 FOR COIL DATA
COIL
SHIELD
% INDUCTANCE LOST
20
15
10
5
0
12
16
20
24
28
32
DISTANCE FROM COIL CENTER IN 32NDS



comes less effective whenever resistance is introduced between the shield can and the chassis. If this resistance varies under operating conditions, noise may be introduced into the circuit. For many years, the conventional method of mounting shield cans was by the use of spade bolts which were riveted to the shield cans and attached to the chassis through the use of nuts and lock washers. Recent advances in mechanical design have produced various kinds of spring mounting devices intended as replacements for spade bolts and purporting the advantages of being far less expensive and much faster to install on the production line.

A successful example of this type of transformer mounting is the U-shaped, spring mounting clip developed and patented by Automatic Manufacturing Corporation, Newark, New Jersey for use with its K-Trans. The design of this clip is such as to assure a permanent, non-oxidizing contact between the shield can and the chassis as well as a means of mounting which satisfies even the strenuous requirements of the Navy shock test.[3]

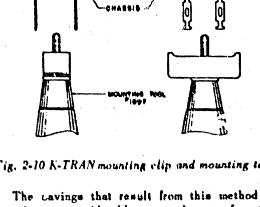


Fig. 2-10 K-TRAN mounting clip and mounting tool.

The savings that result from this method of mounting are considerable as may be seen from the fact that spade bolt mounting requires the use of a total of 10 small parts[4] compared to the single mounting clip described above and pictured in Fig. 2-10. A further advantage is to be found in the ease with which shield cans designed for use with with this clip can be mounted and demounted. The simple snap-action of the clip responds readily to the simplest of tools, and with proper care the clips may be used over and over again — a point which could be of considerable importance when making repairs under emergency field conditions.

A number of other types of spring-actuated mounting devices have been made available from commercial sources. While offering certain advantages over spade bolts, especially in the case with which they may be attached to a chassis, these devices all require riveting to the shields with the consequent handling of a minimum of 6 small parts. This fact in itself reduces the attractiveness of these devices which, while adequate for many civilian applications, are not believed to be sufficiently sturdy for the average military requirement.

Since the shield can is an essential part of a high-frequency transformer, and since such transformers will operate successfully only when firmly attached to the chassis, it follows that for those units requiring shield cans of a size other than the "3/4 inch", the most reliable mounting method is that involving the use of spade bolts. If miniature components are being used, it would seem wise to give consideration to the obvious advantages attached to the U-shaped, spring mounting clip.

DESIGN SUMMARY

From the foregoing discussion of shields, it would seem that good high frequency transformer design practice calls for —

1. The use of standard sizes of drawn aluminum shield cans supplemented by magnetic shielding in miniature and subminiature units or where extremely high Q must be obtained in small spaces.
2. Specification of normal commercial tolerances on all dimensions including internal radii.
3. The use of a substantial mounting method consisting of spade bolts, nuts, and lock washers for the larger sizes of shield cans and either the same or the patented U-shaped spring mounting clip for "3/4 inch" cans.

[3] Demonstrated at Signal Corps Engineering Laboratories during the course of work on Signal Corps Research and Development Contract No. DA-36-039-SC-15321.

[4] Consisting of 2 spade bolts, 2 rivets, and 2 washers to attach the spade bolts to the shield, plus 2 nuts and 2 lock washers to mount the shield to the chassis.

TEMPERATURE CONVERSION TABLE

The numbers in italics refer to the temperature in either centigrade or Fahrenheit which is to be converted to the other scale. To convert Fahrenheit to centigrade, read the left hand column. To convert centigrade to Fahrenheit, read the right hand column.

-73.3	*-100*	-148.0	-48.9	*- 56*	- 68.8	-24.4	*- 12*	10.4
-72.8	*- 99*	-146.2	-48.3	*- 55*	- 67.0	-23.9	*- 11*	12.2
-72.2	*- 98*	-144.4	-47.8	*- 54*	- 65.2	-23.3	*- 10*	14.0
-71.7	*- 97*	-142.6	-47.2	*- 53*	- 63.4	-22.8	*- 9*	15.8
-71.1	*- 96*	-140.8	-46.7	*- 52*	- 61.6	-22.2	*- 8*	17.6
-70.6	*- 95*	-139.0	-46.1	*- 51*	- 59.8	-21.7	*- 7*	19.4
-70.0	*- 94*	-137.2	-45.6	*- 50*	- 58.0	-21.1	*- 6*	21.2
-69.4	*- 93*	-135.4	-45.0	*- 49*	- 56.2	-20.6	*- 5*	23.0
-68.9	*- 92*	-133.6	-44.4	*- 48*	- 54.4	-20.0	*- 4*	24.8
-68.3	*- 91*	-131.8	-43.9	*- 47*	- 52.6	-19.4	*- 3*	26.6
-67.8	*- 90*	-130.0	-43.3	*- 46*	- 50.8	-18.9	*- 2*	28.4
-67.2	*- 89*	-128.2	-42.8	*- 45*	- 49.0	-18.3	*- 1*	30.2
-66.7	*- 88*	-126.4	-42.2	*- 44*	- 47.2	-17.8	*0*	32.0
-66.1	*- 87*	-124.6	-41.1	*- 43*	- 45.4	-17.2	*1*	33.8
-65.6	*- 86*	-122.8	-41.1	*- 42*	- 43.6	-16.7	*2*	35.6
-65.0	*- 85*	-121.0	-40.6	*- 41*	- 41.8	-16.1	*3*	37.4
-64.4	*- 84*	-119.2	-40.0	*- 40*	- 40.0	-15.6	*4*	39.2
-63.9	*- 83*	-117.4	-39.4	*- 39*	- 38.2	-15.0	*5*	41.0
-63.3	*- 82*	-115.6	-38.9	*- 38*	- 36.4	-14.4	*6*	42.8
-62.8	*- 81*	-113.8	-38.3	*- 37*	- 34.6	-13.9	*7*	44.6
-62.2	*- 80*	-112.0	-37.8	*- 36*	- 32.8	-13.3	*8*	46.4
-61.7	*- 79*	-110.2	-37.2	*- 35*	- 31.0	-12.8	*9*	48.2
-61.1	*- 78*	-108.4	-36.7	*- 34*	- 29.2	-12.2	*10*	50.0
-60.6	*- 77*	-106.6	-36.1	*- 33*	- 27.4	-11.7	*11*	51.8
-60.0	*- 76*	-104.8	-35.6	*- 32*	- 25.6	-11.1	*12*	53.6
-59.4	*- 75*	-103.0	-35.0	*- 31*	- 23.8	-10.6	*13*	55.4
-58.9	*- 74*	-101.2	-34.4	*- 30*	- 22.0	-10.0	*14*	57.2
-58.3	*- 73*	- 99.4	-33.9	*- 29*	- 20.2	- 9.4	*15*	59.0
-57.8	*72*	- 97.6	-33.3	*- 28*	- 18.4	- 8.9	*16*	60.8
-57.2	*- 71*	- 95.8	-32.8	*- 27*	- 16.6	- 8.3	*17*	62.6
-56.7	*- 70*	- 94.0	-32.2	*- 26*	- 14.8	- 7.8	*18*	64.4
-56.1	*- 69*	- 92.2	-31.7	*- 25*	- 13.0	- 7.2	*19*	66.2
-55.6	*- 68*	- 90.4	-31.1	*- 24*	- 11.2	- 6.7	*20*	68.0
-55.0	*- 67*	- 88.6	-30.6	*- 23*	- 9.4	- 6.1	*21*	69.8
-54.4	*- 66*	- 86.8	-30.0	*- 22*	- 7.6	- 5.6	*22*	71.6
-53.9	*- 65*	- 85.0	-29.4	*- 21*	- 5.8	- 5.0	*23*	73.4
-53.3	*- 64*	- 83.2	-28.9	*- 20*	- 4.0	- 4.4	*24*	75.2
-52.8	*- 63*	- 81.4	-28.3	*- 19*	- 2.2	- 3.9	*35*	77.0
-52.2	*- 62*	- 79.6	-27.8	*- 18*	- 0.4	- 3.3	*26*	78.8
-51.7	*- 61*	- 77.8	-27.2	*- 17*	1.4	- 2.8	*27*	80.6
-51.1	*- 60*	- 76.0	-26.7	*- 16*	3.2	- 2.2	*28*	82.4
-50.6	*- 59*	- 74.2	-26.1	*- 15*	5.0	- 1.7	*29*	84.2
-50.0	*- 58*	- 72.4	-25.6	*- 14*	6.8	- 1.1	*30*	86.0
-49.4	*- 57*	- 70.6	-25.0	*- 13*	8.6	- .6	*31*	87.8

0	*32*	89.6	29.4	*85*	185.0	58.9	*138*	280.4
0.6	*33*	91.4	30.0	*86*	186.8	59.4	*139*	282.2
1.1	*34*	93.2	30.6	*87*	188.6	60.0	*140*	284.0
1.7	*35*	95.0	31.1	*88*	190.4	60.6	*141*	285.8
2.2	*36*	96.8	31.7	*89*	192.2	61.1	*142*	287.6
2.8	*37*	98.6	32.2	*90*	194.0	61.7	*143*	289.4
3.3	*38*	100.4	32.8	*91*	195.8	62.2	*144*	291.2
3.9	*39*	102.2	33.3	*92*	197.6	62.8	*145*	293.0
4.4	*40*	104.0	33.9	*93*	199.4	63.3	*146*	294.8
5.0	*41*	105.8	34.4	*94*	201.2	63.9	*147*	296.6
5.6	*42*	107.6	35.0	*95*	203.0	64.4	*148*	298.4
6.1	*43*	109.4	35.6	*96*	204.8	65.0	*149*	300.2
6.7	*44*	111.2	36.1	*97*	206.6	65.6	*150*	302.0
7.2	*45*	113.0	36.7	*98*	208.4	66.1	*151*	303.8
7.8	*46*	114.8	37.2	*99*	210.2	66.7	*152*	305.6
8.3	*47*	116.6	37.8	*100*	212.0	67.2	*153*	307.4
8.9	*48*	118.4	38.3	*101*	213.8	67.8	*154*	309.2
9.4	*49*	120.2	38.9	*102*	215.6	68.3	*155*	311.0
10.0	*50*	122.0	39.4	*103*	217.4	68.9	*156*	312.8
10.6	*51*	123.8	40.0	*104*	219.2	69.4	*157*	314.6
11.1	*52*	125.6	40.6	*105*	221.0	70.0	*158*	316.4
11.7	*53*	127.4	41.1	*106*	222.8	70.6	*159*	318.2
12.2	*54*	129.2	41.7	*107*	224.6	71.1	*160*	320.0
12.8	*55*	131.0	42.2	*108*	226.4	71.7	*161*	321.8
13.3	*56*	132.8	42.8	*109*	228.2	72.2	*162*	323.6
13.9	*57*	134.6	43.3	*110*	230.0	72.8	*163*	325.4
14.4	*58*	136.4	43.9	*111*	231.8	73.3	*164*	327.2
15.4	*59*	138.2	44.4	*112*	233.6	73.9	*165*	329.0
15.6	*60*	140.0	45.0	*113*	235.4	74.4	*166*	330.8
16.1	*61*	141.8	45.6	*114*	237.2	75.0	*167*	332.6
16.7	*62*	143.6	46.1	*115*	239.0	75.6	*168*	334.4
17.2	*63*	145.4	46.7	*116*	240.8	76.1	*169*	336.2
17.8	*64*	147.2	47.2	*117*	242.6	76.7	*170*	338.0
18.3	*65*	149.0	47.8	*118*	244.4	77.2	*171*	339.8
18.9	*66*	150.8	48.3	*119*	246.2	77.8	*172*	341.6
19.4	*67*	152.6	48.9	*120*	248.0	78.3	*173*	343.4
20.0	*68*	154.4	49.4	*121*	249.8	78.9	*174*	345.2
20.6	*69*	156.2	50.0	*122*	251.6	79.4	*175*	347.0
21.1	*70*	158.0	50.6	*123*	253.4	80.0	*176*	348.8
21.7	*71*	159.8	51.1	*124*	255.2	80.6	*177*	350.6
22.2	*72*	161.6	51.7	*125*	257.0	81.1	*178*	352.4
22.8	*73*	163.4	52.2	*126*	258.8	81.7	*179*	354.2
23.3	*74*	165.2	52.8	*127*	260.6	82.2	*180*	356.0
23.9	*75*	167.0	53.3	*128*	262.4	82.8	*181*	357.8
24.4	*76*	168.8	53.9	*129*	264.2	83.3	*182*	359.6
25.0	*77*	170.6	54.4	*130*	266.0	83.9	*183*	361.4
25.6	*78*	172.4	55.0	*131*	267.8	84.4	*184*	363.2
26.1	*79*	174.2	55.6	*132*	269.6	85.0	*185*	365.0
26.7	*80*	176.0	56.1	*133*	271.4	85.6	*186*	366.8
27.2	*81*	177.8	56.7	*134*	273.2	86.1	*187*	368.6
27.8	*82*	179.6	57.2	*135*	275.0	86.7	*188*	370.4
28.3	*83*	181.4	57.8	*136*	276.8	87.2	*189*	372.2
28.9	*84*	183.2	58.3	*137*	278.6	87.8	*190*	374.0

88.3	*191*	375.8	90.0	*194*	381.2	91.7	*197*	386.6
88.9	*192*	377.6	90.6	*195*	383.0	92.2	*198*	388.4
89.4	*193*	379.4	91.1	*196*	384.8	92.8	*199*	390.2
						93.3	*200*	392.0

SPECIFICATIONS

QQ-A-318b
"Aluminum alloy 52S; plate and sheet"

QQ-A-359c
"Aluminum alloy 3S; plate and sheet"

QQ-A-561b
"Aluminum alloy 2S; plate and sheet"

QQ-C-576(1)
"Copper plates, sawed bars, sheets, and strips"

BIBLIOGRAPHY

Bogle, A.G.
"The Effective Inductance and Resistance of Screened Coils"
Journal Institute Electrical Engineers, September, 1940

Howe, G.W.O.
"Q Factor of Single Layer Coils"
Wireless Engineer, June, 1949

Pender and McIlwain
Electrical Engineers' Handbook, Fourth Edition
John Wiley & Sons, Inc., New York, 1950

Sturley, K.R.
Radio Receiver Design, Part I
Chapman & Hall Ltd., London, England, 1951

Terman, Frederick E.
Radio Engineering, Third Edition
McGraw-Hill Book Company, Inc., New York, 1947

Welsby, V.G.
The Theory and Design of Inductance Coils
Macdonald and Company, London, England, 1950

Section 3

MAGNETIC MATERIALS

INTRODUCTION

The use of solid iron as a core for an electromagnet was utilized as far back as the time of Michael Faraday. The inefficiency of a solid-iron core for alternating-current applications was quickly recognized because of the excessive amount of heat generated within the core. The electrical loss producing this heat was found to be due to eddy currents induced within the iron. These losses were reduced by substituting iron wire or flat laminations which reduced the path of the circulating currents.

As usage developed in the higher-frequency range, it was discovered that smaller and smaller laminations were necessary. As far back as the late eighteen hundreds, iron filings imbedded in wax or shellac were used for high-frequency applications. This eventually led to the realization that finely-divided iron, treated to insulate each particle from the other, could be bound together by the addition of a binder, molded into the desired shape and heat treated to harden the binder, thereby producing a low-loss high-frequency core.

It was not until about 1930 that high-frequency powdered iron cores manufactured by mass production methods appeared. W.J.Polydoroff and Hans Vogt were early pioneers in this work.

Prior to World War II, iron cores were used in many high-Q antenna coils, especially in automobile radios and in permeability tuners in place of gang capacitors. Permeability tuned i-f transformers made their appearance but were expensive and, therefore, not popular.

Thread-grinding equipment for mass production, developed during World War II, made possible the inexpensive permeability-tuned i-f transformer as we know it today. Relatively few capacitor-tuned units are manufactured now.

Increasing demands for smaller coils for use in miniaturized equipments forced designers to look for other magnetic materials which would permit size reduction without sacrifice in the quality of performance. One such class of materials, ceramic in nature and called *ferrites*, was introduced as far back as 1909, but did not receive much attention until a more extensive investigation of this material was made by Philips Gloeilampenfabricken of Eindhoven, Holland in 1933. During World War II a considerable amount of further research was conducted and in 1947 J.S. Snoek published his well known book, "New Developments in Ferromagnetic Materials" (Elsevier, N.Y.), covering the work of that period.

After the War, a number of industrial concerns in this country as well as the military departments initiated ferrite development programs aimed at exploiting this very promising material. At this writing almost every television receiver and many radio receivers in domestic and military usage utilize this material in one way or another.

ELECTRICAL PROPERTIES OF MAGNETIC MATERIALS

Let us first consider the justification — other than possible economic reasons — for the use of a magnetic core and what it can do for the coil designer. Why not design all inductors around air cores which would have lower losses?

A magnetic core basically performs one or all of these functions:

(a) Miniaturization
(b) Inductance Variation
(c) Shielding

Miniaturization, as it is most frequently recognized and applied, involves the reduction in the physical size (and often weight) of an inductor without degrading its electrical performance. If the original inductors were designed around magnetic cores, they can frequently be further reduced in size by utilizing a more efficient core, a novel core design, or by utilizing a material having more favorable characteristics. Typical examples of this would be cup-core designs to replace simple windings having cylindrical cores; even if the same magnetic material were used, this change would permit using a smaller number of turns because of the more efficient use of the magnetic material, and would therefore lead to a physically smaller assembly

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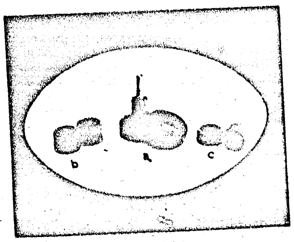
with the same or better Q characteristics. A somewhat different approach yielding a similar reduction in size would be to use a ferrite core in place of a powdered iron core without resorting to form factor modifications.

A more subtle aspect of miniaturization is one involving the improvement of electrical performance, generally by obtaining a higher Q, without increasing the physical size of the inductor. This, too, can often be accomplished in the manner just outlined for reducing size.

Inductance variation is another important function that can be readily accomplished using magnetic cores. This function, referred to as permeability tuning, is based on changing the reluctance or flux distribution in the magnetic circuit of the coil by physical displacement of the magnetic core. The simplest illustration of this is in the use of a movable cylindrical core in a solenoid winding. Basically, the same concept is used in the more complex magnetic core structures (i.e. the almost completely closed magnetic circuit in cup-core assemblies Fig. 3-1) where some portion of the magnetic core is made physically adjustable.



(a) 3 piece assembly (adjustable center core)
(b) 1 piece cup core (plain or with external threads)
(c) 2 piece cup assembly (non-adjustable)

Fig. 3-1 Typical cup cores and cup core assemblies

Permeability tuning is used almost exclusively in i-f transformers and in the tuners for the broadcast band in automobile radios. Oscillator coils, peaking inductors, filter reactors, and numerous other coils requiring adjustment after assembly into an electronic circuit use this simple and effective means of varying the coil inductance.

Non-magnetic core tuning, though somewhat foreign to the subject matter of this section, nevertheless, should be mentioned in discussing inductance variation since it is an effective method for coil adjustment in certain cases. This technique uses a disc or core of silver, brass, copper, or aluminum in the magnetic field of the coil. The non-magnetic core reduces the inductance in proportion to the magnetic-flux lines that it intercepts, so that physical movement of such a core in the magnetic field will cause inductance variation. The losses introduced by the core can be kept small by limiting the range of inductance adjustment. Examples of this type of tuning are high-frequency i-f transformers having silver plated brass tuning cores and television tuners having brass or aluminum threaded cores inside spacewound r-f and oscillator coils.

This type of tuning is of significance at higher frequencies since it reduces instead of increases inductance as does an iron core. Very-high-frequency windings, in general, have only a few turns, and iron cores for adjusting purposes only make such a winding more difficult to manufacture, whereas the non-magnetic core requires extra turns to make up for the loss of inductance due to the core, which is an advantage in many cases.

Magnetic shielding is the third function of powdered iron or ferrite cores. Such shielding confines the field of a high frequency inductor thereby permitting other circuit components to be placed nearer without deleterious effects and interaction from conflicting magnetic fields.

BASIC PARAMETERS

(a) Permeability
(b) Q
(c) Dielectric Constant

Permeability is defined as the ratio of the magnetic induction to the magnetic intensity and is represented by μ. Mathematically it is

$$\mu = \frac{B}{H}$$

where B is the induction in gausses and H is the field strength in oersteds.

The initial (or true) permeability is determined by the slope of the normal induction curve at zero

magnetizing field. This characteristic is most frequently determined by measuring the inductance of a coil wound on a toroidal core of the magnetic material and comparing this value with the inductance of a similar toroidal coil having an air core.

Of greater interest to the coil designer is the *effective permeability* (μ_{eff}) which is usually defined as the ratio of the inductance of a given coil with and without the core. This is an important working parameter to the designer since it reflects the composite effect of the true permeability of the core and the geometry of the specific coil and core combination including distributed capacity effects.

A practical method of measuring μ_{eff} requires the use of a Q-Meter. A suitable test coil is resonated (without the core) at a frequency which is within the range of the Q-Meter capacitor, say 100 $\mu\mu f$. The core is inserted into the test coil and the Q-Meter again resonated by changing the frequency without changing the capacity value used for the previous reading. Call the first reading f_1 and the second reading f_2. The effective permeability can be calculated from the following formula

$$\mu_{eff} = \left(\frac{f_1}{f_2}\right)^2$$

A practical example:

Coil without core:	$C_1 = 100\ \mu\mu f$	$f_1 = 1000$ kc
Coil with core:	$C_2 = 100\ \mu\mu f$	$f_2 = 500$ kc

$$\mu_{eff} = \left[\frac{1000}{500}\right]^2 = (2)^2 = 4$$

An alternate Q-Meter method utilizes a constant frequency and varies the capacity for resonance. If this method is used a value of C_1 must be chosen sufficiently high so that C_2 will be within range of the Q-Meter resonating capacitor when the core is inserted into the coil. This limits the range of effective permeability that can be measured to approximately 10, which is relatively low when ferrites are considered. The following formula applies:

$$\mu_{eff} = \frac{C_1}{C_2}$$

where C_1 = capacity for resonance without core and C_2 capacity for resonance with core.

A practical example:

Coil without core:	$f_1 = 500$ kc	$C_1 = 400\ \mu\mu f$
Coil with core:	$f_2 = 500$ kc	$C_2 = 100\ \mu\mu f$

$$\mu_{eff} = \frac{400}{100} = 4$$

Effective permeability may vary from slightly over one for high-frequency iron-oxide type cylindrical cores to as much as several hundred for certain types of ferrites made into closed cup-core designs having small air gaps. More specific examples are μ_{eff} of 1.5 to 3 for a 3/8" dia x 3/4" long cylindrical core in a universal winding and typical cores used for broadcast band tuning (535 to 1650 kc — .200" dia x 1 1/4" long) having an effective permeability of 9.5 or greater (see Figs. 3-3a-b-c and 3-4c).

The permeability of a powdered-iron core is determined by the basic powder, the method of insulating and binding the particles together and the pressure used in forming the core. It can, therefore, vary within a range of several percent when similar cores are made by different fabricators. Cores made by the same fabricator may vary in permeability because multicavity presses do not always have identical tools in all stations. Uneven tool wear and variation of applied pressure also affect permeability. The coil designer must recognize these factors and allow for reasonable tolerances. In general, the closer the tolerance, the more expensive the core. Generally accepted tolerances for permeability are ± 2% or ± 4%. Cores having closer tolerances generally have to be selected which results in a percentage of unusable cores on either side of the nominal value.

It is suggested that the designer familiarize himself with "Tentative Electronic Iron Core Preferred Dimensional Specification" No.11-55-T* which covers in detail preferred mechanical and certain electrical tolerances as adopted by the electronic core manufacturers.

Because of the newness of the art, there has been no similar Standard set up for ferrite cores. In general, ferrite materials being ceramic in nature follow the usual accepted mechanical tolerances for electrical-grade ceramics. Because of wider mechanical tolerances for ferrite than for iron, it is necessary to allow broader tolerance for permeability. Common tolerances are ± 10% with ± 5% or ± 3% generally held only by selection with a resulting higher percentage of rejects.

Q is a term loosely used to designate the factor of merit of a magnetic core. Actually this is a non-existent term since Q is in reality the factor of merit of an inductor (with or without a core) and is defined as the ratio of the reactance to the equiva-

* Metal Powder Association, 420 Lexington Avenue, New York 17, New York.

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lent series resistance.

$$Q = \frac{2\pi fL}{R} = \frac{\omega L}{R}$$

Since there is no direct way to determine a core *factor of merit* without an associated coil, it is generally accepted practice to refer to the Q of a core as if it was an inherent parameter. It should be realized that the *core and coil* Q will rarely ever be the same for any two test coils and that core Q is, therefore, more of an effect than a characteristic. Suitable test coils are in general those comparable to a working type of winding. The Q obtained is a relative value for comparison with other cores of similar form factor but not necessarily of the same material.

The Q, then, is obtained by inserting the core in a suitable test winding and resonating the coil on the Q meter at the desired frequency (see page 3–9 for proper choice of a test coil).

The Q is largely a function of the type of iron powder used and method of processing. It does not vary greatly from core to core due to molding pressure. Typical tolerances for Q are ± 10%, ± 7 1/2% and ± 5%. Tolerances closer than this would normally be obtained by a 100% test and selection and might result in a number of rejects thereby increasing the core cost.

Ferrite materials are the result of chemical reactions and many factors in their production (such as firing temperature and atmosphere) affect the Q. It is difficult at the present state of the art to manufacture them to close Q tolerance. Larger tolerance on Q is, therefore, necessary than is generally expected of similar powdered-iron cores.

Effect of Inserts:

Iron cores with screw inserts are frequently used in r-f coil design. The effect of introducing the screw is, of course, to reduce the effective Q and if the screw is grounded through the mounting device and shield assembly, the capacity from winding to ground also is often increased.

The conditions under which cores are used greatly influence the reduction of Q due to a molded-in screw, but in general it can run as high as 35% for steel screws at 1000 kc and 20% at 15 Mc, to as low as 3% for brass screws at 1000 kc and 2% at 15 Mc. Since steel screws generally reduce the effective Q four or five times as much as do brass screws, their use should be avoided whenever possible.

This reduction of Q and added capacity effect can be minimized to a great extent by using cores which have screws insulated from the magnetic material. This is accomplished by molding the screw into a phenolic bushing which is attached to the end of the iron core. The length of the bushing governs the proximity of the metal screw to the iron core and therefore affects the Q. This type of core is more expensive than one having the screw molded directly in the iron and should be avoided unless required for electrical reasons.

Ferrite cores, by their very nature, cannot have screws molded in during manufacture, but must have them cemented into a cavity in the end of the core subsequent to the final firing operation. This in itself tends to discourage the use of screws in ferrite bodies.

The dielectric constant of powdered iron and ferrite material has been given little attention in the literature. Manufacturers have not been able to produce iron cores having widely different dielectric constants. Even though isolated projects have possibly indicated a need for iron cores of lower dielectric constant, none are known to be offered commercially.

The dielectric constant of typical commercial Carbonyl E iron cores varies from as low as 20 to as high as 90, or even more. This, of course, is high compared to coil form materials such as phenolics, paper and ceramics which have dielectric constants between 2 and 7. Ferrites, because of many different chemical compositions, have a wide range of dielectric constant. Measurements taken on typical ferrite materials indicate the dielectric constant to be as low as 20 and as high as 460. The average value used in commercial production of r-f and i-f coils and transformers is probably between these extremes. It is known that some commercially produced ferrites have dielectric constants as high as 10,000 and dielectric constant values as high as 100,000 have been reported, but there is no information available regarding the other characteristics of these special compositions.

To the average coil designer, the dielectric constant is not of too much importance if it is relatively low as in the case of iron materials. It is difficult to reduce the dielectric constant of iron compositions. Ferrites, because of the almost limitless body compositions offer more latitude for change, if, due to a particular design, a core having a high dielectric constant should cause undesirable capacity coupling between windings or circuits or increase the distributed capacity of the coil to an unreasonable value. Because each different body composition usually requires its own tools to allow for shrinkage from the initial pressing to the final firing, it is recommended that the coil designer work with established commercial compositions for which tools are available.

The Curie temperature is defined as that temperature at which a magnetic material ceases to have magnetic properties. Because, in the case of ferrites, this may be in the useful working range it is an important characteristic that can be a limitation to their usage.

Ferrite compositions differ in temperature characteristic. Some have Curie temperatures as low as 250F and others greater than 400F. In general, bodies having high permeability and low Q have low Curie temperature and bodies having low permeability and high Q have high Curie temperature. Compositions intended for use at higher frequencies generally are of the higher Q type and tend to have higher Curie temperature.

The curves of Fig. 3-2 illustrate the characteristic of permeability vs. temperature. Some materials show little change until near the Curie temperature (curve a) while others increase appreciably in permeability over the entire temperature range (curve b) until the Curie temperature is reached. As an example of the useful range, materials having a temperature characteristic resembling curve "a" could be used up to 300 or 325F while a material illustrated by curve "c" would be useful only to 250F. A material such as "b" is of little value at any temperature unless some form of compensation is provided for a coil assembly using this composition.

Observations have been made of bodies having a negative temperature characteristic although they are not believed to be produced commercially at the present time.

DESIGN FACTORS

In order to utilize a magnetic material to the

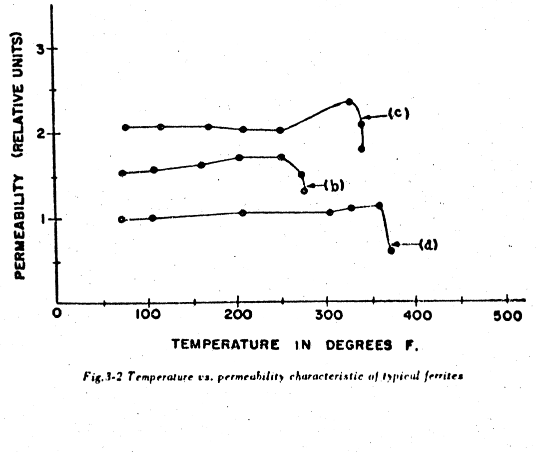


Fig.3-2 Temperature vs. permeability characteristic of typical ferrites

best advantage, certain principles of design must be clearly understood. Basically, a given type of powdered iron or ferrite material has an established true permeability. How nearly this can be realized in actual practice depends upon the core and coil design.

The true permeability is determined by the closed-core (toroid) method. The effective permeability of any particular design of core more nearly approaches the true permeability as the core configuration approaches the toroidal, or closed core configuration. Concentrating the winding near the core tends to reduce the stray or leakage flux thereby more nearly approaching the ideal condition for maximum permeability. See curves of Fig.3-3 showing effective permeability of windings having various form factors all using the same powdered iron core.

The form factor (length to diameter ratio) of the core itself is of extreme importance in realizing the maximum effective permeability. The curves of Fig. 3-4 illustrate the increasing effective permeability as the length to diameter ratio increases. Permeability tuners designed to cover the broadcast band require a length to diameter ratio in the order of 6 to 1 when using common types of powdered iron. Ferrite cores having higher true permeability will also have a higher effective permeability than will iron for the same length to diameter ratio thereby permitting tuners to be designed with shorter core travel if ferrite cores are used.

The proximity of the winding to the core is also an important factor. The curves of Fig. 3-5 show the decreasing effective permeability as the ratio of mean turn diameter to core diameter is increased.

The effect of approaching the toroidal or closed core configuration is shown by Fig. 3-6 wherein cylindrial cores are compared to open and closed cup cores for effective permeability.

Tolerances:

The coil designer should always be cognizant of practical mechanical tolerances on all elements. This is especially true of magnetic cores. Modern commercial manufacturing methods result in well established tolerances since little, if any, machining other than external thread grinding is performed subsequent to pressing. This means that the core as pressed has no further sizing operations

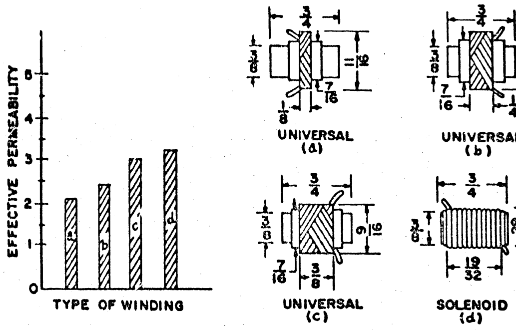


Fig.3-3 Typical coil form factors with same core to illustrate variation of effective permeability.

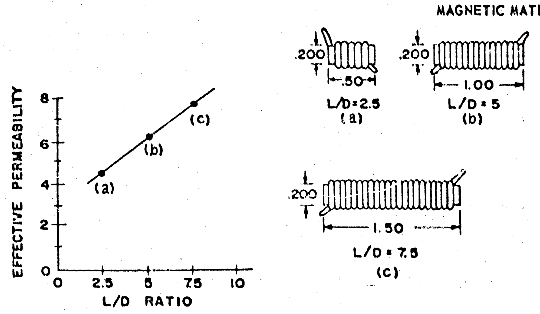


Fig.3-4 Variation of effective permeability with length to diameter ratio.

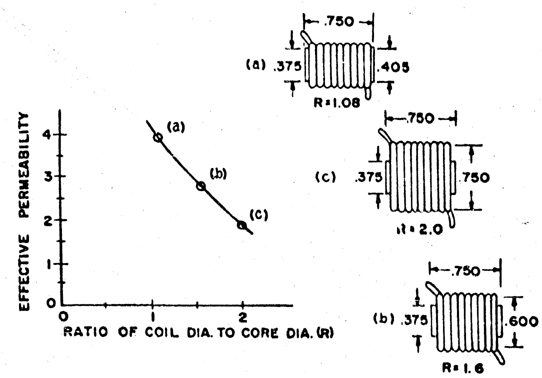
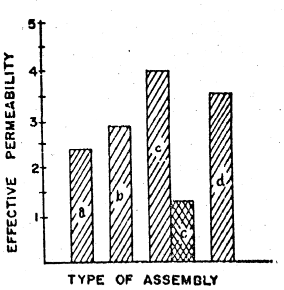


Fig.3-5 Variation of effective permeability with varying ratio of coil to core diameter.



(a) cylindrical core (see Fig.3-3b)
(b) one piece cup (see Fig.3-1b)
(c) three piece assembly (see Fig.3-1a)
adj core in *adj core out*
(d) two piece assembly (see Fig.3-1c)

Fig.3-6 Permeability for typical assemblies.

with which to correct dimensional errors. The following tolerances have become fairly well established for iron cores:

External Diameter:

(Because of die wear cores tend to increase in diameter so tolerance is always stated on positive side), i.e. .195" − .000" + .005" (.005" is generally accepted regardless of diameter).

Length:

(This is governed by pressing on all but side pressed -- and requires a larger tolerance):

up to ½" long + .010"
over ½" to 1" long ±.015"
over 1" to 1½" long ±.020"
over 1½" to 2½" long ±.030"

Internal Diameter:

Same as for external diameter except as a negative tolerance i.e. .140" + .000 −.005".

It is suggested that the Metal Powder Association "Tentative Electronic Iron Core Preferred Dimensional Specification" No.11-55T be consulted for the latest information on above tolerances. This specification will be kept up-to-date whereas it is beyond the scope of this manual to predict changes that may take place in normal commercial practice.

Ferrite cores being of a ceramic nature follow a slightly different tolerance pattern. In general, the tools are subject to wear as are iron-core tools. The tolerance is generally stated as plus or minus since wider tolerance is required to allow for the shrinkage involved between pressing and final firing. Accepted commercial practice is as follows:

External Diameter	±.005" or ±1% which ever is larger
Length	±.010" or ±2% which ever is larger
Internal Diameter	±.005" or ±1% which ever is larger

Since industry standards for ferrite cores have not yet been coordinated by an agency representing the majority of producers, it is suggested that the designer consult with the manufacturer of his choice.

Practical shapes are naturally those that fulfill the design requirements and are economical to produce. Since most cores, either iron or ferrite, are pressed, it is in the interest of economy to utilize readily pressed parts. Round external and internal shapes are easiest to tool and maintain. Blind holes should have taper, the closed end being smaller of diameter than the open end. The length, in the pressing direction, should be as short as practical and the length to diameter ratio should be kept as small as possible. Wall sections should be as thick as possible and preferably not thinner than 3/64 inch.

Hollow cylindrical cores of either iron or ferrite sometimes used as external coil shields inside of metal cans are frequently extruded. It should be remembered that extruded parts can only be made with a uniform cross section. There can be no tapers, offsets or blind holes. This method of production is relatively inexpensive for those types of cores that are practical for extrusion.

TESTING OF IRON AND FERRITE CORES

Quantity production of any component can be expected to be no more uniform than the materials and parts that go into its construction. This is especially true of inductors and transformers having high-frequency magnetic cores. It is, therefore, important to the coil design engineer to be certain that the specifications prepared for the magnetic cores adequately describe the mechanical and electrical parameters and that unnecessary tests or tolerances that serve no useful purpose are not also included.

Mechanical:

Mechanical and physical characteristics are the easiest to evaluate and should be checked first. If the mechanical dimensions of the magnetic core are not within the required tolerances the electrical characteristics are of little importance.

Outside diameter, length and other easily accessible dimensions are most readily measured with a micrometer. Inside diameters, blind hole depths and similar dimensions are best inspected with plug gages. Eccentricity of molded-in screws is measured by chucking the core in a lathe or other suitable fixture and measuring the screw run out with an indicator gage (see Metal Powder Association Tentative Electronic Iron Core Preferred Dimensional Specifications No. 11-55T). In actual practice just the opposite is done, since the core is used by holding the screw and revolving the core within a coil form. Measurement of core eccentricity by chucking the screw introduces errors that are difficult to reconcile between supplier and user. For this reason the former method has been adopted by the majority of the industry.

Electrical:

The electrical parameters require specialized test equipment. A Q-Meter and a Megohmmeter are the most useful. Most production and laboratory measurements of permeability and Q are made with the Q-Meter. Absolute measurements are rarely used but more often comparative measurements are made to a previously established *standard or reference core* (Selection of standards is described on page 3-10). Permeability and Q tolerances are stated as ± deviations in per cent from the established standard.

Permeability and Q:

In general a test procedure which approaches as nearly as possible the conditions under which a core will function has been proven to be the most satisfactory. This is not always possible or practical. The practical approach, then, is to use a test coil which will best show up the most important parameters. That is, if permeability is the most important characteristic for a given application then the test coil should be constructed to best differentiate between small differences in permeability. The same applies to Q or any other important characteristic.

Example: A coil of approximately the same length as the core wound on a thin wall tube produces the highest effective permeability. Used as a test coil, this winding can be expected to best differentiate between small differences in permeability.

Cores used for wide range tuning (to cover broadcast band), having large length to diameter ratio, frequently have a satisfactory overall (or total) permeability but do not have a proper distribution of permeability throughout the length of the core. Such cores are said to be non-homogeneous. The maximum and minimum frequency coverage will be correct but they will not track one with another. It is necessary to classify such cores into groups having similar permeability distribution if they are to operate in cascaded or associated circuits. The simplest way to accomplish this is to provide stops in the bottom of the test coil used for the overall permeability test so that cores can be matched to a reference core at intermediate points say 1/4, 1/2, and 3/4 insertion.

A somewhat more complicated but also more satisfactory type of test is to employ two identical test windings. One winding has the reference core and the other has the core to be tested. Each winding is in the tuned circuit of an oscillator, the outputs of which are combined to produce a beat frequency. As the cores are simultaneously moved into their respective coils an audible note is produced if the frequencies of the oscillators and, there-

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fore, the permeabilities of the cores are not alike. By use of a suitable audio frequency meter the setup can be calibrated to indicate the extent to which the permeability of the core under test deviates from standard and the direction of the deviation.

Sleeve cores such as those used for shielding purposes are generally tested for permeability and Q by inserting them inside of a test coil wound on a tube having an inside diameter which will accept the outside diameter of the sleeve. The winding length is generally about 95% of the core length and the core is centered in this winding.

Cup cores present a more complex test problem. There are a number of types, and each must be treated in a different manner. The more common are shown in Fig. 3-1.

Type A, the three piece assembly, is treated as individual parts to avoid having to keep them as units during subsequent assembly. The center adjusting core having the screw insert is tested as any cylindrical core previously described. The two halves of the cup proper must be tested as individual pieces since there is no means of conveniently handling them in pairs during normal manufacturing operations. The usual procedure is to prepare a test coil having the center core and one cup (bottom) with the winding on the center core. The standard cup is used to complete the assembly as a top core. After the setup is made the standard cup is removed and successively replaced by the cups to be tested.

Type B cups are tested by inserting the central pin into a winding similar to that used in a typical production assembly. The C cups are similar in construction except generally of shorter length and are commonly used in pairs. This means that the outer shell and the center pin must be of identical length to avoid air gaps either around the shell or at the center pin. A jig using one cup as the lower element with a winding on the center pin is the most convenient. The cores to be tested are placed on top to complete the closed cup assembly.

New shapes of cores will present slightly different test-coil problems which can be solved by the application of the above principles and ideas.

After chosing a suitable test coil and a standard core the test method using a Q-Meter must be established. It was previously suggested that a test procedure approximating operating conditions is generally most satisfactory. This includes the test frequency. It is common practice to resonate the test coil with 100$\mu\mu f$ so that deviations in capacity from the nominal value will represent percent of tuning capacity which can be interpreted as percent of effective permeability. It is, of course, necessary when designing the test winding to make the inductance of the proper value to resonate with 100$\mu\mu f$ at the desired frequency.

The test procedure is as follows: The standard core is inserted in the test coil which is connected to the coil terminals of the Q-Meter and the main capacitor dial is set at 100$\mu\mu f$. The vernier dial is set at zero. The frequency is then adjusted for maximum Q and the value noted. The standard core is removed and the cores to be tested inserted into the test coil and resonated with the vernier dial (without changing the frequency dial or the main capacitor dial). The vernier will accomodate a variation of ± 3% in effective permeability when the main tuning capacitor is set at 100$\mu\mu f$. Larger deviations from standard must be observed by leaving the vernier dial at zero and resonating with the main capacitor dial noting the deviation from the original setting of 100$\mu\mu f$. The percentage can be computed. (Note that an increase in the value of capacity indicates a decrease in effective permeability). Q can be read directly on the meter and compared to the Q of the standard core.

Resistance:

The resistivity of a magnetic-core material, especially powdered iron, is also regarded as an important characteristic. Normally, the resistivity could be determined from simple measurement of the resistance between the parallel faces of a cube of the material, but this method is not practical when dealing with commonly used shapes such as cups and cylindrical cores. A more practical test, frequently used for production test purposes, is to contact points on the core surface arbitrarily 1/4" apart, or under certain circumstances even on opposite sides of a core, and measure the resistance with a suitable megohmmeter. It is not uncommon for high resistance cores to measure between 5,000 and 50,000 megohms between such test points. Irregular shaped cores may have high density sections caused by conditions of manufacture. The resistance of these areas should not be permitted to be unreasonably low. If a core is broken, the resistance should be substantially the same for all sections.

SELECTION OF STANDARDS

After the test coil is chosen it is necessary to have a *standard core* to use as a point of reference. In most cases this standard core will be furnished by the core manufacturer and will have been selected as having average electrical characteristics. Pro-

duction parts can be expected to fall, within the prescribed tolerances, on either side to this standard.

If a standard core is not available from the core manufacturer it can be selected from a representative group, preferably a reasonably large production run, by the following method: An 8 x 10 cardboard is ruled with lines about 1" apart running the width of the sheet. The center line is marked zero and lines to the right are marked +1%, +2% etc. and lines to left are marked −1%, −2% etc.

A core is chosen arbitrarily and labeled *temporary standard* (for permeability) and the Q-Meter set up using this core. It is then placed on the *zero line* of the cardboard. 25 or 30 cores are picked at random and measured and then placed on the cardboard in the place representing their permeability with respect to the temporary standard. After this is completed it will generally be found that the majority are in a group which may or may not be centered around the original temporary standard. The core nearest the imaginary center of this group is now selected as the final standard for permeability, providing the Q is about average. If the Q varies considerably from core to core a suitable *Q standard* can be then selected from a group of near nominal permeability cores by the same method just described for permeability. The standard selected should be appropriately tagged and several duplicates selected for future use.

PREPARATION OF A PURCHASE SPECIFICATION

In order to insure that the standard selected is duplicated by any manufacturer who may be called upon to produce the core, a specification adequately describing the part must be prepared. Many core specifications have been issued which are almost meaningless when critically examined. A complete specification should include the following:

(1) A drawing showing all dimensions and tolerances including color coding or other marking.
(2) Electrical specifications to include:
   (a) Permeability tolerance (Permeability to be compared to approved standard)
   (b) Q tolerance (Q to be compared to approved standard)
   (c) Resistance (if required) and how measured.
   (d) Test frequency or frequencies, especially to be used for Q. This should cover the operating frequency range.
   (e) Complete drawing and specification of the test coil unless supplied along with the standard core.
(3) Miscellaneous
   (a) Rust proofing treatment, surface etching, or other special requirements.
   (b) Physical strength requirements and how measured.
   (c) Precautions regarding resistance to particular solvents or coil waxes.
   (d) Lubrication if required (on threaded cores)
   (e) Other specifications as may be required for special application.

It should be noted that unnecessary specifications should not be included just because the above outline mentions a specific item, i.e., if a given application does not require the core to be subjected to high humidity and if samples submitted by the vendor are suitable, there is little justification to give detailed specifications on rust proofing treatment.

It is recommended that r-f core manufacturers or their catalogs be consulted before preparing final core specifications so that mechanical dimensions and electrical standards will conform to items currently in production or for which production tooling is available. Special shapes and special dimensions will require new tooling or special machining which can be expensive operations with the consumer bearing the cost.

TYPES OF IRON POWDERS AND THEIR USAGE

The trial and error method was long used for selecting core materials. As more and more was learned about the behavior of iron-dust cores, and as iron powders were improved, the art of powdered metal cores became established on a more scientific basis.

The modern engineer need no longer depend upon hit and miss methods but can choose a core material based upon proven knowledge of its performance characteristics. The majority of the iron powders used in electronic applications today can be grouped into the following four general types:*

1. Reduced
2. Electrolytic
3. Oxide
4. Carbonyl

*Refer to Fig.3-7 for the general frequency characteristics of these types of iron powders, also to Fig.3-10 for additional recommendations of high Q at various frequencies.

Reduced Iron:

This type is produced from iron oxide such as mill scale and chemically reduced in an atmosphere of hydrogen or other suitable gas. The final product which is a relatively pure iron is pulverized by grinding or ball milling and classified as to particle size either by screening or air-classification.

This also includes what is commonly known as *sponge iron* which uses ore as the basic material for reduction. This iron is generally recommended for use under one megacycle.

Electrolytic Iron:

This type is produced from plate iron which is first electroplated on a suitable cathode. The plated iron is then stripped from the cathode and pulverized in a manner similar to that described above for reduced iron. Electrolytic iron is relatively pure and has high permeability. The frequency range of a finished core is somewhat dependent upon the particle size and how well the individual particles are insulated. In general, this type of iron powder is best suited for application under 2 megacycles.

Oxide:

Oxide (Fe_3O_4) either natural (commonly known as magnetite) or synthetic is frequently used for r-f cores. The natural oxide is pulverized iron ore and is generally of relatively large particle size. It is relatively inexpensive and can be used for quite a wide frequency range. It has lower permeability than most other iron powders and is therefore not suitable for wide-range tuning purposes.

The synthetic oxides are extremely fine and have relatively high Q at higher frequencies. Some of these oxides are suitable up to 200 megacycles or higher.

Carbonyl Powder:

Probably the most widely known powder is the carbonyl group. There are several types, their basic difference being particle size and particle hardness. Each has a definite usage and since at the present time there are at least ten types it is suggested that information regarding frequency limitations be obtained from the manufacturer* of the powder. Table (Fig. 3-7) shows values of μ_{eff}, Q and frequency range for several of the more common carbonyl irons.

Briefly, these powders are prepared from iron-pentacarbonyl $F_e(CO)_5$, a yellowish liquid with a boiling point of 101.5 C. The iron pentacarbonyl is decomposed at a high temperature and the starting materials, iron and carbon monoxide are re-formed. The resulting iron particles are of spherical shape and vary in diameter from 3 to 20 microns. The useful frequency range varies from 50 or 100 kilocycles to 100 or 200 megacycles. The smaller particle-sizes are, of course, used at the higher frequencies.

The spherical nature of this material results in many desirable characteristics, among which are ease of insulating and pressing. It is also extremely uniform and can be depended upon to produce finished cores to reasonably exacting specifications.

METHOD OF MANUFACTURE OF MAGNETIC CORES:

It may seem that the coil design engineer has little concern with the problems involved in insulating and fabricating magnetic powders or with the manufacture of ferrites but a working knowledge of these operations can be extremely beneficial in choosing the most suitable iron or ferrite core not only from the standpoint of operation but from the uniformity to be expected.

Iron Cores:

The manufacture of iron cores essentially consists of the following operations:

a. Insulating the powder
b. Adding the binder (synthetic resin)
c. Granulating and classifying the agglomerates
d. Pressing or extruding
e. Polymerization of the resin binder
f. Final test

Any form of particle insulation tends to reduce the effective permeability of the finished core and to increase the Q by reducing eddy-current losses. If the insulating medium is of the surface-coating type, i.e., insulating varnish or resin, it takes up space that could otherwise be occupied by iron particles in a finished core. If the insulation is a chemical conversion of the particle surface, it reduces the pure iron content of the particle and the converted surface occupies space that could otherwise be occupied by active iron. The effect is the same in that the permeability is reduced by either method.

Resign Coating:

The simplest and most elementary method of insulating iron powder is to make use of the resin used for binding the particles together. This

* Antara Chemicals a Sales Division of General Aniline & Film Corp.

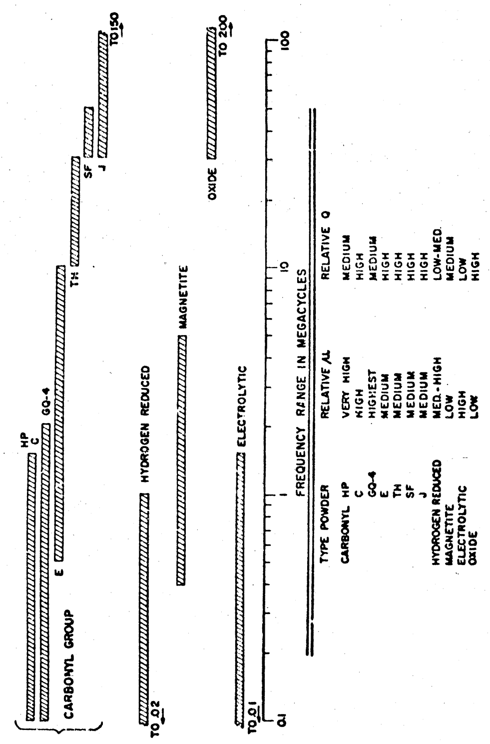

Figure 3-7 Characteristics of typical iron powders.

resin, generally a phenol-formaldehyde, is applied by wetting the iron particles with resin solution in a suitable solvent. This mixture of iron powder and resin is mixed until the solvent is evaporated and large agglomerates are formed.

Theoretically, of course, each particle is thereby coated with a thin layer of resin. In practice, it is impossible to effectively produce a high degree of insulation by this process and it is generally only used for cores intended for operation at lower frequencies (up to 1000 kc) where high insulation resistance is not of paramount importance. Surface resistance between points 1/4" apart may be as low as 1000 ohms or lower.

Sodium Silicate:

Another similar method uses sodium silicate which produces relatively high insulation resistance. Sodium silicate is hygroscopic in nature and when exposed to humidity, the insulation resistance rapidly decreases – a decided disadvantage for most applications. For this reason, this method has been largely superseded by chemical insulation.

Chemical Insulation:

The most effective insulating method is the so-called *chemical process*. Briefly, this uses any one of several materials (liquid), usually phosphoric acid solutions sometimes with the addition of other ingredients. The liquids react chemically with the iron, producing an iron phosphate coating on the surface of each particle. This coating has high insulation resistance and is reasonably tough.

This insulation provides no adhesive between particles and the usual phenolic resins or other media are used for binding the finished core together. Under normal room conditions of temperature and humidity (70 F and 50% relative humidity), it is not uncommon to measure resistance values between contact points on the finished core surface 1/4" apart of ten thousand megohms or greater. The insulation resistance stands up well under conditions of high humidity and can be used up to 200 megacycles or higher. A certain amount of rust proofing is also provided by this type of insulation.

Addition of Binder:

In the case of uninsulated cores the first step (or the second step for insulated cores) is the addition of the binder. This is usually a synthetic resin, commonly a phenol-formaldehyde. For processing convenience, most resins are introduced as a solution (dissolved in a solvent) and mixed with the iron powder in a suitable heavy-duty mixer. The resin content varies depending upon the intended application and the strength required. A reasonable average is between 2 and 4%. Sufficient solvent is used to enable thorough coating of the particle surfaces. Mixing is continued until the solvent is evaporated and the iron particles and resin form a putty-like mass.

In order to obtain uniformity by pressing in small cavities which are volumetrically filled, it is necessary that the putty-like mass be reduced to a fairly uniform particle size – about like granulated sugar or table salt. This is accomplished by hammer mills or other suitable equipment either while the material is slightly damp with solvent or after the solvent has been removed.

Subsequent to granulation, the powder is classified, generally by screening, and the oversize and undersize particles removed for further processing. The selected particles are now ready for forming into the required shape.

Pressing into final shape is accomplished either by single-cavity or multi-cavity presses either of the reciprocating or rotary type. Many presses are equipped to apply pressure at each end independent of the other end thereby allowing the maximum flexibility of adjustment to insure uniform permeability from end to end, and reducing die friction in critical areas so that intricate shapes can be readily fabricated.

The variety of shapes that can be pressed varies from solid and hollow cylindrical cores, either with or without screw inserts in one end, to the various cup and shell cores having center stems or holes for feed-through adjusting elements. Fig. 3-8 shows typical examples of end-pressed powdered iron cores.

Extruding is sometimes used to produce parts such as hollow cylinders having uniform cross section. The powder is insulated in the same manner as required for pressing but the binder is varied to give the proper plasticity to permit extrusion through suitable dies. In general, a slightly moist (with solvent) powder is required which after extrusion is air dried and cut to length.

After either pressing or extruding, it is necessary to *cure* or *polymerize* the resin in order to give the core mechanical strength. This is a time and temperature cycle. The time and temperature are dependent upon the type of resin used, the size and shape of the core, and the ultimate strength desired.

The curing temperature must not exceed that

which will damage the iron particles or resin electrically, and is generally in the range of 275 to 350 F. The time varies from a few minutes up to 24 hours or even more.

It should be recognized that a properly cured core is practically insoluble in common solvents such as alcohols, ketones, and hydrocarbons. This should be checked if cores are to be used in connection with coil lacquers or varnish impregnants.

After the core is completely finished, it should be tested for electrical and mechanical parameters as previously described under *Testing of Iron and Ferrite Cores.*

Nearly all types of powdered iron cores can be machined or ground providing reasonable care and proper equipment is employed. Rough shaping may be accomplished on standard machine shop equipment providing tungsten carbide tool bits are employed. Iron cores should never be machined to final size but allowance should be made for grinding to final dimensions. Wet grinding may be employed for very fine finishes. Care should be exercised when dry grinding cores since the extremely fine iron particles ignite very readily from the heat generated by grinding and will continue to glow and burn. These fires often start in ventilating ducts

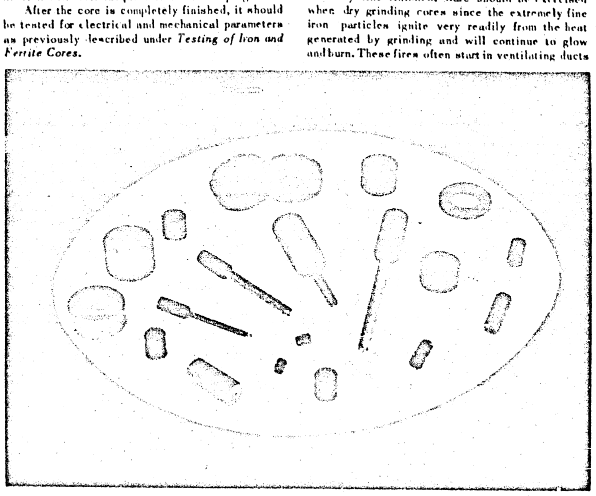

Fig. 3-8 Typical end-pressed cores

Machining Iron:

During the course of coil development, it is often necessary to try shapes that are not readily available or for which production tools have never been provided. It is desirable to fabricate such shapes from more available forms without resorting to expensive tool setups.

or in the chip pans of lathes or other machine shop equipment.

For grinding, high rotational speeds are recommended. The following abrasive wheels have been found to produce satisfactory results:

"Radiac" POR-OS-WAY...9A46-1-2VOS manu-

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factured by A.P.DeSanno & Son, Phoenixville, Pennsylvania.

"Carborundum...GC-120-111VR" manufactured by Carborundum Company, Niagara Falls, New York.

"Robertson...WA-521-J5V manufactured by Robertson Manufacturing Company, Trenton, New Jersey

Even in production quantities, certain machining operations can be performed economically. An example is the various shapes of cup and slug cores having external threads used for tuning i-f transformers. Grinding equipment available since World War II has made this type of core economically possible.

It should be remembered that a running thread is relatively easy to produce whereas it is practically impossible to grind threads on parts having more than one external diameter. Fig. 3-9 shows typical externally-ground threads.

Machining operations should be carefully controlled so that the surface particles are not damaged thereby short-circuiting one to another which reduces the effective Q of the core.

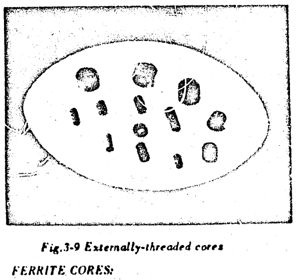

Fig.3-9 Externally-threaded cores

FERRITE CORES:

Contrasted to iron cores, ferrite cores are manufactured by entirely different processes. It was previously shown that iron was the principal ingredient of the iron core and the magnetic properties are inherent in this basic material. In contrast, ferrite materials are made from metallic oxides, the kind and proportions depending upon the required performance characteristics.

These oxides when properly processed form a crystalline compound having magnetic properties. The basic ingredient of all ferrites is iron oxide Fe_2O_3. This is combined with other bivalent metallic ions. Zinc is probably the next most common ingredient and following that nickel, manganese, copper and lithium, not necessarily in that order.

Because of the complex nature of these chemical compounds many characteristics can be obtained by proper choice of ingredients and method of processing. It is beyond the scope of this manual to go into detail on the manufacture of ferrites, but the basic steps in producing a typical ferrite will be outlined. The following operations are generally followed:

(a) Mixing
(b) Calcining
(c) Milling
(d) Adding the binder
(e) Granulation and classification
(f) Pressing or extruding
(g) Firing
(h) Final Test

The selected oxides are thoroughly mixed in a suitable mixer, such as a ball mill so that there will be an intimate contact between the various particles. This mixing may be either wet or dry. If wet mixing is used, it is necessary to remove the water from the resulting slurry either by evaporation or by the use of a filter press. After drying, the mixture is pulverized.

Following mixing, it is not uncommon to *calcine*. Calcining is heating the mixture in suitable containers at a temperature somewhat lower than the final firing temperature to initiate the chemical reaction which converts the oxides into a ferromagnetic material and reduces the amount of shrinkage by driving off the chemically combined water and volatile gasses.

Subsequent to calcining, it is necessary to reduce the resulting caked mass to powder. Following this operation, an organic binder is added and the powder agglomerated and classified into particles of suitable size for pressing.

Instead of granulating the powder, it may be extruded by an hydraulic extrusion press into rods or cylinders having almost any uniform cross section. These rods or cylinders are then cut to proper size before firing.

Pressed or extruded parts are fired in an electric or gas furnace at about 1300 C. The exact temperature and length of firing is dependent upon the material and the characteristics desired.

The material does not display the final ferromagnetic properties until the firing operation is complete. For this reason electrical control tests cannot be performed until the final stage of manufacture. At this point, it is too late to make corrections. Final tests are very similar to those described for iron cores, i.e., permeability, Q, and mechanical dimensions.

MACHINING FERRITES:

Because of the nature of ferrite material, cores can only be produced economically by machine pressing or extrusion using expensive tools. This is somewhat of a deterrent to development work requiring only a few samples of a given shape.

It is possible to accomplish certain operations, mainly grinding, by the use of diamond grinding or cutting wheels or by less expensive silicon-carbide wheels properly cooled by water or a soluble oil. It is advisable to start with a piece as near to the finished dimensions as possible and to grind threads or perform the fewest possible operations. Great care must be exercised especially with respect to the pressure applied by chucks or other holding devices. The material is extremely brittle and will crack if slightly stressed.

If the parts to be made are not complicated by thin sections, it is sometimes possible to perform grinding or other machining operations prior to final firing much more easily than they can be accomplished on completely fired parts. Allowance must be made for shrinkage in the final firing operation. Partially fired parts are not as brittle but neither do they have the compressive strength of finished parts.

CONCLUSION

The core technologies, the operational guides, and the design suggestions given in this section have been aimed more at providing practical orientation for the designer rather than at laying down step-by-step instructions for selecting specific cores for specific applications. The latter task, if at all possible, would require an exhaustive treatment of the innumerable variations available in commercial core compositions, and the endless combinations of electrical and physical requirements that could be demanded in practice (size, tuning range, inductance, Q, coil shape, etc.) which bear directly or indirectly on the selection of the core material. The guidelines furnished should require only one additional ingredient for the successful selection of cores — namely experience.

FREQUENCY	RECOMMENDED MATERIALS
20 kc	Carbonyl L, HP Magnetic Powders MP-1, MP-24 C.K. Williams IRN-2, IRN-31
100 kc	Carbonyl L, C, HP, E Magnetic Powders MP-1 C.K. Williams IRN-3
465 kc	Carbonyl E
4.3 Mc	Carbonyl TH, SF
10.7 Mc	Carbonyl TH, SF
27 Mc	Carbonyl TH, SF, J
41 Mc	Carbonyl SF C.K. Williams IRN-8, IRN-9
60 Mc	Carbonyl SF C.K. Williams IRN-8, IRN-9
100 Mc	Carbonyl SF C.K. Williams IRN-8, IRN-9

Fig.3-10 Core Materials Having High Q at Various Frequencies.

NOTE: The above data was taken from the Final Report by Radio Corporation of America on Signal Corps Contract No. DA-36-039-SC-42475, "Study and Application of R-F Cores to Miniaturization of R-F Coils", covering period of July 1, 1952 to March 31, 1954. This chart was prepared from data taken on cores of the various materials as supplied by several core manufacturers. These materials produced the highest Q in the selected test coils. For further details of the test coils it is recommended that the referenced report be studied.

BIBLIOGRAPHY

Iron Cores At Radio Frequencies –
Foster and Newlon
Proceeding of the IRE – May 1941 - pg.266

Powdered Magnetic Cores –
W.J.Polydoroff
Tele Tech – Feb. 1953 – pg.69

Incremental Permeability Tuning –
W.J.Polydoroff
Radio – Oct.1944 – pg.31

Permeability Tuning –
W.J.Polydoroff
Electronics – Nov.1945 –pg.155

The Theory and Design of Inductance Coils –
V.G.Welsby – Mac Donald & Co., London (1950)

Carbonyl Iron Powders –
George O. Altman
FM and Television – June 1949 – pg.29

Radio Frequency Cores of High Permeability –
George O. Altman and Dr. H. Beller
Electronic Industries – Nov. 1945

Effective Permeability of High Frequency Iron Cores –
W.J. Polydoroff and A.J.Klapperich
Radio – Nov. 1945

Losses in Magnetic Ferrites –
G.S.Hipskind
Electrical Manufacturing – Aug.1955 – pg.108

Basic Core Materials –
Electronic Equipment – March 1955 – pg.30

Magnetic Ferrites, Core Materials for High Frequencies –
Snyder, Albers – Schoenberg and Goldsmith
Electrical Manufacturing – Dec. 1949

Evaluation of Magnetic Powdered Core Materials –
D. Elders and Dr. E. Both
Signal Corps Engineering Laboratories,
Components Division,
Electronic Parts and Materials Branch,
Ft. Monmouth, New Jersey

Study and Application of R-F Cores to Miniaturization of R-F Coils –
Final Report on Contract No. DA-36-039-SC-42475
July 1, 1952 to March 31, 1954
Contractor, R.C.A.
Signal Corps Engineering Laboratories,
Components Division,
Electronic Parts and Materials Branch,
Ft. Monmouth, N.J.

Section 4

ELECTRONIC HARDWARE

INTRODUCTION

As used in this manual, the term *electronic hardware* is a broad and inclusive expression taking in such items as solder lugs, terminals of all types, screw machine parts, eyelets, bushings, mounting brackets, tension devices, and coil form holders – to name a few. For want of a better place, solder and soldering fluxes are also included in this discussion.

It is in the selection of such items as those listed above that the designer may easily affect both the price and performance of his unit. Over-design – that is to say, specification of parts with accuracies or other features in excess of requirements – will rapidly run up cost on an end item. It is also true that too much attention to cost and not enough to the detailed requirements of the specific application in question may bring forth a transformer whose performance will be far below anticipated standards.

MANUFACTURING PROCESSES

A knowledge, however basic, of the way in which items are made can assist an engineer in making a choice consistent with both cost and performance. Toward this end a number of manufacturing processes including punching, stamping, rolling, spinning, machining, die casting, surface and centerless grinding, and others of equal importance are briefly described in the following paragraphs. Suggestions for economical practical design procedures are included for most processes.

PUNCHING

Punching is a general term for all cutting operations performed on a punch press. In general, it is a high speed operation most commonly carried out on strip stock which may be either metallic or non-metallic in nature. The work is done in punch presses of a size suited to the particular job, and the final operation is usually the result of lesser operations performed on the stock as it progresses through the various stages of multiple dies.

There are many common punch press operations – among them *blanking*, which consists of cutting the outside contour of a punched part. When a cutting operation is conducted inside a punched part, it is called *piercing*. Contour cutting and or

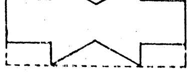

Fig. 4-1 Blanking

piercing of materials such as rubber, lead, fibre, paper, etc. are carried out by pressing a sharp, thin, steel edge through the sheet material by a process known as *dinking*. The operation which shears a punching from a strip or bar is known as the *cut-off*. A special form of piercing frequently used for locating terminal boards within a shield is the process of *lancing*, which is a special form of piercing where the entire contour is not cut but the blanked material remains as a tab.

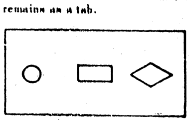

Fig. 4-2 Piercing

ACCURACY OF PUNCHED PARTS

Punched parts may be said to have the accuracy of the die that made them. For this reason it is important to consider carefully the accuracy required in the finished parts, as by so doing it is possible to influence the cost – and the life – of the die.

Barring accidents, die life is largely a factor of the dimensional variations that can be tolerated in the punched parts. It is common practice in die design to anticipate wear and to begin the operation of the tool on that side of the dimensional tolerance which is opposite to the direction of wear. For example, when a hole is to be punched in strip stock, the punch will wear more rapidly than will the female portion of the die. As a result, the size of the hole becomes smaller as the die wears. If a hole with a diameter of 0.250 inch is to be punched, and its diameter held with a tolerance of ± 0.003 inch the die maker would start the tool at a point where the hole would be running somewhere between 0.252 and 0.253 inch because he knows that as the tool wears, the size of the hole will gradually become smaller; and by starting with it on or near the maximum allowable dimension, the useful life of the tool will be considerably extended.

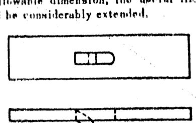

Fig. 4-3 Lancing

DESIGN OF PUNCHED PARTS

In the design of punched parts, it is well to remember that blank shapes should be constructed with true radii wherever possible unless the blank can be made from strip stock with a cut-off tool in which case square corners would be preferable to rounded corners. Any projection on the piece or any slot within its contours should have a width at least one and one-half times the thickness of the material when the material is 1/16 inch and heavier, and for thinner materials, sections should be not less than 3/32 inch for economical manufacture and tool maintenance. To punch holes smaller than those recommended above would involve the use of punches too fragile to stand up under continuous operation.

In the interest of both original tool cost and tool maintenance, round holes should not be specified as having a diameter less than the material thickness, nor should any hole be smaller than 0.030 inch. The distance from the edge of a hole to the edge of either another hole or of the blank itself should be at least equal to the thickness of of the stock and as much greater as is possible in order to avoid distortion of the hole or of the edge of the blank.

To avoid excessive tool cost, tolerances on all dimensions of punched parts should be as large as possible. Generally speaking, a tolerance of ± 0.010 inch on outside dimensions and on the distance between holes will permit economical die manufacture with a minimum of grinding operations. It is interesting to note that normal wear of a blanking die will result in a larger piece, which fact means that outside dimensions actually need carry only a plus tolerance, while pierced openings, as has been previously shown, grow smaller with wear and therefore need carry only a minus tolerance. To be sure, the hole in the die which accommodates the piercing punch also becomes larger with wear, thus serving to increase the clearance between the punch and and the die — a condition which adds to the burr on the finished part. For those cases where excessive burr will interfere with the successful operation of the part, the maximum allowable burr should be specified; however, the smaller the permissible burr, the shorter the productive life of the tool.

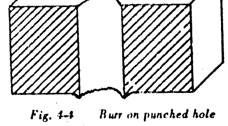

Fig. 4-4 Burr on punched hole

BENDING AND FORMING

The distinction between *bending* and *forming* may be of interest to a designer. *Bending* consists of a plastic deformation which exceeds the elastic limit of the material and is an operation which is carried out by clamping the piece along one side of the bending line and lifting it along the other side, thus forcing a bend along the edge of the clamp. *Forming* is a die process in which the metal is made to conform to the die shape with the result that the inner radius of curvature can be held accurately. Forming may be an act by itself or it may take place within a multiple die where it is carried out in conjunction with piercing, blanking, or other similar operations.

One operation which is particularly difficult to control is forming a 90-degree angle, which means that if some deviation cannot be permitted, the necessary forming die will be greatly complicated. The grain of the metal is important in either bending or forming, and bends should be made at right angles to the grain of the metal wherever possible; otherwise cracking is likely to occur along the bend.

SPINNING

Spinning is a method of forming round sheet metal parts at comparatively low cost. Aluminum, copper, zinc, brass, and certain steels, including stainless steel, are among the materials which are commonly processed in this manner. The work is carried out on a lathe similar to a wood-turning lathe except that it is larger and more powerful. A form, usually of wood, is first made to the exact shape and size of the article to be spun. Fastened to the head-plate of the lathe, this form has a round sheet-metal blank pressed against it by the tail stock of the lathe. When the machine is started, both the blank and the form are rotated at relatively high speeds; and the operator, beginning near the center of rotation, uses either hard wood or metal rods to push the metal until it completely covers the wooden form, thus shaping the desired object.

UPSETTING

Punch press operations are not necessarily limited to flat stock but may be performed upon wire. A relatively common process of this sort, known as *heading*, is an upsetting operation which serves to gather a mass of material for producing, for example, screw and rivet heads.

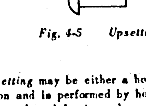

Fig. 4-5 Upsetting

Upsetting may be either a hot or cold working operation and is performed by holding one portion of the metal and forcing other portions toward it. *Cold heading* is an upsetting process by which many small parts are produced in an economical manner from wire stock. In this operation, the material is supplied in wire form, and the heading machine takes the wire, cuts it off at the required length, and carries it through the forming operation in a completely automatic manner at rates of from 150 to 300 pieces per minute. Any malleable metal can be used in this process so long as it does not work-harden excessively. Low carbon steel, low alloy steel, and almost all non-ferrous alloys work well in this process.

KNURLING

Knurling is a cold working process in which a series of sharp serrations on a hardened steel roller are pressed into the material being knurled. This process is used to roughen surfaces for thumb screws and other devices to be turned by hand and also to roughen the surface of metal parts designed for embedment in plastics. Knurling is sometimes applied to the surface of a stud over which a coil form must be pressed. In such cases, the knurl serves to improve the tightness of the fit, since the roughened

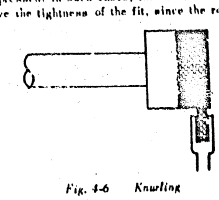

Fig. 4-6 Knurling

surface of the stud consists essentially of tiny pyramids of metal which will deform slightly, thus providing a means of minimizing the effect of minor variations in the sizes of the two pieces. A secondary advantage of knurling in this sort of application is found in the increased strength obtained when the mating parts are cemented. This improvement is, of course, the result of the many small recesses in the metal in which the cement can harden and cling.

The knurling operation is usually performed on screw machines or on lathes.

DRAWING

Drawing is a stretching process which may be performed in two ways. In one method, metal is pulled through a die as in the manufacture of tubes, rods, or wire. In another form of drawing, material is pressed into a die by a ram which causes the material to drag along the walls of the cavity, thus producing cups or shells. It is this type of deep drawing which is used in the manufacture of the shield cans commonly used in electronics.

ROLL THREADING

Closely related to heading is the process of *roll threading*. This operation is carried out by rolling the part to be threaded between two hardened serrated, plane dies that form the threads by squeezing the metal and causing it to assume the form of threads. The blanks are fed automatically, making this a high-production, low-cost operation which produces threads meeting the requirements of a Class 2 Fit and which are actually stronger than cut threads.

EXTRUSION

It is frequently desirable to provide for threaded holes in relatively thin stock. This is done by calling for *extruded* holes which are formed by a swedging process which is actually a cold die-forging operation in which metal is confined and thus forced to flow into a cylinder of sufficient length to provide the number of threads required for safe operation.

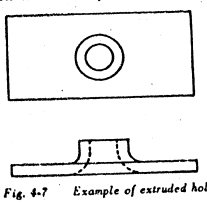

Fig. 4-7 Example of extruded hole

COLD ROLLING

Cold Rolling is a drawing operation carried out between rolls to which no torque is applied. This operation is performed below the recrystallization temperature for the purpose of achieving smaller size, closer tolerances, increased hardness, and higher tensile strength. A majority of the strip stock and sheet stock used for metallic parts in electronic equipment is of the cold-rolled variety.

LATHE WORK

There is probably no single machine of greater versatility and more overall usefulness than the engine lathe. The basic lathe operation consists of turning cylindrical forms by rotating the work against the cutting tool which is arranged to move parallel to the axis of the work rotation. By moving the cutting tool perpendicular to the axis of rotation, it is possible to form plane surfaces and to cut off the work. Drilling, boring, reaming, tapping, threading, contouring, and knurling are among the operations which are commonly performed upon engine lathes.

Metals, wood, and plastics may be easily worked on a lathe by selecting the proper tools and speeds. Ceramic materials in the green (unfired) state, glass bonded mica, and powdered metal objects, such as the iron cores for use at radio frequencies, may also be worked on a lathe, but in the use of such materials care must be taken to avoid damaging the machine by the abrasive action of the dust formed by the machining process. It is important to keep all parts of the lathe as free from this dust as is possible and particularly to keep the ways clean and well lubricated. The dangers of wear due to abrasive dust are even greater in the case of materials such as ferrites which are often threaded on a lathe by means of a tool-post grinder.

SCREW MACHINE PARTS

So-called *screw machine parts* are relatively common in precision-built electronic equipment. These are parts which are formed from rod or bar stock of various cross-sections and may be in the form of bushings, studs, screws, or other shapes which would normally be formed on a lathe. A screw machine is essentially a turret lathe operating in an automatic manner and performing a required series of operations in sequence. As a general rule, automatic screw machines are used only for large quantity runs since tooling for each part is an individual affair and machine set-up is a lengthy and involved process. Turning, drilling, threading, tapping, and and counter-boring are among the operations which are commonly performed on automatic screw machines.

In the design of small parts to be made on an automatic screw machine, care should be taken to avoid square shoulders and sharp corners on the finished pieces. Acceptance of parts bearing " machine finish " is also important for reasons of economy, since grinding is not possible on a screw machine, and to insist upon a ground finish would necessitate transferal of parts from the screw machine to another machine for the grinding operation.

DRILLING

Drilling is a process which is so common that it will not be discussed here in any detail. However, it is well to remember that good design calls for the use of drills of standard size, the avoidance of unusually small drills, the avoidance of deep holes and blind holes (those in which the drill does not break through the material), and, if blind holes are a design necessity, allowance of space for the

drill point at the bottom of blind holes. A drill must be considered as a roughing tool and not as a finishing tool. This means that if holes must be finished

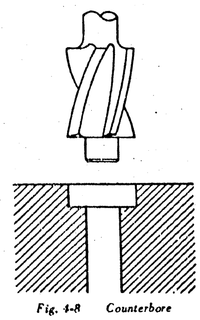

Fig. 4-8 Counterbore

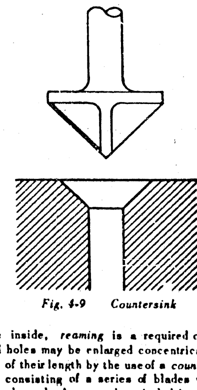

Fig. 4-9 Countersink

on the inside, *reaming* is a required operation. Drilled holes may be enlarged concentrically for a portion of their length by the use of a *counterbore* – a tool consisting of a series of blades which cut only on the ends. A counterbore is led into the hole by a pilot which is slightly smaller than the drilled hole. A *countersink* is used when it is desired to *chamfer* the edge of a drilled hole. This is often done to remove the burrs from the edge of a hole or to provide easy entrance for a tap.

TAPPING

Tapping is a finishing operation used to produce an internal thread in a hole. A tap consists of several cutting blades attached to a supporting body. The end of the tap which enters the hole is beveled for a distance of from two to five times the distance between threads. It is this beveled portion which does the cutting of the internal thread, and the remainder of the tap serves to drive the cutting end into the hole. The majority of the chips formed by the tapping operation go ahead of the tap; hence in blind holes provision must be made for accumulation of these chips at the bottom of the hole. Tapping is made easier if the hole is countersunk first, and good design calls for the use of standard threads of a fit no more accurate than is absolutely necessary for the successful operation of the part.

THREADING

External threads may be produced on a lathe by the use of a single-point tool. The more common method of producing external threads is by means of a threading die, consisting of several cutting blades held in such a manner that when rotated with respect to the article to be threaded, threads of the required characteristics are cut on the outer surface of the material. Externally threaded parts should utilize a standard thread, and a chamfer should be provided at the start so that threading will be facilitated.

The accuracy of fit of threaded parts has been covered in *Screw-Thread Standards for Federal Services* – a publication which is available from the United States Government Printing Office. Beyond the general reminder that screw threads are classified into Class 1 through 5 Fits, with a larger number indicating a closer fit, the reader is left to refer to the above-mentioned publication for more specific information regarding thread forms and tolerances. Class 2 Free Fit is most commonly used thread in practice.

GRINDING

Grinding is a very common finishing process for metallic and certain plastic parts. For those cases where machine finish is not acceptable, grinding is most often specified. Materials which are too hard to permit ordinary machining — ferrites, for example — are usually shaped to final form by grinding.

The actual process of grinding is carried out by a grinding wheel which is made up of tiny particles of crystalline grit bonded together by a material such as rubber, shellac, or phenolic resin. For best results, the cutting surface of the wheels should travel at a speed of between 3000 and 5000 feet per minute.

The nature of grinding wheels is such that they may be considered as self-sharpening cutting tools since as the surface grits become dull, the increased forces needed for cutting serve to tear out the dull particles and expose new, sharp grits with which to continue the cutting process.

A great advantage of grinding as a precision finishing process is found in the thinness of the chip which may be taken. Actually, it is necessary only to exceed slightly the thickness of the cutting edges of the grit, which, in the case of fine wheels, means an exceedingly thin chip. This fact coupled with the very light wheel pressures required makes grinding an ideal means of extremely accurate machining.

CENTERLESS GRINDING

Centerless grinding is an operation of interest to the electronics designer because of the fact that it is used as a means of sizing tubing used for coil forms. This operation is carried out on a machine consisting of three basic parts — the grinding wheel, the regulating wheel, and a work rest. Since it has no shoulders, tubing can be centerless ground by the "feed through" method in which the work is passed from one side of the machine to the other between the wheels, entering the machine with variable OD and possibly with an uneven surface, and emerging from the machine with a uniform OD and a smooth, ground surface.

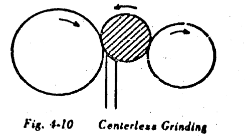

Fig. 4-10 Centerless Grinding

The principle of the machine is simple with the grinding and regulating wheels rotating in the same direction and with their outer surfaces separated by a distance somewhat less than the diameter of the piece being ground. The result of this arrangement is that the piece which is being ground is supported between the surfaces of the two rotating wheels and a tool rest extending upward between the wheels. With its rotational speed and rate of progress through the machine controlled by the regulating wheel, the work receives a fine, dimensionally accurate finish in a process well adapted to the large scale production of cylindrical parts.

LAPPING AND HONING

For those cases where finer surfaces or closer tolerances than can be obtained by grinding are required, *lapping* or *honing* will usually prove satisfactory.

Lapping is a process utilizing fine abrasives carried in oil and applied to the work surface by a device called a *lap*. Laps are usually made of a metal which is somewhat porous and soft — lead and and cast iron are common materials — and shaped according to the surface which is to be lapped. Lapping is done with a reciprocating motion and produces an extremely fine finish on the work.

Honing is a finishing process most often applied to internal bores. A *hone* consists of one or more sticks of fine-grained, bonded adhesive which are applied to the work by a combination of reciprocating and twisting movements. The result is a somewhat crisscross surface pattern of fine grain. Honing is a means of maintaining very close tolerances.

MILLING

Milling is a process whereby excess material is removed from a piece by means of a revolving cutter of an appropriate size and shape. Milling

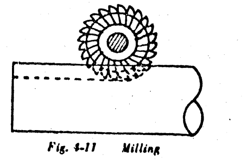

Fig. 4-11 Milling

cutters are available in a wide variety of forms including the plain milling cutter which cuts only on the periphery and is particularly useful for machin-

ing flat surfaces. For milling keyways, slots, or grooves a side milling cutter is used. Cutters of this type are supplied with teeth not only on the outer circumference but on the sides as well, thus having three cutting surfaces in contact with the work.

End milling cutters, supported and driven at one end and with teeth on both the periphery and on the end, are useful for reaching surfaces that cannot be reached with conventional cutters. While this type of cutter cannot bore directly into a piece because of lack of teeth at the center, it can provide required radii within cavities, machine flat spots on sliping surfaces, cut away the radius left by a side milling cutter at the end of a keyway, and perform other similar operations.

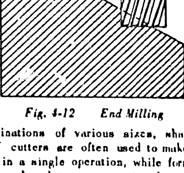

Fig. 4-12 End Milling

Combinations of various sizes, shapes, and types of cutters are often used to make special contours in a single operation, while form cutters can be employed to provide curved or otherwise special cross-sections.

In the design of parts that must be made by milling, it is important wherever possible, to allow for the use of standard milling cutters, thus avoiding the expense and delays involved in procuring special cutters.

BROACHING

Broaching is a very fast and relatively simple means of providing a desired contour — usually internal, although surface broaching is becoming an increasingly common operation. The process consists of pushing or pulling the tool, called the *broach*, through (or across) the work. A broach resembles a coarse file in that it is provided with a large number of cutting edges so arranged as to gradually change the contour of the work from that of the original piece to the desired final form.

Broaching is a fast operation since one pass over the work is all that is required. The pieces are left with a good finish because the tool has fine teeth in that portion which is last in contact with the work. For the most part, broaching is limited to large-scale operations since both the tools and the machinery for using them are initially expensive.

IMPORTANCE OF PROCESS INFORMATION

It is recognized that the foregoing are by no means the only ways in which the many small, relatively precise parts used in electronic components are machined or manufactured. The primary purpose of this discussion of fabrication methods is purely to make an engineer engaged in the design of high frequency transformers cognizant of some of the problems faced by those who must produce his component parts. It is common knowledge through-out the industry that far too often a design engineer will work out a new design in his laboratory, forgetting completely that while there is no particular problem connected with making one or two of almost anything, there well may be serious problems connected with quantity production of the same items. It is hoped that this discussion, brief though it may have been, will help to prevent issuance of specifications which are not practical and therefore must be revised at the cost of both time and money before production can begin.

TERMINALS AND SOLDER LUGS

Terminals and solder lugs are available from a number of sources[1] in a wide variety of sizes and shapes. Almost without exception, the material used in these very necessary bits of equipment is brass. In the case of small solder lugs or ground straps, copper is sometimes used; and for those cases where successful operation depends upon spring action, beryllium copper may be specified. As will be pointed out later in this section, the use of any metal other than one of non-ferrous composition will make soldering difficult even though the surface be electroplated with an easy-to-solder metal.

Good terminal design involves certain practical considerations. To begin with, it is best to utilize existing forms and sizes of terminals rather than to insist upon new designs unless the new terminal will materially improve the performance or the usefulness of the end product. Terminals which can be produced by other than lathe or screw machine techniques are highly desirable from a cost angle and

[1] Aircraft-Marine Products, Inc; Bead Chain Manufacturing Company; Cambridge Thermionics Corporation; U.S. Engineering Company, Zierick Manufacturing Corporation; and many others.

rarely is there a need for greater accuracies in dimensions than can be obtained from upsetting, cold heading, or other similar and more economical processes.

It is important to remember that a well-designed terminal will make it easy to connect lead wires in a manner such as to make the connection mechanically secure before soldering — a requirement that is extremely important. The size and type of the wire which will be used for the connection must be considered. Generally speaking, the use of terminals with holes through which the wires must be pushed in not recommended because of the extra expense involved in this operation. This fact is especially true for those cases where stranded wires are used since only a slight excess of solder on the wire following tinning or a bit of fanning of the individual strands will often be sufficient to prevent entrance of a lead into a terminal hole.

In connection with terminal design, it is well to give some consideration to the means to be employed in holding the terminal to its insulating board. The most common methods of fastening terminals can be listed under the four general headings of *fastening by the use of screws, riveting, spinning, and staking.*

Of the above listed methods, the most expensive and the least used is probably the method involving the use of screws. If a screw is to be used to hold a terminal in place, it means that the terminal must itself be either tapped or threaded, as the case may be. A lock washer is required if the terminal is to remain tight under normal conditions of use which means additional parts to handle and a resultant increase in cost.

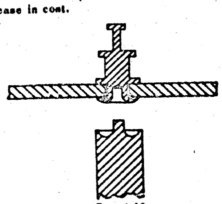

Fig. 4-13
Riveting as a means of mounting terminals

Riveting is a process whereby that portion of the terminal which protrudes through the terminal board is crushed, rolled, or otherwise deformed in such a manner as to prevent withdrawal of the terminal. Properly done, this method provides adequate holding power. Two methods of preventing rotation of a riveted terminal are often used. One method calls for the use of square holes in the insulating board so that the riveting action will cause the metal to crowd outward and into the corners of the square, thus effectively preventing twisting of the terminal. A second method for keeping terminals from turning requires either a knurl or a set of serrations of the shoulder of the terminal which is in contact with the surface of the plastic or other insulating material to which the terminal is attached. For most applications, either of these methods will be found satisfactory.

Spinning is a slower, more expensive method of anchoring terminals in which that portion of the metal which extends through the hole in the insulating board is not crushed but instead is actually rolled by means of a rotating tool until it is in contact with the supporting base. Adapted only to those terminals which have hollow shanks to accommodate the pilot of the spinning tool, the process not only anchors the terminals securely but at the same time eliminates from the deformed portion all protuberances and general roughness. Closely related in nature to the spinning process described earlier in this section, when properly performed, it results in uniform, smooth, well-formed, tight "rollovers" against the insulation. In the case of components which will operate at very high voltages, spinning is often resorted to as a means of reducing the points from which corona may originate.

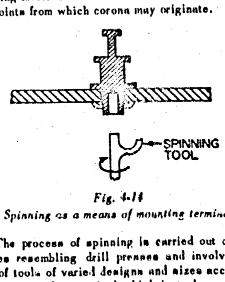

Fig. 4-14
Spinning as a means of mounting terminals.

The process of spinning is carried out on machines resembling drill presses and involves the use of tools of varied designs and sizes according to the particular terminal which is to be spun. As might be expected, the nature of the operation is

such as to require more time than riveting or staking; hence it is not wise to specify that terminals shall be mounted by spinning unless it is considered vital to the performance of the unit to have all terminals free from irregularities and projections. It should further be noted that common design practice often calls for the application of solder to terminals, frequently at the end by which they are attached, and this solder, properly applied, can almost always eliminate completely the need for spinning terminals to the insulating board.

Staking is a very quick and generally efficient means of mounting properly designed terminals. In general, this process is limited to terminals which are flat and rectangular in cross section. The operation is performed in a press and requires the use of tools specially designed for the particular terminal and application. The process of staking consists of displacing a certain amount of metal in a way that anchors the terminal to its supporting surface. Because of a general lack of resiliance in both the terminal and the insulator, it is difficult to keep staked terminals sufficiently tight to prevent movement. This feature alone is sufficient to transfer emphasis from this type of terminal attachment to one of the previously named methods.

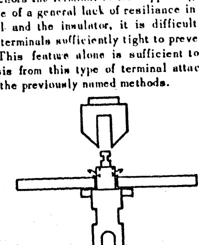

Fig. 4-15 Terminal Mounted by Staking

MOUNTING BRACKETS

Mounting brackets and similar pieces of hardware are usually stamped from cold-rolled steel although brass is sometimes used, particularly in the smaller sized pieces. As supplied commercially, hardware of this sort is most often cadmium plated after being punched from strip stock.

CORE DRIVE AND TENSION DEVICES

The increased use of permeability-tuned transformers has resulted in a number of fundamental types of devices for driving and controlling tension (torque) on the cores which tune the units. Not only must there be a means of moving the core with respect to the coil, but provision must also be included for holding the core once it has been properly located. In other words, a core drive mechanism must provide both threads and tension. In certain instances a third task is given these devices — they are called upon to hold the coil form as well as to provide drive and tension for the core.

Devices of this latter type are available in a number of sizes from at least two sources[2] in this country. The design of these particular parts has been fairly well standarized by the manufacturers. The material used is steel, so treated as to provide satisfactory spring action without the brittleness which might lead to breakage. Reference to Fig. 4-16 will show how, as these coil form holders are locked into position, they grip the coil form tightly — an operation that is facilitated by tiny points bumped out from the surface that contacts the coil form.

Depending upon the manufacturer and also somewhat upon the size of the unit, these combination coil form holders and tension devices drive the core screw by one of a variety of punched shapes which surround a hole of a diameter approximately equal to the pitch diameter of the screw. Since the material used in making these parts is relatively thin (usually about 0.010 to 0.012 inch), to drive the core it is necessary only to make those portions of the device that are immediately around the center hole conform to the thread shape, since the thickness of the material allows them to fit between successive threads of the core screw. Tension then becomes a matter either of deliberate misfit between the core screw and the formed portions of the device which make up the threads, or, as in one commercial version, it is derived from two tension fingers which rub against the core screw at a point somewhat outside the main body of the device. Since these combination coil form holders and tension devices are made from a flexible metal by a stamping process, there is sufficient uniformity among pieces to insure reasonable consistency in the torque of the core screws.

Millions of these devices have been used in commercial applications within the electronics industry. In a majority of cases they have provided satisfactory performance. In the case of military equipment, however, the requirements for resistance to shock and vibration are such as to cast serious doubts as to the wisdom of using this type of tension device in military applications. This is true

[2] The Palnut Company and Tinnerman Products, Inc.

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largely because the core screw is held by only one thread — a condition which leaves the assembly especially vulnerable to vibration unless the core when the washer is applied to the screw in the manner of a nut. The washers themselves are somewhat spherical in shape with cutouts in the form of

Fig. 4-16. Example of combination coil form holder and core tension device.

is a tight fit within the coil form, which is itself a poor design practice.

Another form of tension device frequently used in commercial applications is that known in the industry as the *umbrella washer.*

Umbrella washers are available in sizes to fit the standard screws normally used in powdered iron cores. As is the case with the previously de-

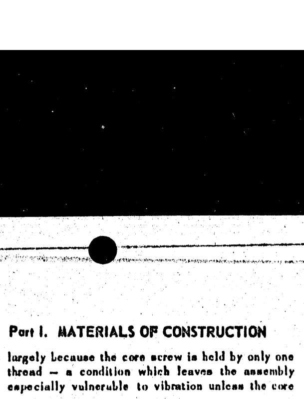

Fig. 4-17. Umbrella washer used as core tension device

scribed combination coil form holders and tension devices, umbrella washers are stamped parts having sections so shaped as to engage the screw threads teeth around their outer peripheries.

To use this form of tension device, the core screw is first threaded through a standard bushing. The umbrella washer is then screwed down over the core screw until its periphery contacts the shield can or base which carries the bushing. Since the washer is made of a springy material such as beryllium copper or spring brass, tightening it against the base compresses the washer, thus applying tension to the screw as a result of the stresses set up between the washer and the bushing. At the same time the teeth are digging into the base — usually provided with radial serrations or holes — thus effectively preventing rotation and subsequent loosening of the washer. Fig. 4-17 shows the operation of this type of tension device and the stresses that are developed in producing the tension.

Study of the above-referenced drawing will show this type of tension device to be dependent for its operation upon the compressibility of the washer. Should the washer collapse — as it may if overtightened — no tension will be applied to the screw. It is therefore apparent that the tension imparted to a screw by an umbrella washer is largely determined by the physical properties of the metal from which the washer is made.

Some years ago, this type of tension device was extremely popular in civilian applications. Recent years have seen the development of simpler and cheaper means of providing tuning cores with sufficient tension for average operating conditions. The principle upon which this device is based is sound, however, and for certain very small designs

where size and weight are important, it is entirely possible that the umbrella washer could provide satisfactory tension in military components.

Next on the list of ways in which torque control can be imparted to a core screw might be listed the *D-spring* type of tension device. The principle of operation in a D-spring tension device is simple. The device is made up of a threaded bushing of a length sufficient to leave room for a slot extending a little less than half way through the bushing at approximately right angles to the axis of the core screw. The width of this slot is such as to provide clearance for a D-shaped clip made from spring wire of a size that will engage the threads of the screw and locate itself at about the pitch diameter.

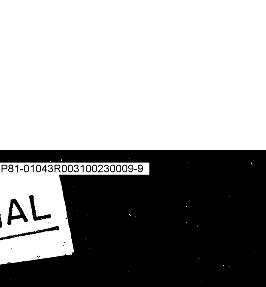

Fig. 4-18 D-spring type tension device

Two schools of thought govern the design of of the slot. The first group holds that this slot should not be cut at exactly right angles to the axis of the core screw but rather should be parallel to the threads of the screw. If the bushing is designed in this manner, tension becomes simply a matter of holding the screw against the back side of the bushing with the resultant tension on the screw a function primarily of spring strength, bushing length, and the finish of the threaded parts.

A second group is of the opinion that the spring slot should be cut exactly 90 degrees from the axis of the screw so that tension becomes a matter not only of contact with the rear area of the bushing but also of forces developed as a result of a deliberate misalignment of spring and threads. It is true that thread wear is somewhat aggravated in this design, but since in the average case it is not necessary to tune a transformer at other than infrequent intervals, the extra wear caused by the slot angle can be considered inconsequential.

The major weakness in both designs lies in the fact that in order to fit between the threads of the core screw, the wire from which the spring is made must be of relatively small diameter. In view of the pressure exerted by the spring upon the core and the fact that tuning is accomplished by relative motion between the screw and spring, it is apparent that it will be extremely difficult to protect the spring against rusting and/or the effects of a corrosive atmosphere. Steel springs protected by cadmium plate followed by Iridite or other protective finish are probably as satisfactory as any, and unless subjected to an excessive amount of adjustment, should provide adequate service.

The performance of D-spring type tension devices is generally satisfactory. The action of the core is smooth, and torque can usually be held within reasonable limits over a considerable period of time, as has been demonstrated in numerous civilian and military applications.

For those cases where the ultimate in protection against the effects of shock or vibration is a design requisite, the use of a *split bushing* provided with a locknut is indicated.

As shown in Fig. 4-19, the basis of this type of tension device is a bushing with four slots 180 degrees apart extending the greater part of its length, threaded on the inside to receive the core

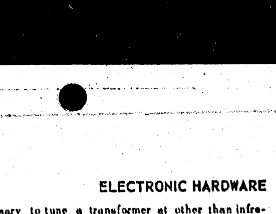

Fig. 4-19 Split bushing tension device

screw and on the outside to receive a locknut. The outside threads are slightly tapered so that as the locknut is tightened on the bushing, the two threaded portions are compressed against the core. By

POOR ORIGINAL

controlling the tightness of the locknut, any desired degree of torque can be applied to the core, or it can be locked against all rotation.

For those cases where extreme conditions of shock or vibration are anticipated, the use of a locking nut with a Nylon or fibre insert[5], or of two nuts tightened one against the other, can be used to provide — and hold — almost any degree of torque up to and including locking against all rotation.

A major objection to this type of core drive lies in its expense. Not only must the bushing be formed, tapped, and threaded — all operations which can be performed on an automatic screw machine — but the slots must be either milled or sawed in a separate operation which adds greatly to the cost of the bushing. A second objection to this type of tension device is the necessity for two different tools when adjustments must be made. Depending upon the form of core screw used, previously described tension devices require only a screwdriver, socket wrench, or other tuning tool to operate the core screw. In the case of a split bushing equipped with a single locknut, some sort of end-wrench is needed, and the tuning process becomes considerably complicated.

It is generally conceded that this type of core drive with its one or two locknuts provides the best control of tension and locking that is presently available. The expense of the required parts and the complications attendant upon their use have greatly retarded the acceptance of this type of core drive in all but the most critical components. Since the probabilities are that another tension device of simpler construction will, in most instances, prove adequate, it seems wise to recommend that that split bushings and locknuts be specified only when no other system will prove the required protection against shock and vibration.

An inexpensive version[6] of the split-bushing type of tension device is shown in Fig. 4-20. Stamped from steel strip stock by a combination of processes, a single part is formed which provides both core drive and tension with the drive coming from threads on the inner surface of each of the two extruded portions extending upward from the base. Tension of torque control is developed by the pressure of the two threaded half-sleeves which are so located as to be pushed apart slightly when a core screw is inserted. Torque is therefore dependent upon the pressure exerted by the two halves of the bushing.

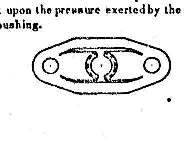

Fig. 4-20 Stamped split bushing type core tension device.

This type of core drive has been used successfully in a number of applications — both civilian and military. A major criticism is to be found in the fact that no control of torque is possible other than through selection of mating parts — a process which is never good production practice. The basic principle of this device is, however, sound, and in non-critical applications this type of core drive can be expected to give satisfactory results.

A somewhat similar type of tension device is basically a split bushing except that in this case, the cutout is at 90 degrees to the screw rather than parallel to it. The remaining collar is then crushed downward almost to the main section of the bushing where it remains supported by two arms formed from the uncut portion of the original bushing.

Designed to serve more as locknuts than as tension devices, this type of core drive mechanism depends upon misaligrment of the bushing and the collar. Both the collar and the bushing are threaded so this misalignment may be either lateral or longitudinal — either type tending to make rotation difficult.

Actually this type of core drive does not make a very satisfactory tension device. The use of steel as a material, combined with the introduction of two 180-degree bends in the slender arms which support the collar, makes this a rather rigid device without the flexibility necessary in a good torque controlling device.

This type of core drive has been used in certain military transformers with units so equipped showing wide variations in torque. Assembly of units using this type of tension device is often difficult and may require adjustment of the collar position before becoming possible. Then too, breakage of the bent arms is by no means infrequent. The use of this type of core drive mechanism is not recommended for either military or civilian applications.

[5] Elastic Stopnut Corporation of America, 2330 Vaux Hall Road, Union, New Jersey

[6] Tinnerman Products, Inc.

MISCELLANEOUS HARDWARE

Since such items as bushings, studs, eyelets, coil form holders, and most of the other small parts used in electronic components are made from free-turning brass, they will, for purposes of this discussion, be considered in one group.

It is well to start the process of specifying such items by checking closely on the standard parts which are available from suppliers already fully tooled to produce them. Selection of such parts always tends to lower cost, to expedite delivery, and to simplify the matter of supply when several manufacturers are to produce the same transformer. Because these standard parts represent the thinking of many design engineers, it is quite probable that in a majority of cases it will be found possible to select nearly all the required hardware from the catalogs of firms regularly engaged in the production of such items.

GOOD DESIGN PRACTICES

Efficient design will, however, often create a demand for small parts which are new and different. It is in the design of such articles that economies or extravagances can be introduced with comparative ease. One fundamentally important design concept is to maintain maximum simplicity in the finished part. Every change in contour and every individual operation that have to be performed in the production of the piece cost money and should be specified only if they serve a useful purpose. Nothing in this discussion is meant to imply that performance should be sacrificed for simplicity. Rather, the intention is to stress the dangers inherent in careless design practices and to emphasize the importance of carefully analyzing every new part to determine whether it can be further simplified without affecting its performance.

As a means of showing what may be accomplished by approaching a new design in a critical manner, the steps that usually precede the introduction of a new transformer design may be considered. Once the basic idea has been formulated and the preliminary sketches made up, the next step is the preparation of a laboratory model. Right here is the place where many design engineers go astray. It is only natural for an engineer to concentrate upon overall performance rather than on design of the individual parts which go to make up the transformer. It is therefore a relatively common practice to go into the Model Shop, pick up whatever stock is available, and turn out on the lathe a part that will do the job at hand — all the time with little attention paid to dimensions, or even to material. This in itself is not necessarily a bad practice. The danger arises later when working drawings must be prepared for the pilot run or for production. Too often are final drawings based on measurements of the original parts used in laboratory models with little or no thought given to improvements or economies in their design.

Regardless of what may have been used in the laboratory models, good engineering procedures demand that at this point consideration be given to specifications that are in accord with normal commercial practices. For example, if free-turning brass is normally supplied in rods having diameters of 13/32 or 7/16 inch, it would seem that in most cases it would serve no useful purpose to specify a bushing with a maximum OD of 0.415 inch. To make such a bushing would require the use of the larger size of brass rod which would then have to be turned down to the specified diameter. Good design requires that sizes should, wherever possible, conform to the standard rod or bar sizes normally available from producers of the raw material. Tolerances should conform to standard commercial practice and should be as broad as possible. In other words, a careful designer will work with available materials and techniques and will resort to new and special designs only when convinced that no other means will provide the desired result. Adherence to these policies will go far toward standardization of components and the elimination of unnecessarily expensive designs.

SOLDER AND SOLDERING FLUXES

The process of soldering is a common one that enters into the manufacture of almost any electronic component or equipment. It is, generally speaking, a simple process although analysis of a soldered connection will show a well-soldered joint to be the result of complex chemical and physical action.

COMPOSITION OF SOLDER

Solder, and this discussion will concern itself only with *soft solders*, is a fusible alloy consisting of tin and lead combined in various proportions, usually in the range of from 40 parts of tin with 60

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parts of lead to as much as 70 parts of tin with 30 parts of lead. The conventional way of describing solder is in terms of its tin-lead content as for example, 55/45 meaning 55 parts of tin with 45 parts of lead, since tin is always the first metal named. It is theoretically possible to have soft solders of any composition as long as they consist solely of tin and lead. Commercial branches of the electronics industry have settled largely upon those solders in the range of 40/60 to 60/40, with 50/50 the most widely accepted formulation. Federal Specification QQ-S-571b sets forth the composition which must be met in the various soft solders and lists 60/40 as the type recommended for use on electrical connections.

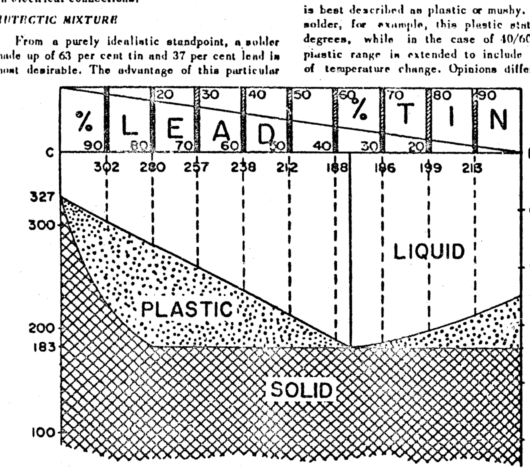

Fig. 4-21 Eutectic chart

EUTECTIC MIXTURE

From a purely idealistic standpoint, a solder made up of 63 per cent tin and 37 per cent lead is most desirable. The advantage of this particular alloy, known as the *eutectic combinations*, lies in the fact that the melting point is sharp – that is, this proportion of tin and lead is either a liquid or a solid depending upon the temperature, and there is no point at which the alloy has a plastic consistency. A further point of interest concerning the eutectic is its low melting point – 361 F – the lowest of all tin-lead alloys, and a particular asset when soldering in close proximity to heat sensitive materials.

Fig. 4-21 is the familiar eutectic chart showing the physical state of all proportions of tin-lead alloys at temperatures from 361 F up to 620 F, the melting point of lead. It will be noted that when the proportion of tin and lead is 63/37, solder passes directly from a solid to a liquid without going through any mushy stage. With any other proportion of tin and lead, there is a temperature range in which the solder is neither a solid nor a liquid, but is best described as plastic or mushy. With 60/40 solder, for example, this plastic state covers 9 degrees, while in the case of 40/60 solder the plastic range is extended to include 99 degrees of temperature change. Opinions differ as to the effect of this plastic period upon the quality of work obtainable from the various grades of solder. One school holds that the presence of a plastic interval is an invitation to movement within the joint during the cooling (plastic) period, and the result is likely to be poor soldered joints subject to easy fracture. This same school contends that since all solders in the liquid state contain eutectic solder as well as the basic combination, prolonged extension of the plastic condition creates a tendency toward crystalline structure within the joint. In support of this contention reference is made to photomicrographs[1] which show that the more one departs from the 63/37 eutectic, the more crystalline the structure of the solder becomes and the lower its tensile strength.

The second school of thought is strongly of the opinion that unless there is a period during which the solder is plastic, there is grave danger of fracture in the joint as a result of vibration occurring at the moment of set. Actually, there seems little reason for this feeling since when solder solidifies, the process is one of freezing, which means that a considerable amount of heat must be liberated even though the temperature does not change. For this reason there is a definite time interval during which crystals are forming in the eutectic, making "movement at the moment of set" a remote possibility at best.

The electronics industry has, in its commercial work, used solders ranging for the most part between 40/60 and 60/40. Experience acquired during many years of study of representative solding operations has shown that substantial savings in time can be effected by using 50/50 solder rather than 40/60. Whether these savings are due to the 16 degree shorter plastic period, to the more homogeneous structure of the solder joint, or merely to the lessened time between the flow and the set of the solder is not clear. It does, however, seem safe to draw upon the experience of the industry and recommend that solder consisting of at least 50 per cent tin be used in electronic applications.

THE SOLDERING PROCESS

The actual process of soldering is both interesting and complex. When two pieces of metal are soldered together, a new alloy, formed from the solder and the metal being soldered, is created. This means that an actual solvent action takes place at the point of joining. Because soldering is an act involving both chemical and physical joining of metals, the strength of soldered joints is greatly in excess of that which would be due to adhesion alone. Proof of this statement is found in the fact that it is impossible to pry solder away from a metal in a manner leaving the metal in its original condition. It is equally impossible to drain or wipe all solder from a metallic surface that has been thoroughly wetted by a film of solder.

[1] "Notes on Soldering" by W.R. Lewis

In this connection, it should be noted that since soldering is a process involving chemical as well as physical action, metals which have been electroplated can be successfully soldered only if the solvent or alloying action extends through the electroplating to the base metal. In other words, the effect of electroplating is one of preserving rather than improving soldering quality – a point which helps to explain the difficulties encountered in soldering cadmium-plated aluminum, and the importance of having electroplating thick enough to protect the base metal.

It is upon the act of wetting metals that the success or failure of a soldering operation depends. It must be remembered that the process of soldering is one involving bare metals. The presence of impurities in any form including oxides, sulphides, paint, enamel, varnish, or any foreign matter, will make soldering difficult, if not impossible. It will be recalled that in Section 1, it was stated that coil leads should not be cleaned far in advance of tinning lest they acquire an oxide coating which would make them difficult to solder.

SOLDERING FLUXES

It is well known that soldering is rarely done without the use of a material known as a *flux*. The actual part played by the flux in the soldering process is probably less well known and will therefore be discussed at this point.

The primary function of a soldering flux is to insure an absolutely clean metallic surface, free from all traces of oxides. It is important to note that the flux does not enter into any chemical union with the solder but merely facilitates the wetting of the metal by keeping its surface free from oxide films.

It must be apparent that if the primary function of soldering fluxes is to remove oxides from the metals involved in the soldering process, they themselves must be chemically active substances. Since ionization is essential to chemical activity, it follows that soldering fluxes must tend to invite corrosion. This is essentially true; there is no soldering flux which is completely non-corrosive – the differences among the various types being of degree of corrosiveness only.

The following might be listed as constituting the requirements for an ideal soldering flux for use on electronic apparatus:

1. It must remove oxide films from metallic surfaces.
2. Having removed the oxide films, it must keep the surface chemically clean.
3. It must be fluid at soldering temperatures.
4. It must not remain between the solder and the metal but must give way completely to the solder as it flows.
5. It must leave a minimum of residue.
6. The residue that is left must be essentially non-corrosive and must not support the growth of fungus.
7. It must not add to the electrical resistance of the joint.
8. It must reduce the surface tension of the molten solder, thus inducing rapid flow.

TYPES OF FLUXES

It is customary to divide soldering fluxes into three general classifications: (1) Acid types; (2) Organic types; and (3) Rosin types. The basic requirements of an ideal flux set forth above at once eliminate from consideration fluxes of the acid type since all varieties are excessively corrosive. Although it must be admitted that their high degree of chemical activity results in fast and easy soldering, acid-type fluxes are entirely unacceptable for all electronic applications.

The second group, made up of weak organic acids, amines, amides, and weak organic bases, is somewhat less active than the acid group and therefore does not permit soldering with the same degree of ease. Admittedly less corrosive than the acid group, all fluxes of this type leave corrosive residues and are, therefore, unacceptable for use in electronics.

It is only the third group of soldering fluxes — that having pure, natural *rosin* as the basic ingredient of all formulations — that is acceptable for electronic use. Rosin is a natural gum produced within the bark of pine trees. Its properties are interesting in that at normal temperatures rosin is completely inactive and therefore non-corrosive. At elevated temperatures, rosin melts and in this state exhibits a degree of activity sufficient to permit it to act as a flux when soldering many non-ferrous metals, particularly those which have been hot-tin dipped.

For all practical purposes, rosin-type fluxes are non-corrosive and are so described in contemporary technical literature. However, as has been pointed out earlier in this discussion, the very fact that a fluxing action can take place under any circumstances is considered by many engineers as evidence of chemical activity — at least at the time of soldering — and it therefore seems wise to recognize the basic principles involved in the action of a flux and use even rosin-type fluxes with caution and restraint.

Because rosin-type fluxes are lacking in the degree of activity required for easy soldering, their wide acceptance is based solely on their freedom from corrosion. Except in the case of a few easy-to-solder metals, the use of rosin does little to help the soldering process, so it is only natural that intensive research should have been devoted to the soldering problem and to ways of preparing fluxes having a high degree of activity but, at the same time, being without corrosive tendencies. The list of materials and combinations tested is seemingly endless with many fluxes developed which are excellent from the standpoint of soldering but which have failed completely when viewed from the aspect of corrosion.

ACTIVATED FLUXES

It was eventually discovered that rosin could be "activated" in a manner affording greatly improved solderability by the addition of certain ingredients and yet cause little, if any, increase in the corrosion-inducing qualities of the flux rosin. Known as activated rosins, the formulations for these fluxes are, for the most part, carefully guarded, although it is known that aniline hydrochloride is an ingredient of certain activated fluxes.

The chemistry involved in activation of rosin is complex and not particularly important in this discussion. It seems sufficient to say that activated-type fluxes have not yet demonstrated a degree of corrosion resistance generally considered acceptable to a majority of the coil industry. In many instances, activated fluxes have been approved for specific uses, but in the case of components using magnet wires of size No. 38 and smaller, there is an understandable reluctance on the part of most engineers to approve the use of an activated flux, and there is almost universal acceptance of the belief that when this type of flux must be used in electronic applications, the quantity involved should be the absolute minimum consistent with good soldering.

FLUX FORMS

Soldering fluxes are available in paste and liquid forms and also as a core within a solder wire. Paste-type fluxes, other than those used in rosin-cored solders, are generally of the acid type and are therefore unacceptable in electronic applications. Liquid fluxes, particularly water-white rosin dissolved in alcohol, are sometimes used, but by far the greatest amount of the solder consumed by the electronics industry is in the form of flux-cored solder where rosin-type flux, either regular or activated, is introduced as a paste into one or more hollow spaces within the wire solder.

The advantages of enclosing the flux within the solder are many. Among them may be listed the fact that the flux is located where it is wanted — at the point where the solder melts — and that no added operations or materials are involved in the fluxing portion of the soldering process. A further advantage of cored solders is found in the accuracy with which the quantity of flux is controlled. Several different core sizes are available, covering a range of approximately 0.5 per cent by weight up to as much as 6.0 per cent. Since wire cored-solders are made by an extrusion process, a high degree of accuracy is maintained in the size of the core and hence in the quantity of the flux contained therein.

SOLDERING METHODS

Within the field of electronic component and equipment manufacture, solder is generally applied either (1) by the use of a soldering iron or other device for melting the solder at the point of application, or (2) by dipping the article to be soldered into a pot of molten solder. It is probable that at the present time more connections are soldered by the first-named method, but it is recognized as a good possibility that rapidly developing interest in printed circuitry, automation, and other mass-production techniques could, in the very near future, easily swing emphasis to pot-soldering methods.

Regardless of the method used, successful soldering involves two basic operations: (1) tinning of the metal surfaces and (2) filling with solder the space between the tinned surfaces. It is not essential that these operations be performed at the same time. In fact, it frequently will be found an advantage to pre-tin connections well in advance of soldering. An instance where this is true is the previously mentioned case of copper wires from which the insulation has been removed but which are not ready for final soldering. If tinned, such leads may be stored for prolonged periods with no detrimental effect upon their solderability, but if left untinned, in only a few hours time they will acquire an oxide coating too heavy to be dealt with efficiently by a rosin flux.

A point of importance in any form of soldering, and one that is often overlooked in iron soldering is the necessity of having the metals being soldered at the temperature of the molten solder. Unless this condition exists, no alloying action will take place and the solder will merely freeze on the metal, forming a "cold joint" which is of little or no value either electrically or mechanically. Because of this necessity for heating not only the solder but the work as well, care must be taken to use soldering irons capable of transferring the required amount of heat.

The irons most commonly used are electrically heated and may consume anywhere from as little as 20 watts up to 700 or more. Almost without exception the actual soldering tips are made of copper and may be of almost any size and shape depending upon the specific use for which they are intended. It is generally accepted that the temperature of the tip should run about 40 degrees Centigrade above the melting point of the solder being used. For the solders recommended for electronic applications, this means that tips should run between about 230 and 275 C during the actual soldering process.

It will be recalled that one requisite of good soldering is to have the work and the molten solder at the same temperature. In the case of iron soldering all the heat used in the process must emanate from the soldering iron and reach the work by way of the soldering tip. Since the tip should always run about 40 degrees C above the melting point of the solder, it is apparent that the expression "for good soldering, the iron must fit the work" is not only meaningful but is based on a number of considerations. Among the most important factors to be considered in the selection of a soldering iron are the following:

1. The heat capacity of the work to be soldered.
2. The size of the soldering tip — particularly of the area which contacts the work.
3. The presence — or absence — of drafts or strong air currents in the work area.
4. The composition of the solder being used.

Since copper alloys with molten solder, and since soldering tips are effective only when well tinned (working surfaces completely covered with molten solder), it follows that there is an erosive action on the tip which causes them to wear away at a surprisingly rapid rate. Many other materials have been substituted for copper, but none is so completely acceptable. It is, therefore, only natural that many attempts have been made to slow down tip wear. The most effective means seems to be to coat the tip with a thin layer of a metal which does not readily alloy with solder. Iron is an example of a metal with the required properties, and iron-plated soldering tips, while not so effective as plain copper, last enough longer and do a sufficiently good job to make their adoption an economical move.

The practice of holding soldering irons on a fixed stand and bringing the work to the iron rather than the iron to the work is fairly universal in the manufacture of electronic components. The capacity (wattage) of the iron, the shape of the tip, whether or not the tip is plated, the size and the composition of the wire solder, and the size of the rosin core, are all factors in soldering which, as has been indicated above, are directly dependent upon the specific requirements of a particular operation. Proper equipment and techniques will produce soldered joints in conformance with the requirements of paragraph 3.10.3 of MIL-P-11268 (SigC) entitled *Parts, Materials, and Processes used in Signal Corps Equipment; General Specification*, which reads as follows:

> "*Soldering — Soldered connections shall be neat. There shall be no sharp points or rough surfaces resulting from insufficient heating. The solder shall feather out to a thin edge, indicating proper flowing and wetting action, and shall not be crystallized, overheated, or underheated. The minimum necessary amount of flux and solder shall be used for electrical connections. Wherever practicable, excess rosin shall be removed with a wire brush and then a dry cloth; any resulting loose flakes or rosin shall be carefully removed from the inside of the equipment. Insulation material that has been subjected to heating during the soldering operation shall be undamaged and parts fastened thereto shall not have been loosened.*"

POT SOLDERING

Pot soldering or, as it is sometimes called, *dip soldering*, is primarily a tinning operation in which the articles to be soldered, having previously been cleaned and fluxed, are dipped into molten solder. The great advantage of this soldering method lies in the number of connections that may be soldered in a single operation.

The size of the solder pot in relation to the size of the work being soldered determines the temperature at which the pot must operate. It is obvious that as work is lowered into the pot, the temperature of the solder will be lowered. It is, therefore, necessary to maintain a minimum pot temperature of at least 60 to 75 C above the melting point of the solder being used. In the case of large pieces and a high production rate, it will be necessary to maintain an even higher differential — in some instances up to 125 or even 200 C.

The nature of pot-soldering being somewhat different from iron soldering introduces slightly different problems. For one thing, a high grade solder — that is, one with a higher tin content — is usually required. There are at least two reasons why this is true: (1) the use of eutectic solder permits minimum pot temperatures, and (2) the higher the tin content, the faster the speed of wetting and the greater the capillary rise of the solder.

Two major disadvantages are attached to the use of eutectic or other high-tin content solder. The higher the tin content of the solder, the more rapid is its alloying action with copper — hence, the more rapid its contamination, and the greater the impoverishment (loss) of tin during continued operation. So important is this latter point that replacement solder added to maintain pot level must be either of higher tin content than the original solder, or it must be pure tin.*

In view of the rapidly increasing interest in pot-soldering, considerable experimentation relative to optimum compositions of solder and flux are now under way. The simplicity of the process is such that it offers many advantages in the assembly of small electronic components.

OTHER SOLDERING METHODS

This discussion of soldering is meant to be representative and not exhaustive. For that reason, such accepted soldering methods as flame soldering, hot-plate soldering, induction soldering, electrode soldering, and resistance soldering will be enumerated, but because of limited acceptance in the industry, they will not be discussed in more detail.

*While this statement is based on generally accepted theory, it is known that at least one major American company engaged in the assembly of electronic equipment by dip-soldering methods has not found this to be strictly true. As originally set up, this particular process used eutectic solder in pots having a capacity of 30 pounds, enough to process approximately 1500 of the units being soldered. To maintain pot levels, pure tin was originally specified. In practice, however, eutectic solder was added to the pots whose contents upon successive analysis always disclosed a tin content within 0.5 per cent of the specified 63 per cent. This fact, coupled with the progressive increase in copper concentration revealed by the analysis, led to the policy of replenishing the pots with 63/37 solder rather than with pure tin, and to the discard of the contaminated solder at regular intervals in order to maintain uniform soldering conditions.

REQUIREMENTS FOR GOOD SOLDERING

At the start of the discussion of soldering, it was stated that the process of soldering is both common and relatively simple. Such a statement is not meant to imply that soldering is a process which is generally well done, for such, unfortunately, is not always true.

The requirements for good soldering on electronic components might be listed as follows:

1. An iron with a tip of a size and shape such as to permit a maximum transfer of heat from the tip to the work.
2. An iron of a size large enough to thoroughly heat both the solder and the work but not large enough to overheat and "burn" the solder.
3. A flux which is non-corrosive and which is sufficiently active to permit good wetting and easy soldering.
4. A means of applying the flux at the point of soldering — as, for example, the use of flux-cored wire solder.
5. Work, the surfaces of which are relatively clean and free from oxide or other coating.
6. Solder with a tin content of not less than 50 per cent and of a form and size suited to the work at hand.

The importance of good materials, good equipment, and techniques based on an understanding of the requirements of good soldering cannot be overestimated. Soldering is an important operation in the manufacture of electronic components and as such is deserving of intelligent consideration by all who participate in the design or production of such items.

BIBLIOGRAPHY

Barber, Clifford L.
"Soft Solder Fluxes"
Industrial and Engineering Chemistry, October, 1937

Greenfield, L.T. and Forrester, P. G.
Properties of Tin Alloys, Tin Research Institute, Fraser Road, Greenford, Middlesex, England

Hobart, J.F.
Soft Soldering & Brazing
D. Van Nostrand Company, New York, 1919

Hydrean, P.
Some Properties of Soft Solders, Federated Metals Division, American Smelting & Refining Company, 120 Broadway, New York

Lewis, W.R.
Notes on Soldering, Tin Research Institute, Middlesex, England, 1948

McKeown, J.
The Properties of Soft Solders & Soldered Joints, British Non-Ferrous Metals Research Association, Euston Street, London N.W.I., England, 1948

National Bureau of Standards
"Soft Solders," LC-937

Young, James F.
Materials and Processes, Eighth Printing, John Wiley & Sons, Inc., New York, 1949

Tool Steel Handbook
Allegheny Ludlum Steel Corporation
Pittsburgh 22, Pennsylvania

SPECIFICATIONS

Federal Specification QQ-S-571b
"*Solder; Soft (Tin, Tin-Lead, and Lead-Silver)*"

JAN-S-627
"*Solder, Low-Melting Point*"

U.S. Army Specification No. 2-89-A
"*Flux, Soft Soldering*"

U.S. Army Specification 4-103
"*Flux, Soldering, Non-Corrosive*"

CATALOGS AND TECHNICAL INFORMATION OF:

Aircraft-Marine Products, Inc.
2100 Paxton Street
Harrisburg, Pennsylvania

Alden Products Company
117 North Main Street
Brockton 64, Massachusetts

Alpha Metals Inc.
58 Water Street
Jersey City 4, New Jersey

American Brass Company
Waterbury 20, Connecticut

American Phenolic Corporation
1830 South 54th Avenue
Chicago 50, Illinois

American Radio Hardware Company, Inc.
152-4 MacQuestion Parkway South
Mount Vernon, New York

Bead Chain Manufacturing Company
110 Mountain Grove Street
Bridgeport 5, Connecticut

Cambridge Thermionics Corporation
445 Concord Avenue
Cambridge 38, Massachusetts

Grayhill
561 Hillgrove Avenue
La Grange, Illinois

Kester Solder Company
Chicago 39, Illinois

Multicore Sales Corporation
164 Duane Street
New York 13, New York

National Company
61 Sherman Street
Malden, Massachusetts

The Palnut Company
61 Cordier Street
Irvington 11, New Jersey

Shakeproof
Division of Illinois Tool Works
Saint Charles Road
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Tinnerman Products, Inc.
Cleveland, Ohio

United Carr Fastener Corporation
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Section 5
CERAMICS

GENERAL

The nature of high frequency transformers requires that certain parts be fabricated from non-conducting materials possessing good dielectric and mechanical properties. Both plastics and ceramics are used for this purpose with ceramic materials more often specified for those transformers intended for high temperature operation. Unlike plastic materials, ceramics neither soften nor char at high temperatures but remain hard and rigid and are not subject to bending or twisting. Ceramics are excellent insulators possessed of good dielectric constant, low dielectric loss factors, and either low (0.1 per cent) or zero water absorption. They have the further advantage of being chemically inert which, to the design engineer, implies almost complete freedom from electrolytic corrosion. A further advantage which is often of importance in electronic applications is found in the fact that ceramics do not encourage the growth of fungus.

NATURAL CERAMICS

The first ceramics to be used by the electronics industry were actually forms of block talc or natural steatite, mined directly from deposits in the earth. This type of material is readily machineable when green (unfired) and after being formed to the desired shape can be fired to its final hardness with relatively low shrinkage (usually less than 5 per cent). Less vitreous (glass-like) and less uniform than the manufactured steatites, natural ceramics are little used in electronics at the present time.

TECHNICAL CERAMICS

The ceramic most common in electronics applications is *steatite* — a member of the "white ware" group of stone-like materials including porcelain and china. Vitreous in nature and possessing extremely low water absorption, ranging from 0 to 0.02 per cent, steatite is available in a number of different mixes, each varying slightly in certain of its characteristics but all of high electrical quality.

A group of ceramics known as *refractories* were, at the time of World War II, highly porous materials, some of which would absorb water in amounts as high as 18 per cent by weight. Included in this group of refractories are zircon, alumina, and cordierite. All of these materials possessed properties which showed considerable promise for electrical applications if only the moisture resistance could be improved. This was especially true of cordierite whose extremely low coefficient of linear expansion — 1.6×10^{-6} in the range of 25 to 100 C — made it an ideal material upon which to wind extremely stable coils.

Because it was recognized by Signal Corps engineers that water absorption in ceramic material should be as close to zero as is possible and in no case should exceed 0.1 per cent, a program of research and development intended to lead to an improvement in ceramic materials of the nature of cordierite and zircon was initiated by the Signal Corps Engineering Laboratories shortly after the close of the War. The success of this program is attested to by the fact that specification JAN-I-10 which had previously listed as approved ceramic materials only steatite, porcelain, glass, and glass-bonded mica was amended in January 1953 to include as additional approved materials cordierite, zircon, wollastonite, forsterite, alumina, and lithia porcelain. Since that time more than one hundred commercial ceramic bodies have been approved under this specification. (See Fig. 5-1)

TEMPERATURE EFFECTS

Many manufactured (technical) ceramics are sintered and vitrified at very high temperatures, usually somewhere between 2000 and 3500 F.

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Fig. 5-1 Property chart of JAN-I-10 ceramics. (From Signal Corps Engineering Laboratories Information Bulletin No. 201)

JAN-I-10 CERAMICS		L, GRADE & TYPE (NOTES 1, 2, & 3)	POWER FACTOR, 1MC	DIELECTRIC CONSTANT 1MC	LOSS FACTOR 1MC (NOTE 3)	DIELECTRIC STRENGTH, V/MIL (NOTE 4)	FLEXURAL STRENGTH, PSI (NOTE 5)	POROSITY (FUCHSINE DYE PENETRATION)
STEATITE	UNGLAZED	L-3A	0.0031-0.0035	5.34-5.9	0.018-0.020	211-335	17,220-18,000	NONE
		L-4A	0.0018-0.0031	5.25-5.88	0.011-0.018	202-224	17,600-21,200	NONE
		L-5A	0.00035-0.0010	6.42-6.03	0.0021-0.0063	210-230	12,300-20,500	NONE
	GLAZED	L-3A	0.0018	5.90	0.022	>180	23,400	NONE
		L-4A	0.0015-0.0026	5.62-6.72	0.009-0.018	198-217	16,700-27,200	NONE
		L-5A	0.0008-0.0014	5.73-6.18	0.005-0.008	210-330	22,000-28,600	NONE
PORCELAIN	GLAZED-UNGLAZED	L-2A	0.0065-0.0107	5.9-6.0	0.038-0.063	150-310	8,000-12,300	NONE
		L-6A	0.0075-0.0090	5.42-6.01	0.041-0.059	196-240	8,500-13,600	NONE
ZIRCON	UNGLAZED	L-3A	0.00801	9.03	0.0101	200	23,100	NONE
		L-4A	0.0012-0.0014	8.40-8.27	0.010-0.012	181-250	20,100-22,900	NONE
		L-5A	0.0008	7.40	0.006	168	14,600	NONE
	[illegible]	L-4A	0.00110	8.98	0.01	181	29,000	NONE
ALUMINA	[illegible]	L-6A	0.0006-0.0008	8.14-9.20	0.006	180-224	24,000-42,200	NONE
GLASS	[illegible]	L-3A	0.0051	4.73	0.024	251	9,400	NONE
		L-4A	0.0030-0.0020	6.62-6.72	0.009-0.011	215-255	8,400-12,000	NONE
		L-6A	0.00074	6.19	0.0031	220	16,000	NONE
	[illegible]	L-8	0.0017	3.78	0.0068	360	8,520	NONE
[illegible]	GLAZED	L-3A	0.0027	7.30	0.020	300	13,400	NONE
		L-4A	0.0017-0.0018	7.08-7.44	0.012-0.013	228-305	16,000-17,300	NONE
[illegible]	[illegible]	L-5B	0.0006	5.9	0.003	181	15,700	NONE
	[illegible]	L-6B	0.00051	6.37	0.0015	195	19,330	NONE
[illegible]	[illegible]	L-3A	0.0048	4.57	0.022	240	13,000	NONE
[illegible]	[illegible]	L-6A	0.0006	6.49	0.003	288	15,700	NONE

JAN-I-10 GRADE REQUIREMENTS
1. TYPE A-B THERMAL ENDURANCE CYCLES FROM 100°C INTO 10°C WATER
2. TYPE B-B THERMAL ENDURANCE CYCLES FROM 100°C INTO 10°C WATER
3. GRADE L-1 L-2 L-3 L-4 L-5 L-6
LOSS FACTOR 0.150 0.070 0.035 0.016 0.008 0.004
4. DIELECTRIC STRENGTH 150 VOLTS PER MIL
5. FLEXURAL STRENGTH 5000 PSI

Nearly all technical ceramics will withstand prolonged heating at 1830 F since they can be used at temperatures almost up to their firing temperature. Some grades of technical ceramics are satisfactory for continuous operation well above this figure, while others such as glass-bonded mica require lower safe-operating temperatures.

Of particular importance to the electronics engineer are the widely differing coefficients of linear expansion to be found among ceramics. Information available from one American manufacturer[1] shows his ceramic materials to have coefficients of linear expansion within the temperature range of 25 to 100 C, varying from a high of 11.3 x 10^{-6} in the case of a vitreous ceramic (steatite) to a low of 1.6 x 10^{-6} in the case of a refractory (cordierite). Another manufacturer[2] produces a lithia-porcelain material with a coefficient of linear expansion of 0.85 x 10^{-6}, a power factor of 0.00424, and very low moisture absorption. Similar materials with zero and with negative coefficients of expansion as great as 6.5 x 10^{-6} are also produced by this same manufacturer. It is largely due to the varying degrees of thermal expansion available in ceramic materials and to the fact that this thermal expansion can be matched to that of various metals including Invar that metal-ceramic seals have been made possible.

Reference to Fig. 5-3 will show the effect of various coil form materials upon the temperature coefficient of universal coils. The superiority of ceramic forms is clearly revealed in this graph where the three types of ceramics tested constituted a group exhibiting maximum temperature stability. While exact values for the thermal coefficient of linear expansion of all coil form materials tested in the course of this program are not available, it is apparent that there is a direct relationship existing between the temperature coefficient of the coils and the thermal coefficient of linear expansion of the forms on which they are wound. This graph is offered as evidence of the value of proper selection of material when maximum coil stability is a design requisite.

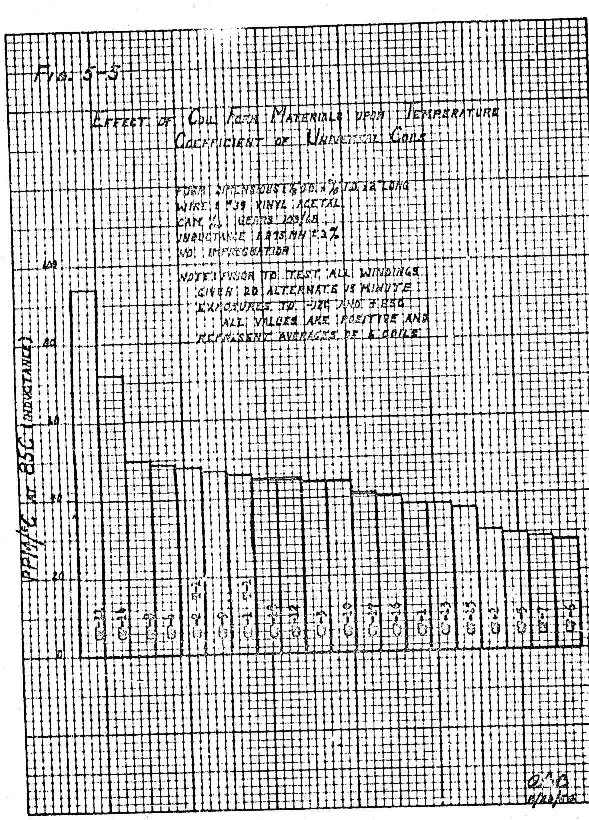
Fig. 5-2 COEFFICIENTS OF LINEAR EXPANSION OF COMMON CERAMIC MATERIALS

Material	Coefficient of Linear Expansion* 25-100 C	25-700 C
Steatite Type L-4A	7.3 x 10^{-6}	8.6 x 10^{-6}
Steatite Type L-5A	6.4 x 10^{-6}	8.9 x 10^{-6}
Forsterite	9.1 x 10^{-6}	10.6 x 10^{-6}
Titanium Dioxide	7.3 x 10^{-6}	8.7 x 10^{-6}
Zircon	3.2 x 10^{-6}	4.1 x 10^{-6}
Cordierite	1.6 x 10^{-6}	2.8 x 10^{-6}
Wollastonite	—	6.1 x 10^{-6}
Alumina	5.1 x 10^{-6}	7.2 x 10^{-6}
Lithia Porcelain	—	0.085 x 10^{-6}

*Expressed in change per unit length per degree centigrade.

[1] American Lava Corporation, Chattanooga 5, Tennessee.

[2] Stupalith manufactured by Stupakoff Ceramic and Manufacturing Company, Latrobe, Pennsylvania.



GLASS

While not a particularly common material for use in high frequency transformers, *glass* does have certain interesting properties and is approved for use in military applications under JAN-I-10.

Glass is an amorphous (noncrystalline) material which is rigid at ordinary temperatures and which softens or even becomes fluid at elevated temperatures and is usually without a definite melting point in between the extremes.

Glass is available in an almost limitless number of formulations, nearly all of which have sand or silica (SiO_2) as the principal ingredient. Other ingredients such as soda, lime, magnesia, alumina, boron oxide, lead oxide, and potash are added to lend particular physical properties to the glass as well as to lower the melting point of the mixture, thereby making fabrication easier.

From the standpoint of electronics design, there is little interest in glasses other than in the high silica[3] variety since only in this type of glass is the power factor sufficiently low to permit its consideration in high frequency applications. Power factors ranging from 0.0001 to 0.005 measured at 1 Mc at 20 C are given as applying to this type of glass as supplied by one major glass manufacturer.[4]

Considerable work has been done with metallized glass inductors, and it is entirely possible that as development work along these lines progresses, the use of this type of coil may become far more common. In general, coil designers will seldom incorporate glass in their high frequency transformers other than in hermetically sealed units where glass-to-metal seals may be used.

GLASS-BONDED MICA

Glass-bonded mica is not a new material. It made its first commercial appearance in 1921 and is currently available under a number of different trade names.[5] In many ways the nature of glass-bonded mica resembles a plastic more than it does a ceramic. However, the fact that it is essentially an inorganic material and that it is capable of operation at temperatures as high as 600 F accounts for its consideration in electronics as a high temperature insulating material -- hence its inclusion with ceramics.

As the name suggests, glass-bonded mica is made up of ground mica particles held together by an easily melted form of glass. Until very recently, the mica used in this material was always of the natural variety. Recent developments in synthetic mica have brought about the use of this new type of material and have made possible the production of a glass-bonded synthetic mica capable of operation at temperatures up to 900 F, whereas with natural mica, 650 F was the top operating temperature.

When this mixture of mica and glass is heated to the melting point of the glass, the material may be formed to shape in steel molds where cooling and subsequent hardening take place. The nature of glass-bonded mica makes it adaptable to both compression and injection molding techniques. Inserts may be molded in and may be of any metal capable of withstanding temperatures of 1300 F under pressures up to as much as 40,000 pounds per square inch. Brass, Monel, and cold-rolled steel are commonly used as material for inserts. There should always be at least 1/16 inch of glass-bonded mica between inserts which themselves should never be located less than 1/16 inch from the edge of the molded piece. While it has been stated that both compression and injection molding techniques are used in the fabrication of glass-bonded mica, it must be recognized that the necessity for preheating to 1300 F and the maintenance of mold temperatures of 800 F necessitate a type of mold and molding equipment vastly different from that used in conventional plastics molding.

In the design of parts to be fabricated from glass-bonded mica, the same general principles apply that pertain to comparable parts to be molded from plastics. A good plastic flow is always desirable, which means that rounded corners and gradually changing contours should be incorporated in the piece being designed so that the cavity may fill in the shortest possible time. The need for wall thicknesses of 1/16 inch or greater, for taper on all vertical surfaces, for generous radii (minimum 1/64 inch, preferred 1/32 inch) on all edges and corners must be recognized as essential to good design practice.

Glass-bonded mica, like steatite, is well adapt-

[3] An example of this type of glass is "Vycor" manufactured by Corning Glass Works, Corning, New York.

[4] Corning Glass Works.

[5] "Mycalex" manufactured by Mycalex Corporation of America, Clifton, New Jersey.

"GE-Mycalex" manufactured by General Electric Company, Chemical Division, Pittsfield, Massachusetts.

"Mykroy" manufactured by Electronic Mechanics, Inc., Clifton, New Jersey.

"Insamol" manufactured by Westinghouse Electric Corporation, Philadelphia, Pennsylvania.

POOR ORIGINAL

ed for use at elevated temperatures. Possessed of a relatively low loss factor, low dielectric constant, high dielectric strength, and water absorption equal to approximately zero, this material has one distinct advantage over steatite in that it can be machined on ordinary machine shop equipment. Tungsten carbide tools give best results. A copious supply of water serving both as a coolant and as a lubricant is essential to the machining process.

Another advantage of glass-bonded mica over ceramic material is to be found in the accuracy with which this material can be molded. The normal tolerance applied to molded glass-bonded mica pieces of a length or width up to 1 inch is ±0.002 inches. Thickness (that dimension which is at right angles to the parting line of the mold) is ordinarily held to ±0.005 inches. Since glass-bonded mica contains neither fillers nor plasticizers, molding consists only of confining the molten material for a period of time sufficient to permit the material to become rigid through loss of heat.

Glass-bonded mica is a material with which every engineer engaged in the design of high frequency transformers should be familiar. Present tendencies toward higher ambient temperatures and miniaturization of components cannot but direct attention toward this material in view of its excellent mechanical and electrical properties.

MATERIAL CHARACTERISTICS

Were it not for their cost, technical ceramics would have a much wider use in the electronics industry. They can be provided with holes either tapped or plain. Some are available glazed[6] or unglazed, although those with low coefficients of expansion are difficult to glaze without having crazing[7] appear at elevated temperatures. Recent developments in surface glazes have substantially lessened the dangers of crazing, and it appears probable that it will soon be possible to glaze all types of ceramic materials without danger of crazing. It is possible to coat ceramics with metal, thereby providing a surface which may be readily soldered. Because ceramics are available in so many expansion rates, it has been possible to develop ceramic-to-glass, ceramic-to-metal, and glass-to-metal seals capable of operation over extended temperature ranges without danger of leaks due to cracking or crazing. Sizes and shapes in which ceramics may be obtained are almost limitless since the pieces may be formed by dry pressing, wet pressing, extrusion, or casting. They may also be machined before firing from pieces of either pressed or extruded material. After firing, the material is too hard to permit machining, but grinding or lapping or honing to accurate dimensions is possible. However these operations are expensive, and good design practice will call for as little post-firing work as will give the desired dimensions in the final piece.

Because of the nature of ceramic materials, there is a substantial amount of shrinkage in size between the green state and the final fired state. This shrinkage varies with the particular composition of the material but may be considered as approximating 15 to 20 per cent for the average electrical grade of ceramics. Development work now in process will undoubtedly result in a substantial reduction in this figure. While fairly predictable, shrinkage is at times difficult to control. Frequently the length and thickness of a piece will show different shrinkages — probably because of a difference in pressures incurred in molding. Ceramics, therefore, require greater dimensional tolerances than would otherwise be the case, with the standard tolerance for ceramic parts being ±1 per cent with nothing less than ±0.005 inches over unglazed surfaces, and ±2 per cent with nothing less than ±0.010 inches over glazed surfaces. To maintain closer limits means either selective gauging or post-firing operations — both of which are expensive — but where cost is no object, ceramic pieces can be delivered within limits of ±0.0005 inches or even less.

USES IN ELECTRONICS

In the design of high frequency transformers, one of the principal uses of ceramic material is in coil forms where operation at elevated temperatures or exceptional temperature stability is a basic requirement. Ceramic coil forms may be either tubes or solid dowels. They may have smooth surfaces, or they may be threaded for use with solenoid windings. In the case of tubes, the thinnest wall thickness obtainable in commercial practice is 3/32 inches in a tube with an ID of 5/16 inches. Ellipticity is a serious problem with thin walled ceramic tubes. According to General Standards of Steatites and Other Electrical Grades of Ceramics:

> *"Ellipticity shall be determined by dividing the maximum diameter by the minimum diameter measured in the same plane perpendicular to the axis. For wall thicknesses of 7 to 10 per cent of the outside diameter, this quotient shall not exceed 1.03 and for wall thicknesses of 10 per cent or more of the outside diameter, this quotient shall not exceed 1.02".*

Because of this problem with ellipticity, designs requiring absolutely round ceramic tubes — for example, permeability tuned transformers — should be approached with caution since such tubes can be obtained only by selective gauging or by grinding.

Other uses of ceramics in high frequency transformer design may involve terminal boards, bushings, capacitor bases, spacers, resistor cores, and other applications where dielectric strength, low dielectric constant, low loss factor, excellent moisture resistance, freedom from cold flow, and mechanical stability at elevated temperatures are requisites of the design.

While vitreous ceramics generally have excellent moisture resistance, for those cases where maximum resistance to the effects of surface moisture is required, ceramics may be given a surface treatment with a *silicone solution*. For maximum effectiveness, this treatment — which consists of dipping the hot ceramic parts in a dilute solution of a silicone[8] in a volatile material such as carbon tetrachloride or trichloroethylene followed by baking at 160 C for one hour or 300 C for one half hour to remove the solvent and set the silicone film — should be done immediately after the pieces are removed from the kiln before they have become contaminated through handling or exposure to atmosphere.

A wide variety of sizes and shapes of ceramic material is available from various manufacturers who are tooled on these items. Familiarity with these standard parts is essential to economical design practice, since their use wherever possible will help to avoid tool costs and production delays. Because of the constant changes which are the natural result of customer requirements, it will be found most helpful to an engineer to maintain close contact with the ceramics manufacturers. By so doing, he can be assured of up-to-date information on both the properties of the various ceramic materials and the standard parts which are immediately available.

DESIGN PRINCIPLES

Since standard parts often will not suffice, a coil engineer should be sufficiently familiar with ceramics to enable him to design parts which will be both practical and economical. In general, the design principles which apply to plastics apply also to ceramics since both are molded materials. Ceramics, however, require heavier wall sections; flat plates to be free from warpage should never be designed less than 1/8 inches thick with 1/4 inches even more desirable. With the exception of glass-bonded mica, metal inserts cannot be molded in most technical ceramics because of the high firing temperatures. Simple shapes adapted to compression molding or to extrusion will be found most economical and most satisfactory.

Wherever possible, ceramic parts should be ordered in accordance with the industry-accepted tolerances quoted earlier in this chapter. Not only will tightening of these tolerances increase the cost of the parts, but it is likely to result in much more difficult procurement with subsequent production delays. It is hoped that the chart appearing as Fig. 5-5, entitled Good Design Practices for Ceramic Radio Insulators, will be found helpful when ceramic parts must be designed.

[6] Glazing is an operation performed primarily to make the surface of a ceramic easy to clean. It does not contribute to lower moisture absorption, as all materials approved under JAN-I-10 must be satisfactory in this respect before glazing.

[7] Tiny surface cracks.

[8] Such as Dow Corning 200 Fluid — 350 centistokes.

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Fig. 5-4 RECOMMENDED PROCEDURE FOR SILICONE TREATMENT OF CERAMIC PARTS

The application of a silicone treatment consists of three simple steps:

1. Cleaning
2. Solvent application
3. Curing

1. *Cleaning*

For optimum results, the surface of the articles to be treated must be dry and free of absorbed grease and electrolytes. Wherever possible the silicone treatment is preferably applied to new, unhandled ceramic parts.

To clean articles that have been used, two methods of cleaning are suitable:

a. Heat to above 400° C for 1 hour or more
b. Degrease by means of a solvent degreasing operation

In either case, if electrolytes might be present on the surface, the article should be immersed in boiling distilled water for 30 minutes and dried prior to heat cleaning or degreasing.

2. *Solvent application*

Only a very thin film of silicone is required and excess should be avoided.

Prepare a 2 per cent solution by weight of Dow Corning 200 Fluid in methylene chloride, carbon tetrachloride, trichlorethylene or perchlorethylene. It is preferable that this solution be prepared fresh. A solution which has stood for several weeks may become less effective.

Dip the article to be treated into this solvent solution; remove and drain; air dry or heat for one-half hour at 100° C to allow the solvent to evaporate.

3. *Curing*

After solvent evaporation, the silicone film must be completely cured — an operation which is easily accomplished by heating for 1 hour at 300° C (572° F) or two hours at 275° C (527° F).

BIBLIOGRAPHY

Alexy, M. Raymond
"Design Considerations in Miniature High-Voltage Airborne Components"
Electrical Manufacturing, May, 1953

Du Bois, J.H.
"Glass Bonded Mica"
Society of Plastics Engineers Journal, April, 1953

Plastics
American Technical Society, Publishers, Chicago, 1947

Hodgman, Charles D., and others
Handbook of Chemistry and Physics, Thirty-fifth edition
Chemical Rubber Publishing Company, Cleveland, Ohio, 1953

Young, James F.
Materials and Processes, Eighth Printing
John Wiley and Sons, Inc., New York, 1949

CATALOGS AND TECHNICAL INFORMATION OF

American Lava Corporation
Cherokee Boulevard and Manufacturers Road
Chattanooga 5, Tennessee

Cambridge Thermionic Corporation
440 Concord Avenue
Cambridge 38, Massachusetts

Corning Glass Works
Corning, New York

General Ceramics and Steatite Corporation
Crows Mill Road
Keasbey, New Jersey

Mycalex Corporation of America
125 Clifton Boulevard
Clifton, New Jersey

National Ceramic Company
500 Southard Street
Trenton 2, New Jersey

National Company, Inc.
61 Sherman Street
Walden, Massachusetts

Stupakoff Ceramic and Manufacturing Company
Box 390 Hillview Avenue
Latrobe, Pennsylvania

SPECIFICATIONS

JAN-I-7
"Insulators, Glass-Bonded-Mica, Radio

JAN-I-8(1)
"Insulators, Steatite, Radio"

JAN-I-9(1)
"Insulators, Glass, Radio

JAN-I-10(3)
"Insulating Materials, Ceramic Radio, Class L."

JAN-I-21(1)
"Insulators, Porcelain, Radio"

QPL-10-8
"Insulating Materials Ceramic Radio, Class L." (Specification JAN-I-10)

HH-I-536
"Insulation Sheet, Electrical, Natural Muscovite Mica"

7220
"Insulating Materials and Parts; Cleaning of"

General Standards for Steatites and Other Electrical Grade Ceramics as adopted by The Steatite Research Council.

Fig. 5-5

GOOD DESIGN PRACTICES for CERAMIC RADIO INSULATORS*

ITEM IN STANDARD SPEC.	STEATITE SPEC. JAN-I-8	GLASS SPEC. JAN-I-21	GLASS BONDED MICA SPEC. JAN-I-7
Overall Dimensions	No requirements.	No requirements.	Dimensions of 0" to 15": ±1/64" Dimensions of 15" up: ±1/64" length, thickness, ±0.020" width
Critical Dimensions	Unglazed Surfaces: ±1% or ±0.005", whichever is greater. Glazed Surfaces: ±2% or ±0.012", whichever is greater. Minimum thickness tolerance: ±0.010" unless otherwise specified.	±1% or ±0.010", whichever is greater.	No requirement.
Non-Critical Dimensions	Dimensions of 0" to 12-1/2": ±2% up to ±1/8". Dimensions of 12-1/2": ±1%. Minimum tolerance: ±1/64".	Dimensions of 0" to 12-1/2": ±2% up to ±1/8". Dimensions of 12-1/2" up: ±1%. Minimum tolerance: ±1/64".	No requirement.
Cylindrical Shapes	OD and ID Unglazed: ±1% or ±0.005", whichever is greater. OD and ID Glazed: Diameters of 0" to 1" -- ±2% or ±0.012", whichever is greater. Diameter of 1" up: ±(1% + 0.010"). Minimum wall thickness ±0.010" unless otherwise specified.	Diameters up to and including 1": ±2% or ±0.010", whichever is larger. Diameters over 1": ±(1% + 0.010"). Minimum wall thickness ±0.010"), unless otherwise specified.	No requirement.
Camber			No requirement.
Wall Thickness of Tubes	Not less than 8% of OD nor more than 20% or 7/8", whichever is smaller. If OD x 3 < L < OD x 6, 20% OD practicable. If OD x 6 < L, 10% OD practicable. Minimum wall thickness: 1/32". Minimum diameter recommended: 0.100".	Not less than 8% of OD nor more than 20% or 7/8", whichever is smaller. If OD x 3 < L < OD x 6, 20% OD practicable. If OD x 6 < L, 10% OD practicable. Minimum wall thickness: 1/32". Minimum diameter recommended: 0.100".	No requirement.
Ellipticity of Tubes	For walls less than 10% of OD: $\frac{\text{Maximum Diameter}}{\text{Minimum Diameter}}$ must be less than 1.03. For walls more than 10% of OD: $\frac{\text{Maximum Diameter}}{\text{Minimum Diameter}}$ must be less than 1.02.	For walls less than 10% of OD: $\frac{\text{Maximum Diameter}}{\text{Minimum Diameter}}$ must be less than 1.03. For walls more than 10% of OD: $\frac{\text{Maximum Diameter}}{\text{Minimum Diameter}}$ must be less than 1.02.	No requirement.

(continued)

Part I MATERIALS OF CONSTRUCTION

Hole Centers	±1% or ±0.005", whichever is greater. Holes perpendicular to axis or ends ±2°	±1% or ±0.005", whichever is greater.	Edge of hole not less than 1/8" from edge of piece. Distance between edges of holes not less than 1/8". Tolerance ±0.007".
Hole Diameters	Glazed: ±2% or ±0.012", whichever is greater. Unglazed: ±1% or ±0.005", whichever is greater.	±1% or ±0.005", whichever is greater.	±0.005" for diameter up to 0.5". ±1% for diameter over 0.5".
Tapped Holes	Avoid if possible. If necessary, 6-32 or larger, countersunk less than one thread depth. Length 0" to 1.5": Minimum 6 threads. Length more than 1.5": Minimum 9 threads. Add 3 threads beyond number required for screw.	Avoid if possible. If necessary, 6/32 or larger, slightly countersunk. Minimum: 6 threads. Maximum: 12 threads. Add 3 threads beyond number required for screw.	Avoid if possible. Threads shall conform to Class 1 fit and shall not be smaller than 6-32. Avoid blind holes. Countersink slightly to prevent chipping.
Wire Groove Threads	90° considered standard for "V" threads. ±3° on thread angles. Slight radius at bottom of all grooves.	90° considered standard for "V" threads. ±3° on thread angles. Slight radius at bottom of all grooves.	No requirement.
Edges	Specify slight radius or bevel.	Specify slight radius or bevel.	±1° for angle or V cuts in edge of flat pieces. Radius of not less than 1/8" at apex of angle or "V" cut.
Thickness for Flat Shapes	Where ratio of maximum superficial dimension to largest dimension at right angles is less than 5, minimum thickness in inches $\sqrt{A}/10$. Where ratio is 5 or more, minimum thickness in inches $\sqrt{A}/8$. A = outline area.	Where ratio of maximum superficial dimension to largest dimension at right angles is less than 5, minimum thickness in inches $\sqrt{A}/10$. Where ratio is 5 or more, minimum thickness in inches $\sqrt{A}/8$. A = outline area.	Flat pieces available in standard thicknesses of 1/8", 3/16", 1/4", 3/8", 1/2", 5/8", 3/4", 7/8", and 1". For maximum dimension 0" to 3", minimum thickness 3/32". For maximum dimension above 3", minimum thickness 1/8".
Fastening of Parts	Cushion with resilient material. Suspend at 3 points or less.	Cushion with resilient material. Suspend at 3 points or less.	No requirement.
Parallel Limits	Satisfactory if thickness limits are met.	Satisfactory if thickness limits are met.	Flatness limits should be 0.0015" per inch of length.
Diameter of Rods	No requirement.	±(0.035" x OD + 0.015") when OD <1.25".	Available 1/4", 3/8", 1/2", 5/8", 3/4", 3/8", and 1". Tolerance ±0.005".

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ACKNOWLEDGEMENT

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Mr. Sam DiVita
Signal Corps Engineering Laboratory
Fort Monmouth, New Jersey

Section 6

PLASTICS

Plastics are synthetic organic materials. They derive their name from their most important property — the ease with which they can be molded. The plastics industry probably has its beginning with the initial production of celluloid in the year 1870. However, it was not until Dr. Leo H. Baekeland announced his discovery of phenolformaldehyde resin in 1909 that the plastics industry really began the growth which has carried it to a point where its sales volume approximates three billion dollars per year, making it the sixth largest industry in the United States.

THERMOPLASTICS

Plastic materials can be separated into two major classifications based upon their behavior when subjected to heat. First, there are the *thermoplastics* which soften under heat and harden again when cool. This change of state is not accompanied by any chemical change and may be repeated a number of times without altering the properties of the material. More than a dozen basic plastic resins fall into this class, and among them are materials with softening points ranging from approximately 140 F to more than 250 F. Thermoplastic materials exist within a wide range of mechanical and electrical properties as well as of softening temperatures. Some have a sharp softening point; others become progressively softer as the temperature rises. From the standpoint of an engineer engaged in the design of electronic components, the most serious drawback of the thermoplastic group of materials is their inability to withstand elevated temperatures.

THERMOSETTING MATERIALS

The second major classification of plastics is the group of *thermosetting* materials. These soften only once under heat, whereupon they undergo a chemical change which leaves them hardened and impervious to further applications of heat up to the charring point. Actually, with the exception of the relatively new alkyds and silicones, there are but three of these plastics in general use — *phenol-formaldehyde, urea-formaldehyde, and malamine-formaldehyde* or, as they are more commonly called, *phenolics, ureas,* and *melamines.* Since these materials do not tend to soften under heat, they are widely used in applications where operating temperatures will range between 150 to 300 F.

FILLERS

Thermosetting resins are rarely used without the addition of fillers. Among the various materials employed for this purpose are wood flour, cotton flock, glass filaments, powdered mica, and asbestos. The percentage of filler to resin varies widely according to the particular resin and filler selected and to the purpose for which the filler is added but may, in certain instances, go as high as 50 per cent by weight.

There are a number of reasons for the use of fillers. Not the least of these is the reduction in cost of the molded pieces resulting from the fact that most fillers are far less expensive than the resins with which they are used. Also of importance is the increased production obtainable from each cavity as a result of the shortened curing cycle made possible by the addition of fillers to the base resin. Another factor which is influenced by the use of fillers is shrinkage of the molded parts during the curing cycle. All plastic materials shrink somewhat during molding — a condition resulting in such effects as the presence of "shrink marks", warpage, and dimensional instability, all of which are lessened by the addition of fillers to the base resin.

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Part I MATERIALS OF CONSTRUCTION

For applications in the field of electronics, the filler of greatest importance is powdered mica. Use of this material substantially improves the electrical properties of almost every thermosetting resin since the electrical losses originating in the mica are far less than those which are inherent in the plastic. In addition to improving the electrical properties of the molded parts, the use of powdered mica also lessens the tendency of the plastic to absorb moisture, while at the same time it increases the dimensional stability of the molded pieces. A disadvantage of mica-filled materials is found in their lowered mechanical strength resulting from failure of the resin to "wet" the mica during the mixing process. Then, too, if it is necessary to perform any machining on the finished pieces, the addition of mica makes these operations much more difficult.

Wood flour is often used to improve the physical appearance of molded parts, but its use in radio frequency components is not generally recommended because of its poor power factor and its tendency to invite corrosion under conditions of high humidity.[1] For applications where impact resistance is of prime importance, cloth fillers are used to produce parts with up to 15 times the shock resistance of similar pieces made from the same resin filled with wood flour. Asbestos is sometimes used as a filler where maximum resistance to elevated temperatures and/or unusually great dimensional stability are required, but because of poor electrical characteristics, it should not be considered for any part which will be located within a radio frequency field.

Thermoplastic materials are, for the most part, used in pure resin form without the addition of fillers. This fact is especially true of the acrylic group, while on the other hand polystyrenes are sometimes used with fillers in amount ranging up to 40 per cent by volume.

MOLDING METHODS

Molding is the most important means of fabricating plastic materials. Three methods are in common use, the best known being *compression molding*. As the name suggests, in this method the plastic material is formed under pressure and heat to its final shape.[2] The simplest molds are those in which the plastic is placed in a cavity either as a *preform* or as a measured amount of powder. (A preform consists of a definite amount of plastic which has not been heated but merely pressed sufficiently to make it hold together in a form which will fit in the cavity of the mold).

With the plastic in the cavity, the mold is placed in a heated press and closed under pressures sometimes reaching as high as 5,000 pounds per square inch. Under heat and pressure the resin is cured at approximately 320 to 325 F for a period of time ranging from a matter of seconds to an hour or more, depending upon the material, its quantity, and the nature of the finished part. During the cure, a change takes place in the resin leaving it hard and in exact conformance with the shape of the mold cavity. When the cure is complete, the mold is opened and the piece is removed.

Compression molding is the generally accepted method for handling thermosetting materials. It can be and sometimes is used for thermoplastic materials in which case provision must be made for cooling the mold below the hardening point of the plastic; otherwise it would be impossible to remove the finished piece. Since this is neither a simple nor a practical method of handling thermoplastic materials, a more commonly used method is that known as *injection molding*.

This process was developed to eliminate alternately heating and cooling the molds when working with thermoplastic materials. As the name suggests, the semiliquid thermoplastic is actually "injected" into the mold cavity under considerable pressure. The cavity is cool when the material enters, and the mold remains closed only long enough for the thermoplastic to cool below its hardening point. The mold can then be opened, the part removed, and the cycle repeated. Injection molding is much more rapid than compression molding, and sometimes as many as 5 "shots" per minute are possible. Since production molds usually contain a number of cavities — more than a hundred in some cases — injection molding lends itself to high production rates and low costs.

For many years the economies of thermoplastic injection molding were denied users of thermo-

[1] See discussion of electrolytic corrosion in Section 8 of this manual.

[2] In certain specific instances as, for example, in the manufacture of powdered iron cores and resin-bonded ceramics, the articles are first molded to shape in cold molds from which they are removed and subsequently cured in a separate baking operation performed outside of the mold. While in reality a form of compression molding, in the plastics industry this procedure is known as cold molding.

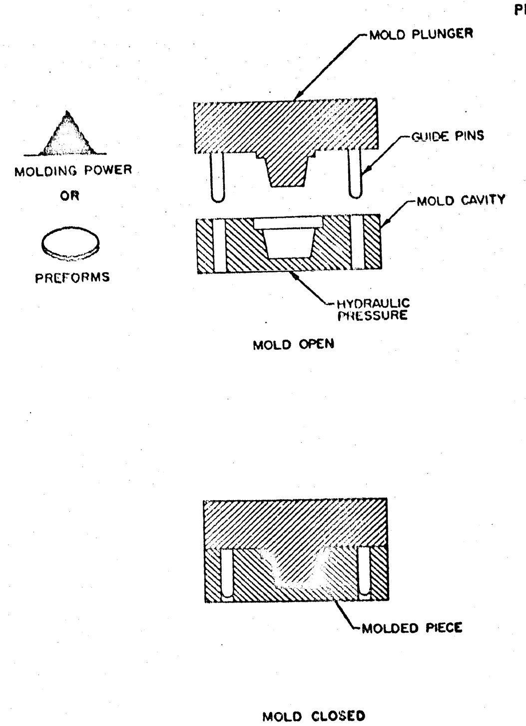
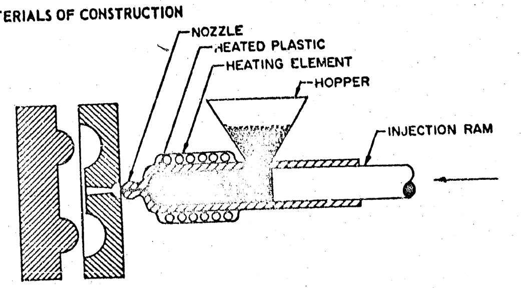
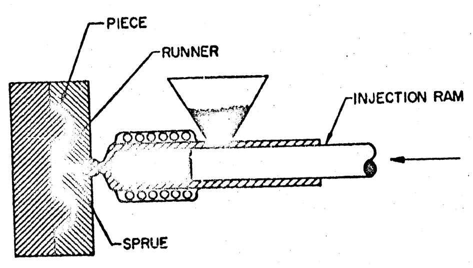


Fig. 6-1 Basic principles of compression molding.

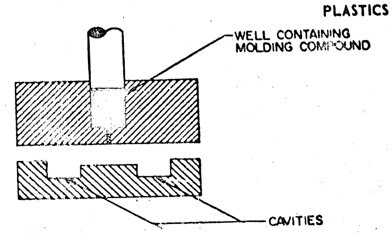


MOLD OPEN

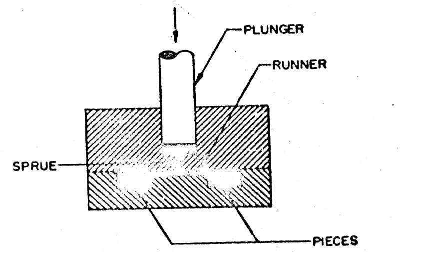


MOLD CLOSED

Fig. 6-2 Basic principles of injection molding.



MOLD OPEN



MOLD CLOSED

Fig. 6-3 Basic principles of transfer molding.

setting materials because the nature of these plastics prevented their use in injection molding. To make thermosetting materials more competitive with thermoplastics, a new molding process known as *transfer molding* was developed and later patented. [3]

Transfer molding allows thermosetting materials to be handled with approximately the same techniques as those used in injection molding. A transfer mold contains a small cylinder or well in which the thermosetting molding material is heated to the flowing point but not to the curing point. The fluid plastic is then injected into the cavity where it remains under heat and pressure until cured. This process is particularly advantageous when small inserts must be molded into the finished piece. Mold design is somewhat complicated, but the advantages of transfer molding are undeniable.

because of uniform cross section can be extruded. The process can also be extended to the point of extruding plastic materials directly on wire for insulating purposes. Dimensional tolerances are necessarily greater on extruded parts than on molded parts but will be found satisfactory for many applications.

LAMINATES

Another process widely used in the fabrication of sheets, rods, and tubes, is known as *laminating*. A *laminated material* is formed from layers of individual filler sheets bonded together by resins into a solid body. Phenolic tubes and sheets so widely used in the electronics industry for coil forms and terminal boards are examples of laminated plastics. Paper, cloth, asbestos, glass fabric, and other similar materials may be used with any thermosetting plastic to prepare a laminate.

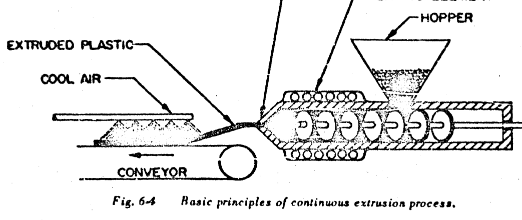


Fig. 6-4 Basic principles of continuous extrusion process.

EXTRUSION METHODS

In addition to being moldable, some plastic materials may be forced through a die to produce a strip of material shaped like the die opening. Only thermoplastic materials can be extruded, since no opportunity is afforded for the cure of a thermosetting plastic.

In general, continuous extrusion is a comparatively simple process requiring neither expensive machinery nor complicated dies; and although not without its problems, it is a very fast and satisfactory method of making ribbons, rods, bars, tubing, and other forms of plastic materials which

Laminates are made in many different grades.[4] Their fabrication begins when the filler sheets are impregnated with a varnish solution of the basic resin. This impregnation is accomplished by passing strips of the paper, cloth, or other materials through the varnish and then through a wringer or squeeze roll to control the plastic content. The strips then enter an oven in which the liquid portion of the varnish is evaporated and the cure of the resin begun — but not completed — for the plastic must be left in a state permitting good bonding and quick and complete curing in the final lamination process.

[3] Shaw Insulator Company, Irvington, New Jersey.

[4] See Bibliography for list of MIL specifications pertaining to various grades of plastics.

Laminated sheet stock of the type used in terminal boards is formed in huge hydraulic presses under pressures of 1200 to 2500 pounds per square inch and temperatures of 300 to 350 F. Stacks of impregnated filler sheets are placed between chrome-plated steel face-plates where, under the combination of heat and pressure, the various layers bond as the resin flows, finally curing into a solid sheet. When the cure is complete, the presses are cooled and the laminate removed.

Laminated tubing is made from the same impregnated filler sheets as is the flat stock and may be *rolled tubing* or *molded tubing*.

Rolled tubing is made by passing the treated strips of filler over a heated roller and onto a steel mandrel of the proper diameter which is centered among three rollers applying pressure to the strips as they are wrapped around the mandrel. This combination of heat and pressure bonds the layers sufficiently to prevent unwrapping while they are being transferred to 275 F ovens for a 18 hour curing cycle. When cured, the tubes are stripped from the mandrel and centerless ground to the proper diameter.

Molded tubing is made by wrapping the required number of layers of impregnated filler on mandrels which are then placed in closed molds where heat and pressure cure the resins. Concentricity is more likely to be a problem with molded tubing, and some weakness may be noted along the "parting line" where the mold closes. Density, however, is higher in molded tubing with consequent lower water absorption — a point of importance in high-frequency applications.

Laminates, especially those with fillers of paper or glass, are of special interest to the electronic components designer since they offer a very desirable combination of mechanical strength, reasonably good electrical properties, and relatively high heat stability. Laminates are available in many different grades which offer a wide variety of electrical and mechanical characteristics.

CASTING

Casting is one more method of utilizing certain plastics. There are obvious advantages to a material which can be poured into a mold and cured under little or no pressure either at room or slightly elevated temperatures. Recent years have seen the development of several such resins. Chiefly of the thermosetting type, some of these resins have extremely high heat-distortion points and excellent electrical properties. Shrinkage is sometimes a serious factor in the use of casting resins especially where inserts are involved, but this can be minimized by the addition of fillers and by control of temperatures during the cure.

Practically all casting resins require the use of a catalyst. Depending upon the nature of the resin and the chosen catalyst, the reaction which accompanies the hardening process may or may not require the addition of heat. In all cases the reaction is distinctly exothermic; and whenever casting is considered, dissipation of the resulting heat must also be considered. As a means of embedment or encapsulation of electronic components, the use of casting resins has much to offer.

So easy are they to handle, that as a means of making a few plastic parts without incurring either the expense or loss of time involved in the design and fabrication of a steel sample mold, casting resins are without equal. They can be used in inexpensive, quickly and easily made plastic molds[6], which will, with reasonable care in handling, permit up to 50 or more pieces before requiring replacement.

This process is adaptable to the use of both ethoxyline (epoxy) and polyester type resins as well as to certain phenolic resins. Requiring a minimum of equipment beyond that found in the average coil laboratory and without the tieing up of appreciable amounts of space or money, the combination of casting resins and plastisol molds offers a highly practical and readily available means of proving out newly designed plastic parts without costly expenditures or delays.

ELECTRICAL PROPERTIES OF PLASTIC MATERIALS

Since nearly all plastics are insulators, their introduction into the field of electronics came about as a result of their high resistance properties. In present-day electronic components, plastics appear prominently in coil forms, terminal board, wire insulation, insulating tapes, treatment (coil impregnation) materials, and many other places. Both thermosetting and thermoplastic resins are used, depending upon the requirements of the particular application.

Polystyrene, together with its copolymers is one of the most satisfactory materials for use within the magnetic fields of radio frequency coils. The greatest drawback of this material is its relatively low heat distortion point, but credit must be given to the manufacturers who, through diligent research over the past ten or more years, have succeeded in moving the heat distortion point upward nearly 40 C. It is, however, unwise to consider polystyrene or any of its copolymers for continuous operation at temperatures much in excess of 85 to 90 C.

Selection of the proper plastic material for a particular application – for example, a coil form – is dependent upon several factors, and it is probable that no material will possess all the desired properties and that a compromise on some points will be necessary.

The most important single characteristic is probably the *power factor* of the material. Power factor may be defined in various ways, but it is essentially a measure of power losses and therefore should be as small as possible. There are no perfect insulators, and any non-conducting material that is introduced into the magnetic field of a coil brings with it a loss proportional to its power factor. Of all the common plastics, polystyrene with its average power factor of 0.0002 makes the best coil form material; while nylon and cellulose acetate, having power factors of 0.040 and 0.055 respectively, will substantially lower the Q's of coils wound upon them.

Moisture absorption is almost as important as power factor and largely for the same reasons. Moisture within the field of a coil means losses, and while proper impregnation may minimize the dangers of a material high in water absorption, it is to the advantage of a design engineer to select a material as low as possible in this particular characteristic. Here again, polystyrene ranks high on the list with a moisture absorption expressed in percentage of weight of water absorbed of 0.04, compared to 1.50 for nylon and 3.80 for cellulose acetate. Best of all the plastics in this regard are the polyesters whose water absorptions average 0.01 per cent.

The *volume resistivity* – actually a measure of the conductivity of the material – is important, particularly as it may affect electrolytic corrosion, and it is extremely important in those cases where fine sizes of magnet wire are to be in contact with the material. Reference to Fig. 6-6 will show polystyrene best in this regard, having an average of about 18 million ohms. It should be noted that there are fewer differences among the various materials in this characteristic than in most of the others.

Completing the list of important electrical characteristics are the *dielectric constant* and the *dielectric strength* of plastic materials. It is desirable that the dielectric constant be as low as possible in order to minimize the distributed capacitance of the windings and also the capacitance of the coil to the core when plastic coil forms are used with tuning cores of any type. Plastics vary widely in this property from certain polyesters with a dielectric constant of 2.3 and polystyrene with 2.6 to fabric and flock-filled phenolics which average between 8.5 and 10.0. Dielectric strength is usually recorded in volts per mil thickness and ranges from polystyrene with 600 volts per mil to fabric-filled phenolics with values as low as 250.

Mechanical characteristics of plastic materials are important to the designer of high-frequency components and will be found to vary greatly. Mechanical strength sufficient to permit handling and assembly operations is of extreme importance, and since plastic coil forms are often used with wall thicknesses as low as 0.008 inch, it is apparent that neither excessive brittleness nor flexibility can be tolerated. Descriptive literature available from nearly all plastics manufacturers will be of great assistance in the selection of proper plastic materials for specific applications.

DESIGN OF MOLDED PLASTIC PARTS

An understanding of plastics molding is essential to the design engineer if economically manufacturable parts are to be produced. Failure to consider the principles of good plastic parts design may lead to high mold costs, excessive piece breakage, warpage, or to other factors having a serious influence on both the cost and the effectiveness of the finished piece.

Intelligent design practice begins with an understanding of the basic principles involved in the design and construction of *plastic molds*. Molds are usually expensive as they are made from the best grades of tool steel, hardened and ground to withstand molding pressures of 2000 to 5000 pounds per square inch. All surfaces in contact with the plastic are highly polished to give a good appearance to the piece as well as to permit its easy removal from the cavity.

Molds – or *dies*, as they are often called – have a normal one-way (usually up and down) motion. This is important to remember as any feature which interferes with this normal motion or which cannot result from it will add to the ultimate cost of the molded piece. It is perfectly possible to put undercuts in a molded piece or to put holes in the side of it – but these and similar operations mean additional mold parts, often accompanied by complicated cam-actuated mechanisms to permit withdrawal of the finished piece. Such requirements should be avoided unless absolutely essential.

The space in the die where the molded part is formed is called the *cavity*. The final tool may contain a single cavity or it may contain a number of cavities which in transfer or injection molds are connected by channels through which the molding material flows to reach the cavities. These channels are called *runners*, while the main entrance into the die is known as the *sprue*.

Since plastic molds are necessarily expensive because of the amount of precision work that goes into them, steps should be taken to keep mold cost as low as possible. One procedure that is often followed is to include more than one type of cavity in the same mold as, for example, a base and cover assembly or right and left hand mating parts. As long as the pieces are of approximately equal size, no serious difficulties are involved in this practice.

In multi-cavity molds it is the generally accepted practice to make the cavities in such a manner that they can be removed and/or replaced without destroying or damaging the mold. This is an important feature of mold design since the heat and pressures encountered in the molding process often result in damage to a cavity which, if not of a replaceable type, would require replacement of the entire mold instead of merely the damaged portion.

Die cavities may be formed in two ways – by *hobbing* and by *machining*. If a large number of pieces are to be made, a die with several cavities will greatly reduce molding costs. In such cases, the cavities will probably be hobbed.

A *hob* is a piece of steel made into an exact replica of the part to be molded and then hardened. The hob is then pushed under tremendous pressures into pieces of low carbon steel where it forms cavities requiring only minor finishing operations, consisting chiefly of polishing and hardening, before being ready for insertion in the production mold. Hobbing offers definite advantages where several identical cavities are required. Difficult machining operations need be performed but once, since barring accidental damage, the hob can be used to form a considerable number of cavities, each exactly like the others.

For production molds with fewer cavities, it is usually more economical to form them by machining. Cavities are seldom machined in one piece but rather are made up from a number of pieces carefully fitted together. Precision work is called for, and even then it is not uncommon for minor differences to appear when the pieces are assembled into the final tool. For this reason, cavities are usually identified by letter or number so that pieces may be traced directly to the cavities that produced them.

Certain basic principles govern good molded plastic design practices. While almost any shape can be molded, the designer will do well to keep the following fundamentals uppermost in his mind:

1. Molded parts must have a certain amount of

[6] For instructions for making plastisol molds, see Section 13 of this manual.

taper or *draft* to prevent sticking in the cavity. While the permissible minimum will vary somewhat with the material and the particular application, it is generally conceded that at least one-half degree of taper must be allowed in all cases, and one degree should be given wherever possible.

2. Wall thickness is an extremely important point. Thin sections (0.040 inch or less) are likely to be troublesome, particularly in thermosetting materials. Uniformity of cross section does much to lessen the danger of warpage and also assists in more uniform cure and less shrinkage. Ribs may be used freely to stiffen thin sections and reduce the danger of warpage.
3. As an aid in designing a molded part, it is helpful to try to picture the cavity that will be required. If the cavity is to be hobbed, raised lettering or other features above the surface will be costly, while in a machined cavity they can be stamped or engraved in the mold without difficulty.
4. Holes in a piece mean pins in the mold. If the holes are small, the pins that form them are necessarily weak and subject to distortion and breakage as well as to comparatively rapid wear. Locating holes too near the edge of a piece will often result in a weakening of the walls at that point.
5. Wall sections joined at right angles should include a fillet at the point of joining, both for simplicity in mold design and for reducing potential breakage in the pieces. The radius of the fillet should be as large as is consistent with the requirements of the piece.

It is not the purpose of this discussion to do more than present the highlights of good molded plastic parts design. The suggestions offered here and in Fig. 6-7 are meant to guide the thinking of a transformer engineer to a point where he will be aware of the problems involved in molding and therefore will specify reasonable designs. The actual design of the mold is a problem for a mold engineer, and the final piece may well represent compromise on the part of both engineers.

Fig. 6-6 IMPORTANT PROPERTIES OF COMMON PLASTICS.

Fig. 6-6a (1 of 10)

MOISTURE ABSORPTION – The percentages by weight of water absorbed by a sample immersed in water. Depends on area exposed.

1. Polyester	0.01
2. Polyethylene	0.02
Hard Rubber	0.02
3. Styrene	0.04
4. Mica Filled Phenolic	0.07
5. Vinylidene	0.10
Shellac	0.10
Resin Filled Phenolic	0.10
6. Asbestos Melamine	0.13
7. Mineral Phenolic	0.18
8. Alpha Melamine	0.30
9. Cast Phenolic	0.35
10. Laminated Phenolic	0.50
11. Wood Flour Phenolic	0.60
Urea	0.60
12. Fabric Phenolic	1.00
Fabric Melamine	1.00
13. Ethyl Cellulose	1.50
Nylon	1.50
Cellulose Nitrate	1.50
Cellulose Propionate	1.50
14. Cellulose Acetate	3.80
15. Casein	10.50

Fig. 6-6b (2 of 10)

POWER FACTOR — In a perfect condenser the current leads the voltage by ninety degrees. When a loss takes place in the insulation the absorbed current, which produces heat, throws the ninety degree relation out according to the proportion of current absorbed by the dielectric. The Power Factor is the cosine of the angle between voltage applied and the current resulting. Measurements are usually made at million cycle frequencies.

Power Factor 10^6 Cycle

1. Styrene	0.0002
Polyethylene	0.0002
2. Polyester	0.0005
3. Hard Rubber	0.0050
4. Mica Filled Phenolic	0.0170
5. Glass Filled Melamine	0.017
6. Vinyl Co-polymer	0.018
7. Resin Filled Phenolic	0.019
8. Cellulose Propionate	0.020
9. Ethyl Cellulose	0.021
10. Butyrate	0.025
11. Alpha Melamine	0.029
Asbestos Melamine	0.029
12. Urea	0.032
13. Fabric Melamine	0.038
14. Laminated Phenolic	0.040
Wood Flour Phenolic	0.040
Cast Phenolic	0.040
Nylon	0.040
15. Vinylidene	0.050
Fabric Phenolic	0.050
Mineral Phenolic	0.050
16. Casein	0.052
17. Cellulose Acetate	0.055
18. Flock Filled Phenolic	0.050
19. Cellulose Nitrate	0.085

Fig. 6-6c (3 of 10)

SAFE TOP OPERATING TEMPERATURE in degrees Fahrenheit.

1. Mineral Filled Phenolic	400
Glass Filled Melamine	400
2. Flock Filled Phenolic	300
Asbestos Melamine	300
Nylon	300
3. Wood Flour Phenolic	280
Mica Filled Phenolic	280
4. Laminated Phenolic	260
5. Resin Filled Phenolic	250
Fabric Filled Phenolic	250
6. Fabric Filled Melamine	220
7. Alpha Filled Melamine	210
8. Polyester	185
Polyethylene	185
9. Vinylidene	180
Cellulose Propionate	180
Styrene	180
10. Lignin	175
Urea	175
11. Butyrate	170
Shellac	170
12. Acrylate	160
Ethyl Cellulose	160
Cellulose Acetate	160
Vinyl Co-polymer	160
Cast Phenolic	160
13. Cellulose Nitrate	130

Fig. 6-6d (4 of 10)

THERMAL EXPANSIVITY (Coefficient of Expansion) – The increase in length per unit length per degree centigrade rise in temperature.

1. Mineral Filled Phenolic	2.1
2. Lignin	2.3
3. Laminated Phenolic	2.4
4. Urea	2.8
5. Fabric Filled Phenolic	3.0
6. Wood Flour Phenolic	3.3
7. Flock Filled Phenolic	3.5
8. Melamine	4.0
9. Resin Filled Phenolic	4.1
10. Casein	4.4
11. Vinyl Co-polymer	7.0
Styrene	7.0
Cast Phenolic	7.0
12. Shellac	8.0
13. Acrylate	9.0
14. Nylon	10.0
Cellulose Nitrate	10.0
15. Cellulose Acetate	12.0
16. Butyrate	14.0
17. Cellulose Propionate	14.5
18. Ethyl Cellulose	15.0
19. Polyethylene	18.0
20. Polyester	18.0

Fig. 6-6e (5 of 10)

THERMAL CONDUCTIVITY is the time rate of the transfer of heat by conduction, thru unit thickness, across unit area for unit difference in temperature.

1. Mineral Filled Phenolic	12.0
2. Polyethylene	8.0
Polyester	8.0
3. Melamine	7.5
4. Urea	7.0
5. Laminated Phenolic	6.5
6. Cellulose Acetate	6.0
Butyrate	6.0
7. Nylon	5.8
8. Cellulose Nitrate	5.5
Wood Flour Phenolic	5.5
Flock Filled Phenolic	5.5
Fabric Filled Phenolic	5.5
Cellulose Propionate	5.5
9. Acrylate	5.0
Ethyl Cellulose	5.0
10. Resin Filled Phenolic	4.5
11. Cast Phenolic	4.0
12. Vinyl Co-polymer	3.7
13. Hard Rubber	3.2
14. Hard	3.0
Styrene	3.0

Fig. 6-6f (6 of 10)

SPECIFIC HEAT of a substance is the ratio of its thermal capacity to that of water at 15 degrees C.

1. Vinyl Co-polymer	0.23
2. Mineral Filled Phenolic	0.30
3. Vinylidene	0.32
Styrene	0.32
4. Hard Rubber	0.33
Fabric Phenolic	0.33
5. Acrylate	0.35
Butyrate	0.35
Laminated Phenolic	0.35
Cast Phenolic	0.35
Cellulose Nitrate	0.35
6. Wood Flour Phenolic	0.38
Flock Filled Phenolic	0.38
7. Cellulose Acetate	0.38
Urea	0.40
Cellulose Propionate	0.40
Nylon	0.40
8. Ethyl Cellulose	0.58
9. Polyester	0.55
10. Polyethylene	0.53

Fig. 6-6g (7 of 10)

VOLUME RESISTIVITY – The resistance in ohms between opposite faces of a centimeter cube of the material; is in the order of millions of ohms.

1. Styrene	18.0
2. Vinyl Co-polymer	16.0
3. Acrylate	15.0
4. Hard Rubber	13.5
5. Vinylidene	13.0
Nylon	13.0
Ethyl Cellulose	13.0
Polyethylene	13.0
Melamine	13.0
Polyester	13.0
Cellulose Propionate	13.0
6. Cast Phenolic	12.5
7. Resin Filled Phenolic	12.0
Urea	12.0
8. Fabric Filled Phenolic	11.5
9. Butyrate	11.0
Wood Flour Phenolic	11.0
Cellulose Nitrate	11.0
Laminated Phenolic	11.0
Mineral Filled Phenolic	11.0
10. Shellac	9.0

Fig. 6-6h (8 of 10)

DIELECTRIC STRENGTH – The voltage that will rupture or puncture the material in question when placed between electrodes of a given size. Dielectric tests are usually made at commercial frequencies, i.e., 60 cycles. The results will vary with the thickness tested. The thinner the section the higher the electrical gradient. Puncture voltage in volts per mil thickness is usually given in tables.

1. Styrene	600
Hard Rubber	600
2. Laminated Phenolic	550
3. Asbestos Melamine	535
4. Acrylate	500
5. Mica Filled Phenolic	475
6. Glass Melamine	460
Polyester	460
7. Cellulose Nitrate	450
8. Polyethylene	440
9. Vinyl Co-polymer	425
10. Cellulose Propionate	425
11. Casein	400
Shellac	400
12. Nylon	385
13. Cast Phenolic	375
14. Resin Filled Phenolic	350
Vinylidene	350
Lignin	350
Wood Flour Phenolic	350
Urea	350
15. Alpha Melamine	340
16. Flock Phenolic	325
Butyrate	325
Mineral Phenolic	325
17. Fabric Melamine	270
18. Fabric Phenolic	250

Fig. 6-6i (9 of 10)

DIELECTRIC CONSTANT – The ratio between the capacity of a condenser with a given dielectric and the same capacity with a vacuum as a dielectric.

	Dielectric Constant 10^6 Cycles
1. Polyester	2.3
2. Styrene	2.6
3. Hard Rubber	3.0
4. Vinyl Co-polymer	3.0
5. Acrylate	3.1
Ethyl Cellulose	3.1
6. Nylon	3.4
7. Vinylidene	3.5
Cellulose Propionate	3.5
8. Resin Filled Phenolic	4.5
9. Butyrate	4.7
10. Cast Phenolic	4.8
11. Nivea Phenolic	5.0
12. Cellulose Acetate	5.1
Wood Flour Phenolic	5.1
13. Fabric Phenolic	5.5
14. Fabric Melamine	5.6
15. Mineral Melamine	5.8
16. Flock Phenolic	6.0
Mineral Phenolic	6.0
17. Glass Melamine	6.0
18. Cellulose Nitrate	6.4
19. Casein	6.5
20. Urea	7.8
21. Alpha Melamine	8.0

Fig. 6-6j (10 of 10)

SPECIFIC GRAVITY – The ratio of the weight of the molded piece to the weight of an equal volume of water.

1. Polyestera	0.92
Polyethylene	0.92
2. Ethyl Cellulose	1.14
Nylon	1.14
3. Acrylates	1.18
4. Cellulose Proprionate	1.20
5. Butyrate	1.21
6. Molded Resin Phenolic	1.28
7. Cellulose Acetate	1.30
8. Cast Phenolic	1.32
9. Casein	1.35
10. Wood Flour Phenolic	1.36
Styramic	1.36
11. Fabric Filled Phenolic	1.38
12. Vinyl Copolymer	1.40
Lignin	1.40
Cellulose Nitrate	1.40
13. Urea	1.48
14. Melamine	1.49
15. Hard Rubber	1.50
Laminated Phenolic	1.50
16. Vinylidene	1.70
17. Shellac	1.90
18. Mineral Filled Phenolic	1.90

POOR ORIGINAL

GOOD DESIGN PRACTICES for MOLDED PLASTIC PRODUCTS

	DO	DO NOT	REMEMBER
Tolerances	Small dimensions: Minimum tolerance ±0.002". Large dimensions: Minimum tolerance ±0.005".	Do not specify close tolerances unless absolutely necessary	The closer the tolerance the higher the cost. Compression molding techniques require additional allowances across parting line of mold.
Draft	Allow a minimum of 1/2° taper (1° much better) on all vertical surfaces.	Do not specify less than 3° to 5° for deep barriers or bosses.	Ample draft is essential if pieces are to come out of mold.
Wall Thickness	Thermoplastics: Minimum 0.040". Thermosettings: Minimum 1/16". (3/32" better) Hold variations in cross-section to less than 3:1.	Do not call for abrupt changes in section. They are apt to cause localized stresses.	Side walls of deep molded pieces should have heavy section at bottom with a taper up to the top.
Holes	Specify round holes wherever possible. The more irregular the shape, the greater the expense. Allow greatest possible tolerance. Do not locate close to edge. Start holes at maximum size to allow for pin wear.	Do not specify oblique holes. Do not specify holes less than 1/16" in diameter or longer than 2-1/2 times the diameter.	Holes cannot be punched in molded phenolic or urea materials. Holes can be punched in most injection molded products.
Plane Surfaces	Use reeds, flutes, stipples, knurls, etc., on large plane surfaces unless willing to accept flow lines and shrink marks.	Do not specify other than simple designs. Intricate work means expensive engraving.	Flat surfaces tend to show shrink marks.
Edges and Corners	Break all corners with radii or fillets. As little as 0.010" radius may be used but 0.030" preferred.	Do not call for sharp edges on molded parts. Tool maintenance will be high.	Failure to provide for radii, bevel, or chamfer on edges and corners will add greatly to tool cost.

Fig. 6-7

Fig. 6-7 (continued)

Inserts	Specify minimum distance between inserts as 1/8" or greater. Use brass inserts for easy machinability, less damage to mold.	Do not use inserts that extend through molded part. Do not call for thin layers of plastic over inserts.	Provide sufficient anchorage by means of knurls, grooves, etc. Delicate inserts may collapse under molding pressures. Use inserts instead of tapped holes whenever possible.
Tapped Holes	Countersink holes before tapping. Specify flock filled materials for applications requiring maximum thread strength. NOTE: These materials generally poor electrically.	Do not call for tapped holes if screw is to be removed frequently.	Use inserts instead of tapped holes whenever possible. Self-tapping and drive screws very effective for light duty fastenings.
Thin Walls, Ribs, Bosses	Specify holes in heavy sections where possible. This reduces thickness and helps obtain proper curing.	Do not call for thin walls or ribs--particularly with sharp corners at base and little draft.	Try to keep wall thickness uniform. Heavy sections joined to thin sections make uniform curing difficult.
Warpage	Allow for minimum ±0.0025" per inch of length.	Do not plan on perfectly flat pieces--mold finish.	May be minimized by use of ribs or structural shapes and crowned surfaces.
Molded Threads	Use standard V-type thread. Avoid where possible--especially if less than 1/4" diameter.	Do not specify more than 32 threads per inch.	When internal threads are required, break edges of hole so that threads will not extend to top of hole.
Lettering	Depressed letters should be 0.008" to 0.010" deep and not more than 1/32" wide. Raised letters and lines 0.008" to 0.015" high.	Do not specify other than simple designs. Intricate work means expensive engraving.	Raised letters often less costly since they may be stamped or engraved in the mold when cavities are constructed. Depressed letters less expensive in hobbed cavities.
Undercuts	Try to avoid. Two pieces often cheaper.	Do not specify-- articularly on small pieces.	Undercuts can be molded only with complicated, expensive molds.
Shrinkage	Know the shrinkage of the material with which you are working and recognize that there will always be some after-shrinkage with aging.	Do not use molded pieces in places which require close dimensional accuracy.	Most noticeable on flat surfaces. Somewhat controllable by length of molding cycle.
Thermal Expansion	Specify slots or oversized holes if pieces are to be fastened to metal.	Do not fail to make allowance for this property of plastics.	Plastics expand and contract more than metals.
Post-Molding Operations	Specify only when absolutely necessary.	Avoid machining operations if possible.	Plastic materials are abrasive and cutting tools wear rapidly. Machining removes surface resin film which possesses superior water resistance and dielectric strength.

Fig. 6-8 COEFFICIENT OF THERMAL EXPANSION (LINEAR)

(10^{-5} per °C)

Substance	Temp. °C	Coefficient
Aluminum	-191 to +40	2.3
Brass (Cast)	0-100	1.8
Casein		8.
Cellulose Aceto-Butyrate ("Tenite" II)		11-16
Cellulose Acetate (Molding Powder)		11-16
Cellulose Nitrate (Pyroxylin)		12-16
Ethyl Cellulose		10-14
Glass (Plate)	0-100	0.9
[redacted]	[redacted]	[redacted]
Iron	-18 to +100	1.14
Lead	18-100	2.94
Methyl Methacrylate Resin ("Lucite" or "Crystalite")	0-75	8.
Paraffine	16-38	13.0
Porcelain	20-790	0.41
Quartz (Crystal)	-190 to +16	0.52
Rubber (Hard)		8.0
Steel	-18 to +40	1.32
Styrene Resin		6-8
Wood-		
Ash	0-100	0.95
Maple	2-34	0.63
Oak	2-34	0.49
Zinc	10-100	2.62

Fig. 6-9

DRILLS-NUMBER SIZES

No.	Diameter Inches	No.	Diameter Inches
1	.2280	41	.0960
2	2210	42	.0935
3	.2130	43	.0890
4	.2090	44	.0860
5	.2055	45	.0820
6	.2040	46	.0810
7	.2010	47	.0785
8	.1990	48	.0760
9	.1960	49	.0730
10	.1935	50	.0700
11	.1910	51	.0670
12	.1890	52	.0635
13	.1850	53	.0595
14	.1820	54	.0550
15	.1800	55	.05 20
16	.1770	56	.0465
17	.1730	57	.0430
18	.1695	58	0420
19	.1660	59	.0410
20	.1610	60	.0400
21	.1590	61	.0390
22	.1570	62	.0380
23	.1540	63	.0370
24	.1520	64	.0360
25	.1495	65	.0350
26	.1470	[illegible]	.0330
27	.1440	67	.0320
28	.1405	68	.0310
29	.1360	69	.0292
30	.1285	70	.0280
31	.1200	71	.0260
32	.1160	72	.0250
33	.1130	73	.0240
34	.1110	74	.0225
35	.1100	75	.0210
36	.1065	76	.0200
37	.1040	77	.0180
38	.1015	78	.0160
39	.0995	79	.0145
40	.0980	80	.0135

Fig. 6-10

LETTER SIZES OF DRILLS

Drill Size	Diameter Inches
A	.234
B	.238
C	.242
D	.246
E	.250
F	.257
G	.261
H	.266
I	.272
J	.277
K	.281
L	.290
M	.295
N	.302
O	.316
P	.323
Q	.332
R	.339
S	.348
T	.358
U	.368
V	.377
W	.386
X	.397
Y	.404
Z	.413

Fig. 6-11 (1 of 2)

TAP DRILL SIZES FOR PLASTICS

The following data are based on experience and is specifically for Thermosetting Plastics. Use these sizes or one size larger for Thermo-Plastics.

Use nearest available drill size, whether Number, Letter or Fractional Drill. If one does not have specified type, use nearest *larger* size. Specification of percentage thread is based upon Diameters in small sizes, and upon Pitch in larger sizes.

The formula used is —

$$D = T - \frac{n\%}{100}(2d)$$

in which

D = Drill diameter

T = O. D. of thread or tap

2d = Double depth of thread

$d = p(.64952)$

$$p = \text{pitch} = \frac{1}{\text{No.thds.per in.}}$$

n = Per cent of thread depth desired

Use the following percentage depths of threads:

50% below No. 6

60% No. 6 thru 14

70% 15 thru 30

70% 1/4 thru 1/2" N. C. S.

70% 1/4 thru 1" N. F. S.

75% 9/16 thru 1" N. C. S.

The "C" before the size shows the Standard N. C. S. size. The "F" shows the Standard N. F. S. size.

Fig. 6-11 (2 of 2)

Tap	Drill	Tap	Drill	Tap	Drill	Tap	Drill
F 0 x 80	55	8 x 30	28	16 x 16	3	N. C. S.	
1 x 56	1/16	C 8 x 32	28	16 x 18	7/32		
C 1 x 64	52	F 8 x 36	27	16 x 20	2	1/4 x 20	5
F 1 x 72	51	9 x 24	27	17 x 16	1	5/16 x 18	G
2 x 48	49	9 x 28	25	17 x 18	A	3/8 x 16	O
C 2 x 56	48	9 x 30	24	17 x 20	B	7/16 x 14	3/8
F 2 x 64	48	9 x 32	24	18 x 16	B	1/2 x 13	7/16
3 x 40	45	C 10 x 24	22	18 x 18	D	9/16 x 12	31/64
C 3 x 48	44	10 x 30	19	18 x 20	E	5/8 x 11	17/32
F 3 x 56	43	F 10 x 32	19	19 x 16	E	3/4 x 10	21/32
4 x 32	42	11 x 24	17	19 x 18	F	7/8 x 9	49/64
4 x 36	42	11 x 28	16	19 x 20	G	1" x 8	7/8
C 4 x 40	41	11 x 30	16	20 x 16	H		
F 4 x 48	40	12 x 20	16	20 x 18	I	N. F. S.	
5 x 30	37	12 x 22	15	20 x 20	J		
5 x 32	36	C 12 x 24	13	22 x 16	L	1/4 x 28	2
5 x 36	36	F 12 x 28	12	22 x 16	M	5/16 x 24	I
C 5 x 40	35	13 x 20	12	24 x 14	5/16	3/8 x 24	R
F 5 x 44	34	13 x 22	10	24 x 16	O	7/16 x 20	25/64
6 x 30	32	13 x 24	9	24 x 18	P	1/2 x 20	29/64
C 6 x 32	31	14 x 20	6	26 x 14	Q	9/16 x 18	33/64
6 x 36	31	14 x 22	5	26 x 16	11/32	5/8 x 18	37/64
F 6 x 40	1/8	14 x 24	4	28 x 14	23/64	3/4 x 16	11/16
7 x 28	1/8	15 x 18	5	28 x 16	U	7/8 x 14	13/16
7 x 30	1/8	15 x 20	4	30 x 14	W	1" x 14	15/16
7 x 32	30	15 x 22	3	30 x 16	X		
8 x 24	29	15 x 24	7/32				

BIBLIOGRAPHY

Du Bois, J. H.
Plastics
American Technical Society, Publishers, Chicago, Illinois, 1947

Hodgman, Charles D., and others
Handbook of Chemistry and Physics, Thirty-fifth edition
Chemical Rubber Publishing Company, Cleveland, Ohio, 1953

Scribner, George K.
A Ready Reference for Plastics
Boonton Molding Company, Boonton, New Jersey, 1950

SPECIFICATIONS

MIL-P-3115B
"Plastic-Material, Laminated, Thermosetting, Sheets, Paper-Base, Phenolic-Resin"

MIL-P-3413
"Plastic-Material, Molding; Rigid Thermoplastic, Polystyrene; For use in Electronic, Communications, and Allied Electrical Equipment"

MIL-P-15037B
"Plastic-Material, Laminated Thermosetting, Sheets, Glass-Cloth Melamine-Resin"

MIL-P-15047B
"Plastic-Material, Laminated Thermosetting, Sheets, Nylon Fabric Base, Phenolic-Resin"

MIL-P-14D
"Plastic-Materials, Molding, and Plastic-Parts, Molded; Thermosetting"

MIL-P-77A
"Plastic-Material, Thermosetting, Cast: For use in Electronic, Communications, and Allied Electrical, Equipment"

MIL-P-79B
"Plastic-Materials, Laminated, Thermosetting Rods and Tubes"

MIL-P-997B
"Plastic-Material, Laminated, Thermosetting, Electrical-Insulating:
Sheets, Glass Cloth, Silicone Resin"

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Boonton Molding Company
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Cleveland Container Company
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Continental-Diamond Fibre Company
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Dow Corning Corporation
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The Formica Company
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Cincinnati 32, Ohio

B. F. Goodrich Chemical Company
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Cleveland 15, Ohio

Insulating Fabricators, Inc.
150 Union Avenue
East Rutherford, New Jersey

The M. W. Kellogg Company
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Jersey City 3, New Jersey

Koppers Company, Inc.
Chemicals Division
Pittsburgh 19, Pennsylvania

Marco Chemicals, Inc.
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National Lead Company
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National Vulcanized Fibre Company
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Naugatuck Chemical
Division United States Rubber Company
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Nuodex products Company, Inc.
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Polymer Corporation of Pennsylvania
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Resistoflex Corporation
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Rex Corporation
Rexolite Division
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Shell Chemical Corporation
500 Fifth Avenue
New York 56, New York

Synthane Corporation
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Taylor Fibre Company
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Section 7

WAXES, VARNISHES, CEMENTS, AND LACQUERS

PURPOSE OF IMPREGNATION

Experience has shown that if a coil is to operate satisfactorily through extremes of temperature and humidity, it must be sealed completely against the entrance of moisture. The general term applying to this sealing process is *impregnation*. Numerous materials and methods of obtaining this end are in daily use.

Since the purpose of impregnation is to seal the coil completely, the most satisfactory process is one in which no voids remain which could serve as moisture traps. A well impregnated coil has not only a completely filled interior but also an exterior thoroughly sealed with a material which will not absorb moisture vapor. Coils wound with textile-served wire are extremely difficult to seal since the textile serves as a wick through which moisture readily enters.

The fact that the impregnating material is dispersed throughout the winding, and is, therefore, in the magnetic field, makes the electrical characteristics of the material of paramount importance. To avoid excessively high distributed capacity, the material should have a dielectric constant as low as possible. To avoid introducing excessive losses into the windings with a resultant decrease in Q, the power factor should be as low as possible. It should be noted that many impregnation materials have satisfactory power factors at room temperature, but these same materials at elevated temperatures may have so great a power factor as to make the materials entirely unsatisfactory for use at radio frequencies. Designers who expect their product to work at temperatures in excess of 60 C will do well to give this point serious consideration. Illustrations of the extent to which Q may be lowered by temperature may be found in Fig. 7-1.

Basically, the materials for impregnation may be sub-divided into three main classes: waxes, varnishes, and lacquers. A possible fourth class could be made up of the newer 100 per cent solids resins which are gaining increased popularity for reasons to be outlined later.

WAX

Waxes enjoy the widest use for applications where the highest operating temperature will not be in excess of 65 to 85 C and where extremely low temperatures will not be encountered. Since wax is essentially a "hot melt" type of material with a relatively high coefficient of expansion, it follows that at high temperatures there will be a tendency for wax to run somewhat or even to drip, while at extremely low temperatures shrinkage may cause cracking to a degree which often makes the treatment of little or no value.

A wide variety of waxes is available. One of the larger manufacturers[1] produces 180 types of waxes, most of which have been designed to meet specific requirements. In general, waxes have a wide range of temperature through which they will operate satisfactorily, although working at a temperature too near the "set point" may result in poor adherence and unsatisfactory moisture protection. Application of waxes is usually easy and frequently determines the type of wax to be used in a particular case. For the most part waxes have low dielectric constants, low power factors, and excellent moisture resistance. With the variety of waxes which are readily obtainable, it is a comparatively simple matter for a designer to select a wax which will afford the desired protection and at the same time will possess those properties adapting it to economical manufacturing processes.

Two general methods of applying wax to coil windings are in common use. The first consists merely of impregnation in the sense that the spaces or voids within the windings are filled by the wax, but no accumulation of wax appears on the surface

[1] Zophar Mills, Inc., 115 26th Street, Brooklyn 32, New York.

of the windings. This method is accomplished by dipping the windings in wax, the temperature of which is sufficiently high to permit the coils to absorb heat so that the surface will drain completely after removal from the wax pot.

A second method of using wax is to place a rather heavy layer over the surface of the coils where it serves as a barrier against the entrance of moisture. Such a coating is usually applied over windings that have been previously impregnated. The process of applying this coating is commonly called *flashing* and consists of dipping the coils into wax which is heated only slightly above its melting point. This dipping operation, while essentially simple, requires a certain amount of skill on the part of the operator since a slight twisting motion must be imparted to the coils both during and immediately after immersion to prevent the

Fig. 7-1

EFFECT ON Q[1] OF CHANGE IN TEMPERATURE OF TREATED COILS[2]

Material	Test Freq. 430 kc	Test Freq. 40 Mc
T-1	95.5	96.0
T-6	95.0	95.5
T-4 (v)	94.3	—
T-16	94.1	96.9
T-22	94.0	96.9
T-18	93.4	95.0
T-23	93.2	93.9
T-4	92.9	94.6
T-24	92.7	94.0
T-7	92.6	96.8
T-2 (v)	92.4	—
T-14	91.8	92.0
T-3	91.7	95.0
T-3 (v)	91.6	—
T-9	91.5	—
T-9 (2B)	91.0	96.0
T-2	90.9	91.6
T-12	90.4	93.6
T-19	90.2	94.7
T-8 (v)	90.0	—
T-8	88.8	94.1
T-26	88.0	93.9
T-25	87.0	91.2
T-15	84.0	89.5
T-17	81.3	92.7

[1]All figures represent the per cent of the *Q at room temperature* which was present at approximately 80 C after the coils had been stable for 5 minutes.

[2]Coil Data:
Form - Material, CF-3
OD, 1/2 inch
Wire - Insulation,
Size, No. 39
Gears - 103/68
Cam - 1/16 inch
Inductance - 1.275 mh ± 3 per cent

formation of air cells beneath the wax coating as well as to insure even distribution of the wax as it sets.

Properly conducted, the flashing process has as its end product a coil completely enclosed in a continuous wax coating which provides substantially improved moisture resistance when compared to similar coils which have been impregnated but not flashed.

Unfortunately, even the best impregnations using wax do not result in completely filled windings. A careful study of large wax-impregnated windings will show that the center portions contain a number of voids. These openings are brought about largely by the high shrinkage rate of the wax as it cools. The outside surface of the coil cools first, and shrinkage therefore begins at that point. As cooling progresses toward the interior of the coil, the wax continues to shrink, and the result is a movement of the impregnant toward the outside of the winding with resultant voids in the center. This action is aggravated by a tendency of the coil to "bleed" as it is withdrawn from the solution. This bleeding is the result of run-off from the little spaces dividing the heated wires. As a result, it is difficult, if not impossible, to fill completely the spaces within a coil using wax as an impregnant, and the condition becomes progressively worse as the size of the coil becomes larger.

IMPREGNATION METHODS

At least three methods of impregnation are widely used in the electronics industry today. The simplest of these consists merely of dipping the windings in the impregnant. For best results, this procedure should be prefaced by a baking cycle under conditions insuring the removal of all moisture from the winding. One-half to one hour in a ventilated, circulating type oven, at 225 to 250 F, is a very satisfactory cycle to accomplish this end. To avoid picking up moisture as well as to assure maximum penetration of the wax into the windings, the coils should go directly from the oven into the wax without being permitted to cool. To secure complete entrance of the wax into the windings, a coil must remain in the wax for a period of time sufficient to heat the copper to the same temperature as the wax. Watching the surface of the wax is a means of determining exactly when the penetration has ceased, since the entrance of an impregnant into a winding is always accompanied by displacement of air, showing up as bubbles rising to the surface of the tank. A common method of specifying the time for which coils must be impregnated is "Immerse in wax for a period of _ minutes or until bubbling ceases – whichever is the longer". For the average universal winding of 2 mh or less, five minutes will be about right, and the time for larger windings can be determined by the careful watching of a few coils as they are impregnated one at a time in a clean pot. Adherence to the above procedure will insure as complete impregnation as is possible by this simple method.

The basic fault to which this system of impregnation is subject is the possibility that all entrapped air within the windings will not be displaced by wax. Suitable agitation of the coils, especially at the time they are immersed in the wax, tends to minimize this danger as does also the fact that the coils are hot when they enter the wax and therefore any air that is entrapped will be relatively free from moisture and in an expanded state.

A second method of impregnation is that known as the *capillary* system. The difference between the capillary system and normal dipping lies in the fact that coils treated under the capillary system are never completely immersed in the impregnant but always have approximately 1/16 inch of the winding remaining above the surface of the liquid. The very fact that a small portion of the coil must extend above the surface of the liquid makes this a much more difficult and expensive operation to perform, but it does provide more complete impregnation because air can move upward through the winding with far less resistance than is encountered when it must come out through the impregnant. Furthermore, in this process the entrance of the impregnant into the winding results not only from normal liquid flow but also from the force of capillarity set up by the movement of the impregnant into the voids of the winding. So effective is this system of impregnation that coils thus treated are scarcely distinguishable from those treated by the third – and generally accepted to be superior – method of *vacuum impregnation*

Vacuum impregnation requires the use of two vacuum tanks with interconnecting piping so arranged that coils placed in one tank may be evacuated and, while free of air, be flooded with the impregnation material which enters from the second tank. Depending upon the need, at this point air pressure above atmospheric may be introduced to drive more completely the impregnant into the interior of the winding. Following this impregnation, the common practice is to send the impregnant back to the storage tank and then to draw another "dry" vacuum on the coils for the purpose of removing from the surface the excess material left from the initial impregnation.

Dipping, capillary, and *vacuum impregnation* procedures may be employed not only with waxes but with any of the common impregnation materials in use today. The actual equipment, however, must be suited to the particular impregnant to be used. For example, wax requires heated tanks, whereas varnish does not. In the selection of electrically heated tanks, particular attention should be paid to the type and location of the heating units which should be so designed as to provide uniform heating over the entire surface of the tank, since concentration of heat and the resultant "hot spots" will result in carbonization of the wax. Safest in this regard are those tanks in which the heat is supplied not by electricity, but by steam jackets. However, electrically heated tanks properly protected by thermostatic controls and having the heating elements located apart from the tank walls (thereby actually using air as the heat transfer medium) are perfectly satisfactory.

It is important to keep wax clean of all impurities, especially carbon and metallic particles or other conductive materials. Wax pots should be cleaned at regular intervals, with their contents either discarded or reprocessed to insure cleanliness. A good check of the condition of a wax pot is to dip a piece of clean white blotting paper into the wax, withdraw it slowly, and then see what it looks like when cool. Should the blotting paper differ appreciably in color from pieces of the fresh wax, contain black specks, or otherwise give indication of contamination, the wax should be discarded at once.

VARNISH

A common and generally satisfactory coil treatment-particularly for applications involving higher operating temperatures — involves the use of varnish. Varnishes consist for the most part of heat-blended mixtures of resins and drying oils dissolved in a solvent and are obtainable in two basic types — the oleoresinous varnishes, similar to those used in coating enameled magnet wire, and the newer and more common synthetic varnishes.

Oleoresinous varnishes were the first to be adopted by the electronics industry but because of the obvious advantages of the synthetic types are of secondary importance today.

The resins in this type of material are usually natural resins or gums and may include resin and Congo copal among others. The fact that these resins are natural products which exude from trees makes them difficult to control from batch to batch and accounts, at least in part, for the change to synthetic resins.

Drying oils, as used in varnishes, (See Fig. 7-2) are derived primarily from nuts and seeds and occasionally from animal fat. All of these materials have the property of taking oxygen from the air whereupon they convert to a solid film as a result of oxidation and, to a certain extent, of polymerization. Probably the two most common drying oils are the familiar linseed oil, which was the first to be used in varnish making, and tung oil, which is extracted from the nut of the Chinese tung tree. Linseed oil is very slow drying and produces films which are flexible, while tung oil dries more rapidly into films which are noted for toughness and moisture resistance. A further advantage of tung oil lies in its uncommon ability to dry quickly and completely, even in relatively thick films. Combinations of these and other oils are often used in the manufacture of varnishes for electrical applications.

It is not uncommon to find the expressions *short oil* or *long oil* used in connection with varnishes. These terms mean simply the amount of drying oil compared to the resin and are given in gallons per hundred pounds of resin. A short oil varnish would use something in the order of ten gallons of oil to a hundred pounds of resin, while a long oil varnish might carry as much as seventy to eighty gallons to an equal amount of resin.

Oleoresinous varnishes are cured essentially by the annexation of oxygen although it appears probable that a certain amount of actual polymerization takes place during the bake. Curing varnishes of this type presents something of a problem since particular care must be taken to prevent the surface of the varnish coating from "setting up" before the solvent has been removed from the interior. The seriousness of this surface sealing will be apparent if one considers that there is but little possibility of the solvent's escaping once the surface has cured. Continued heating can then, at best, succeed only in driving out the remaining solvent through openings in the surface

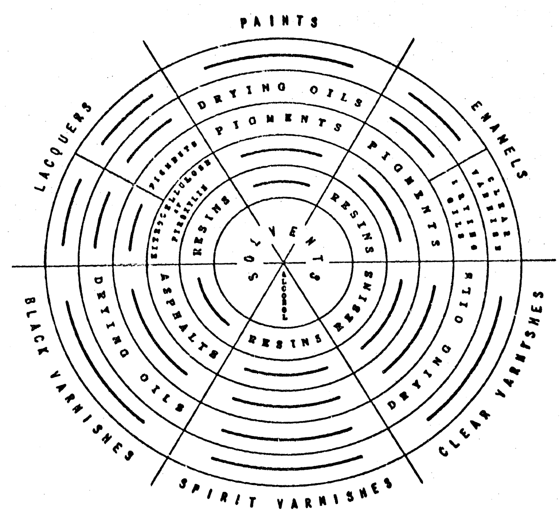

Fig. 7-2 Composition of varnishes, lacquers, paints, and enamels.

film, formed as a result of the vapor pressure of the entrapped solvent which, of course, leaves the windings with surfaces roughened by crater-like openings, the presence of even one of which will effectively destroy the moisture sealing properties of the treatment.

The most practical way of getting rid of solvents without damage to the protective coating usually involves a two-stage cure following a period in which the coils are permitted to drain and air dry. The actual baking (curing) cycle should have the first bake at about 180 F and the second and final cycle at somewhere in the order of 250 F. The exact duration of the treatment cycle is a function of the particular winding and the selected treatment material and, for best results, should be individually determined for each coil. An illustration of a typical treatment specification developed for a specific coil is included in this discussion as Fig. 7-3.

SOLVENTS

Because commercial solvents of the type used in varnishes may be broadly grouped into three general classes, and because these three classes vary substantially in their solvent powers, it seems advisable at this point to discuss briefly these groupings beginning with the aliphatic hydrocarbons which include such liquids as heptane, VM&P (varnish makers' and painters' naptha) and mineral spirits. These materials have the lowest solvent power of all commercial solvents and therefore can

Fig. 7-3 SAMPLE COIL TREATMENT SPECIFICATION

THE ALPHA COIL COMPANY

Treatment Specifications

No. V14-7

To be used on — Effective date June 17, 1954

Philhouse 7X-1535Z

Materials and equipment required.

29-5319	Coil Racks
75-9168	Baking Varnish
86-1907	Thinner
Vacuum impregnation equipment	
180F	Circulating oven
250F	Circulating oven

Minimum length of treatment cycle:

6½ hours plus handling time

PROCEDURE:

1. Rack coils on 29-5319 racks, dressing leads into clips.
2. Bake racked coils 1 hour at 220-230F.
3. Place coils in vacuum tanks.
4. Draw dry vacuum (minimum of 29 inches) and hold for 20 minutes.
5. Without releasing vacuum, flood coils with 75-9168 varnish from storage tank.
6. Maintain vacuum (minimum of 29 inches) for 30 minutes.
7. Release vacuum and admit atmosphere.
8. Allow to stand for 15 minutes.
9. Return varnish to storage tank.
10. Draw dry vacuum (minimum of 29 inches) and hold for 15 minutes.
11. Release vacuum.
12. Remove coils from tank and allow to air dry for ½ hour.
13. Bake at 170-180F for 1 hour.
14. Bake at 245-255F for 2½ hours.
15. Remove from racks and place in tote boxes.

NOTES:

1. Solids content of varnish must be checked before start of each work period.
2. Solids content of varnish must be held between 45 and 50 per cent. Use 86-1907 thinner as required.
3. Racks must be cleaned following every fifth cycle.
4. Refer all questions to V.T. Lambda, Treatment Engineer.

Specifications written by:

G.P. Beta

Approved:

I.Q. Sigma

be expected to have the least effect upon tapes, wire insulations, cements, and similar products with which they may come in contact. Fortunately most electrical-grade varnishes use one or more of the above-named solvents.

The group next to be considered is made up of the aromatic hydrocarbons such as toluol, benzol, xylol. Considerably higher in solvent power and found in a number of currently available varnishes, these materials should be used with caution in any electronic application. Wire insulations such as enamel and Formex and pressure-sensitive tapes, even of the thermo-setting variety, are somewhat subject to attack by solvents of this group, and immersion in any material containing aromatic hydrocarbons should be held to an absolute minimum. For those instances where no other material has the desired properties, it is advisable to follow the procedure outlined elsewhere in this manual and to provide the component with a preliminary coating of a material utilizing an aliphatic hydrocarbon solvent, thus preventing the more active agents from contacting the vulnerable portions of the unit.

Most powerful of all solvents, and as a consequence rarely used in varnish making, are the esters, ketones, and chlorinated hydrocarbons. This group of solvents is not recommended for any electronic application where there is even a remote possibility of the materials coming in contact with wire insulations, tapes, or other similar organic substances. Ethyl acetate, amyl acetate, methyl ethyl ketone, methyl iso-butyl ketone, acetone, and carbon tetrachloride are among the solvents belonging to this group.

Synthetic varnishes are closely related to the thermo-setting plastics which are so common in the industry and most often are of the pheno-formaldehyde or melamine type (See Fig. 7-4). As received, these materials contain the resin together with its drying oils in a solvent and usually have an actual solids content of somewhere between 50 and 60 per cent. The curing process consists first of ridding the varnish of the solvent, and secondly, of the actual conversion or polymerization of the resin. Varnishes of this group are less susceptible to surface sealing than are those of the oleoresinous type and also tend to cure more completely, thus providing more uniform electrical characteristics. Use of synthetic varnishes under proper conditions results in a hard, smooth, and highly water-resistant film, surrounding a reasonably well-filled coil. Since these materials are at best only 60 per cent solid matter, it is apparent that complete impregnation of the winding cannot be accomplished in one attempt. Actually it would require an infinite number of immersions to fill every void, but practice has shown that two or three impregnations, each followed by dipping in the varnish and then baking until cured, are satisfactory for all except the most severe cases.

A common complaint in the use of varnish treatments is the irregular surface made up of crater-like depressions, having sharp edges, which appear all too frequently on varnished coils. This condition can invariably be traced to improper curing — usually insufficient drying time before being placed in the oven — or more rarely, to too high an initial temperature in the baking oven. As has been mentioned previously, from 40 to 60 per cent of a varnish is solvent which must be evaporated before the resin can be converted. If freshly impregnated windings are placed in an oven at elevated temperatures, the tendency is for the varnish immediately to skim over, making a film which retards the evaporation of the solvent. While generally held to be not so serious in the case of synthetic varnishes as with oleoresinous materials, this situation still is undesirable and is almost certain to result in the formation of bubbles or blisters, leaving a rough surface which affords poor moisture protection; and, in the case of high voltage components, the condition may provide points from which corona can readily originate. A minimum of one-half hour air-drying between impregnation and baking is essential to prevent excessive roughness of surface.

Another point not to be over-looked is the importance of using ventilated, circulating ovens in which a sufficient amount of fresh air is always present to insure against excessive vapor pressures which will retard solvent evaporation.

Ovens used for baking varnished coils must be kept clean and must be so constructed as to keep the possibility of explosion to a minimum. Door latches which will release under internal pressure, a limit switch to prevent overheating in the event of a thermostat failure, low surface temperature heaters, and a direct vent to the outside, are items of extreme importance — all of which are available in the products of reputable oven manufacturers.

HIGH TEMPERATURE PROTECTIVE TREATMENTS

The need for coils so treated as to permit operation at temperatures up to 125 C has brought many

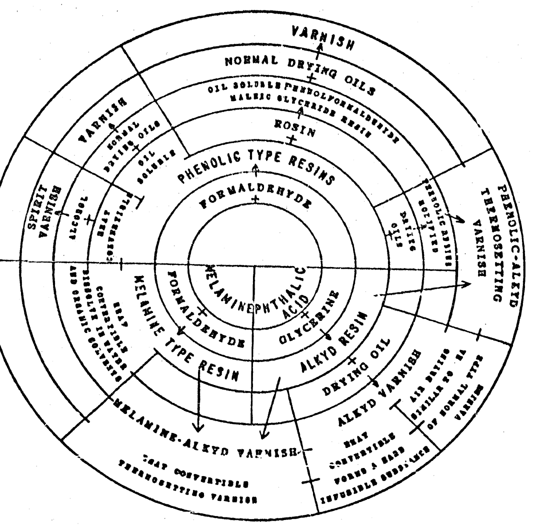

Fig. 7-4 Basic composition of common varnishes

new problems. This temperature is well within the safe-operating range of silicone varnishes, but their use presents many difficulties, inasmuch as they are made with xylol or toluol as a solvent and therefore tend to attack both enamel and Formex wires. In addition, their complete cure cannot be effected at less than 200 C — a temperature too high for the average wire insulation.

Since many of the currently available synthetic varnishes can be used safely at temperatures up to 125 C, a satisfactory treatment procedure[2] which also has excellent moisture resistance properties consists of a first coat of any good thermosetting synthetic varnish, preferably diluted approximately 50/50 with VMP naptha to permit thorough coating of the wire insulation. Over this light coat of varnish, which is applied mainly for the protection it affords the wire insulation, is placed a coat of silicone varnish, such as Dow Corning DC 996. This silicone varnish should not be fully cured but should remain in a slightly tacky condition[3] to insure maximum moisture protection. A baking cycle

[2]Developed by Automatic Manufacturing Corporation under Signal Corps Contract No. Da-36-039-sc-15321.

[3]By "slightly tacky" is not meant a surface sufficiently soft to collect dust but rather one which is soft only by comparison with a fully cured silicone varnish film. Actually it is difficult to distinguish between a fully-cured film and the type which is arbitrarily called "slightly tacky".

of approximately eight hours at 150 C has been found to provide satisfactory resistance to moisture. The advantages of silicone varnish are made available in this manner without the introduction of any deleterious effects upon wire insulations since the conventional varnish coating effectively seals off the windings, thereby preventing contact between the xylol or toluol and the wire insulation.

LACQUERS

Air-drying lacquers, sometimes called "spirit-varnishes", are another type of coil treatment materials. Lacquers are basically air-drying materials and are usually solutions of thermo-plastic resins in alcohol or other rapidly evaporating solvents. Low power factors and low dielectric constants give materials of this sort good electrical properties, but the moisture protection provided by lacquers is usually inferior to that of wax or varnish. For non-critical applications they have the advantage of ease of handling since they may be applied by any conventional method and permitted to dry from periods of 30 minutes to several hours, depending upon the nature of the winding and the degree of penetration achieved. Unfortunately, it is difficult to determine when a lacquer film is completely solvent-free since tests[4] demonstrate that a 2 mil film may retain up to 25 per cent by weight of solvent when apparently hard and dry and after as much as 48 hours of air-drying may still retain 2 per cent of solvents.

A common use of lacquers is in the treatment of solenoid windings, where the material is applied either by brush, dipping, or by rolling on a saturated felt pad. In those instances where several coats are required, it is generally better to conduct the drying in an oven operating at a temperature of between 140 F and 200 F. Best results may be expected from a longer cycle carried out at a lower temperature, since the process is solely one of ridding the mixture of solvents — a process made more difficult by the formation of any surface skin.

100 PER CENT SOLIDS RESIN IMPREGNANTS

Interest is growing rapidly in the newer types of 100 per cent solids resins for impregnation purposes. These resins are generally of the ethoxylene or polyester type. Since these materials are 100 per cent solids, it is now possible to secure a degree of impregnation which has hitherto been impossible. Complete filling of all spaces within a winding is desirable, both from the standpoint of moisture prevention and for protection from corona in the case of high voltage applications. Electrically, the polyester resins exhibit somewhat superior qualities, but from the standpoint of moisture protection the ethoxylene resins with their superior bonding qualities provide better sealing. Both types of material have a serious drawback in that they require catalysts for their conversion and have as a result a limited pot life. Depending upon the particular resin and the choice of accelerator, this pot life may be as short as five minutes, or it may extend to a matter of days. One property which makes these materials — particularly the polyesters — of particular interest is their high heat distortion point, which means that many of them can be operated without damage at temperatures far in excess of 125 C.

Reference to Fig. 7-1 will show the effect of using resins of these types at various frequencies and temperatures. Particular note should be paid to the tendency of ethoxylene resins to bring about lower Q's when operated at elevated temperatures as well as to the fact that this tendency can be minimized by the addition of small amounts of relatively inert oils such as Kel-F No. 3.[5]

CEMENTS

Cements are of considerable importance in the manufacture of electronic components. One very important use of adhesive materials is to anchor the starting turns of a winding to the form; another is to hold the finish turn, thus terminating the windings.

Cements used for these purposes fall roughly into two categories — those which are "hard" when "set" and those which remain in a flexible state. Both types have their advantages, but one should not overlook the value of the flexible type as a means of terminating windings that are made up of small wires. When such windings are terminated with a hard cement, the result is a sharp edge against which the wire must flex. Excessive lead breakage may be expected under these conditions. The use of a flexible cement such as Pliobond[6] which is rubbery in nature results in a flexible bond with sufficient "give" to minimize to a large extent lead breakage at the point where the wire leaves the surface of the coil.

In all cases where treatment materials or

[4]Conducted in the Components and Materials Branch of Signal Corps Engineering Laboratories.

[5]Manufactured by the M.W. Kellog Company, Jersey City, New Jersey.

[6]Manufactured by Goodyear Tire & Rubber Company, Akron, Ohio.

POOR ORIGINAL

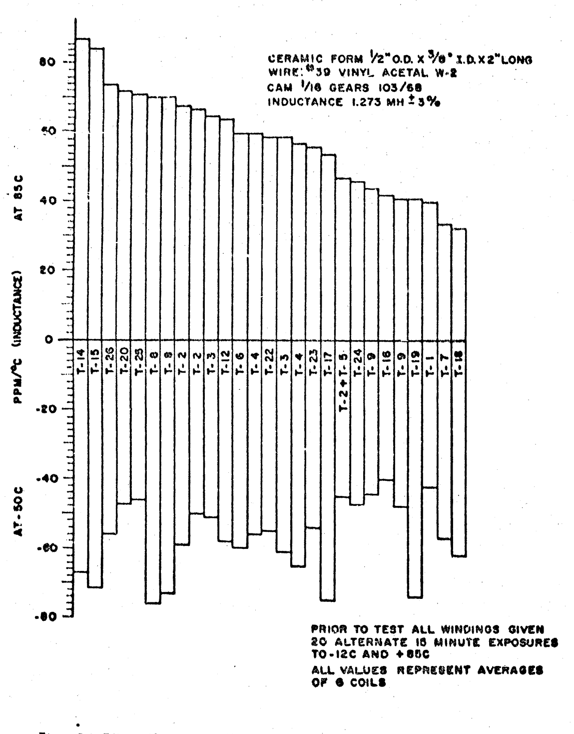

Figure 7-5 Effect of Impregnation Material and Methods upon Temperature coefficient.

cements are used on windings, care must be taken that the particular materials chosen are compatible with the insulation of the wire. In Fig. 1-9, a list of common wire insulations and the solvents which degrade them is given. Any varnish containing xylol or toluol — particularly in combination with alcohol — is unsafe for use directly over Formex wire. Another case of incompatibility would be the use of a cement containing acetone on a wire insulated with cellulose-acetate lacquer film or celanese textile serving. Of even greater importance is the resistance offered to corrosion by all sorts of treatment materials and cements. Nearly all coils have some d-c voltage across them when in use. It takes only a minute quantity of ionizable material joining with the moisture and impurities in the air to start an electrolytic action which soon erodes away the tiny copper conductors.

New materials under consideration should, regardless of their source, be carefully checked for possible corrosive effects. In Section 8 of this manual may be found a more detailed discussion of electrolytic corrosion, its causes, and recommended test procedures.

SELECTION OF IMPREGNANTS

Selection of the best impregnants and of the best method of impregnation is largely dependent upon the requirements of the particular coil in question. If the coil is to be operated under severe conditions of humidity and high temperature, the designer's choice is automatically directed toward either a dual varnish treatment or one of the new 100 per cent solids resins. If, however, the maximum operating temperature of the unit will not exceed 85 C, and the requirements are not too severe with respect to humidity, wax will be perfectly satisfactory and far less expensive except for those cases where poor low temperature characteristics make it impractical. Since the choice of treatment materials has a definite effect upon the temperature coefficient of the coil, this too must be taken into consideration. In Fig. 7-5 may be seen the comparative effects of various treatment materials upon the temperature stability of coils of identical types. A glance at this table will show that where maximum temperature stability is a requisite, the best impregnation material is probably wax. Should the temperature requirement for the coil be such that wax could not be used, the next best material would appear to be silastic rubber.

Fig. 7-6 shows the effect of exposure to static humidity upon the Q of coils impregnated with various materials which, in some cases, are applied by more than one method. Here again, wax is indicated as superior, providing the operating temperature range does not exceed that recommended for wax. If a wider temperature range is required, either a suitable cement or a silicone rubber is recommended. Many other materials are compared and all have desirable characteristics for specific applications.

An important aspect of design work is the selection of those materials and methods which will most economically produce a finished unit capable of a specified performance. The importance of both the treatment material and its method of application cannot be disregarded since many aspects of ultimate coil performance are directly related to the treatment accorded the coil subsequent to winding. To assist in the selection of both the impregnation material and its method of application, it is recommended that the data presented in this section be studied carefully and a treatment which appears suitable, selected for the impregnation of experimental coils. If, after test, the selected treatment fulfills the requirements a specification patterned after that shown in Fig. 7-3 can be issued. This information is intended to provide the design engineer with background material upon which to call when a treatment must be specified. Properly interpreted, it can be of definite value in the determination of the most satisfactory coil treatment.

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POOR ORIGINAL

Fig. 7-6

EFFECT UPON Q[1] OF EXPOSURE TO STATIC HUMIDITY[2] OF COILS[3] IMPREGNATED WITH VARIOUS MATERIALS

Treatment	Drying time following exposure			
Material	1/2 hour	4 hours	24 hours	48 hours
T-1	96.0	97.1	98.1	98.4
T-18	75.6	95.5	98.1	98.4
T-16	89.0	90.9	96.8	97.8
T-9 (2B)	92.1	93.7	95.9	96.8
T-2	89.4	93.7	93.7	96.5
T-7	95.5	95.9	95.5	96.5
T-17	40.2	60.8	88.2	95.9
T-6	93.8	94.6	94.8	95.8
T-4	89.5	91.4	93.5	95.2
T-24	84.3	89.6	94.0	95.0
T-3 (v)	82.5	87.0	94.5	94.9
T-23	93.5	93.5	93.5	94.5
T-25	73.5	82.5	92.4	94.5
T-2 (v)	86.2	87.0	93.2	94.3
T-22	88.5	92.5	92.5	93.1
T-8	83.3	88.8	91.9	93.0
T-8 (v)	83.0	89.0	91.4	92.0
T-12	72.6	91.0	91.0	91.8
T-15	87.6	88.7	90.8	90.8
T-19	73.4	75.1	81.1	90.4
T-14	84.5	85.9	88.1	89.8
T-3	80.1	85.0	86.3	89.6
T-26	79.8	79.9	83.1	85.0
T-9	81.1	83.0	83.5	83.8
T-4 (v)	55.5	60.9	65.9	68.0

[1]All figures represent the per cent of *Q before exposure* which was present at the indicated time. Test frequency was approximately 430 kc.

[2]95 per cent relative humidity and 40 C for 200 hours.

[3]Coils made to the same specifications as those in Fig. 7-1

BIBLIOGRAPHY

Young, James F.
Materials and Processes
John Wiley & Sons, New York, 1949

Moses, Graham Lee
Electrical Insulation, First Edition
McGraw Hill Publishing Company, New York, 1951

CATALOGS AND TECHNICAL INFORMATION OF:

The Acme Wire Company
New Haven, Connecticut

Bakelite Company
30 E. 42nd Street
New York, New York

Biwax Corporation
345 Howard Street
Skokie, Illinois

Bond Adhesive Company
537 Johnson Avenue
Brooklyn, New York

John C. Borthig Company, Inc.
P.O. Box 115
Rutherford, New Jersey

Ciba Company, Inc.
627 Greenwich Street
New York 14, New York

Communications Products Company, Inc.
Marlborough, New Jersey

John C. Dolph Company
Monmouth Junction, New Jersey

Dow Corning Corporation
Midland, Michigan

Emerson & Cuming
869 Washington Street
Canton, Massachusetts

General Electric Company
Chemical Division
Pittsfield, Massachusetts

Houghton Laboratory
Olean, New York

Insulation Manufacturing Corporation
565 West Washington Boulevard
Chicago, Illinois

The Insl-X Company, Inc.
Ossining, New York

Irvington Varnish & Insulator Company
Division of Minnesota Mining &
Manufacturing Corporation
Irvington 11, New Jersey

Linde Air Products Company
Division of Union Carbide &
Chemical Corporation
30 East 42nd Street
New York, New York

The Marblette Corporation
37-27 30th Street
Long Island City 1, New York

Minnesota Mining and Manufacturing Company
St. Paul, Minnesota

Mitchell-Rand Insulation Company, Inc.
51 Murray Street
New York 7, New York

Naugatuck Chemical
Division of U.S. Rubber Company
Naugatuck, Connecticut

Sauereisen Cements Company
Pittsburg, Pennsylvania

Zophar Mills, Inc.
115 26th Street
Brooklyn 32, New York

SPECIFICATIONS

JAN-T-152
"Treatment, Moisture and Fungus Resistant, of Communications, Electronic, and Associated Electrical Equipment; General Process for"

JAN-V-1137
"Varnish, Insulating (Electrical)"

JJJ-W-141(1)
"Wax: Carnauba"

MIL-V-173A
"Varnish, Moisture and Fungus Resistant, For the Treatment of Communications, Electronic, and Associated Electrical Equipment"

TT-V-130(1)
"Varnish; Spirit (Shellac Varnish Replacement)"

TT-P-141b
"Paint, Varnish, Lacquer, and Related Materials; Methods of Inspection, Sampling, and Testing"

TT-T-266a(1)
"Thinner; Dope and Lacquer (Cellulose-Nitrate)"

TT-T-306(2)
"Thinner; Synthetic-Enamel"

TT-O-364(2)
"Oil; Linseed, Boiled (for Use in Organic Coatings)"

TT-O-367
"Oil: Linseed, Heat-Polymerized (Bodied), For Paint, Varnish, and Enamel"

TT-O-369(2)
"Oil; Linseed, Raw (For Use in Organic Coatings)"

TT-O-371a(1)
"Oil; Linseed; Replacement (For Use in Organic Coatings)"

TT-P-381(2)
"Pigments-In-Oil; Paint Color"

TT-O-388(1)
"Oil, Soybean, Refined (For Use in Organic Coatings)"

TT-O-395(1)
"Oil; Tung (China-Wood, Raw) (For Use in Organic Coatings)"

TT-E-489(1)
"Enamel, Gloss, Synthetic (For Exterior and Interior Surfaces)"

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New York 20, New York

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Signal Corps Engineering Laboratories
Fort Monmouth, New Jersey

Mr. A.J. Raffalovich, Materials Section
C & M Branch
Signal Corps Engineering Laboratories
Fort Monmouth, New Jersey

Section 8

TAPES AND FILM INSULATIONS

USES IN ELECTRONICS

In the electronics industry, tapes are used primarily as tools. Unlike tools in the ordinary sense, to be completely acceptable for use on coils and transformers, tapes must not only do the job to which they are assigned but must do it without adding new problems such as electrolytic corrosion or failure during impregnation.

Of the two basic functions performed by tapes—holding and protection—the first named accounts by far for the greatest number of applications of tape to coils and transformers. For example, tapes are commonly used to hold the wire on the coil form at the start of a winding. For this purpose, small pieces of tape are simply pressed over the start lead, thus holding it against the coil form until the winding begins to build up.

Similarly, narrow pieces of tape are frequently used to hold down the finish lead of a winding, particularly until the cement has had time to set. At this point the tape may or may not be removed, depending upon the type of tape employed and the treatment which is to be accorded the winding. For those applications where it is probable that the finish lead may be subjected to some strain, it has been found that a narrow strip of tape completely encircling the winding as shown in Fig. 8-1 often provides the support necessary to prevent the lead from breaking loose.

Some of the more common uses of tape are actually a combination of the two basic functions in that they provide protection while simultaneously serving to hold objects which would otherwise be without support. An example of this type of application is to be found in those transformers where extremely high coupling is required between two windings and is obtained by placing the second winding directly on top of the first. In such cases, it is customary to place a layer or two of tape around the first winding — an operation which accomplishes the dual objective of

(1) affording protection to the first coil during all subsequent handling and winding operations and
(2) providing a more level surface upon which to wind the second coil.

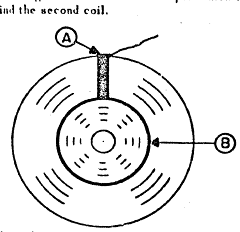

Fig. 8-1. Uses of tape in a typical high-voltage coil.

It is frequently necessary to insert a tap within a winding and to provide that tap with both mechanical and electrical protection. For such applications tape is an ideal tool since it serves to hold the wires in place, while, at the same time, it provides insulation from adjacent wires and protection against damage from the winding button. In this as in all applications where the tape remains as a part of the finished coil it is especially important to use tape which is of the highest electrical quality and of a composition compatible with subsequent impregnation materials and processes.

Protection against the introduction of electrolytic corrosion is frequently entrusted to electrical grade tapes. In windings for intermediate frequency

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transformers, for example, it is sometimes necessary to use coil forms of materials that are subject to corrosion when exposed to humidity and d-c voltage.

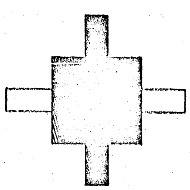

Fig. 8-2. Example of two windings separated only by electrical-grade tape.

For years it has been an accepted practice in such instances to wind one coil – preferably the one which is to operate at the highest potential – on top of a layer or two of acetate tape wound around the coil form. (See Fig. 8-3). In this manner the winding is insulated from the coil form by a material not subject to breakdown in the presence of humidity and voltage, and thus the assembly is provided with some degree of insurance against damage from corrosion caused by electrolysis.

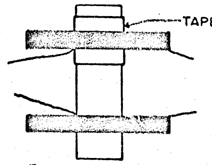

Fig. 8-3. Tape used to insulate winding from coil form.

A practice which is not particularly common at the higher frequencies but which has been used in power applications for many years calls for entire windings to be wrapped in tape prior to impregnation or encapsulation. In this manner almost complete protection against loose turns and/or mechanical damage to the windings is assured. A further advantage to be found in this procedure lies in the fact that the turns of tape provide an excellent base for treatment materials, thus insuring maximum pro-protection to the enclosed winding.

While on the subject of applications for tapes of electrical quality within the field of electronics, one cannot overlook the very important cases in which tape serves to insulate from their shield cans such assemblies as i-f and r-f transformers. This sort of application is becoming increasingly important as more and more miniature and sub-miniature units are developed where in it is frequently necessary to insure against a high potential lead shorting to the shield can under conditions of shock or vibration. The degree of dielectric strength obtainable in a thin film makes tape ideal for applications of this kind with the presence of the pressure-sensitive adhesive adding value by insuring against movement of the insulating film once it has been placed in position.

The above mentioned uses of tape by no means complete the list of cases where tapes have been employed in electronic components. The examples listed are believed to be representative and are intended to indicate something of the extent to which tapes have become a tool of the industry.

INDUSTRIAL AND ELECTRICAL GRADES

Commercially available tapes can be classified in various ways, the most important of which is undoubtedly that which sets forth the distinction between *industrial* and *electrical* grade tapes. The outward appearance of the two grades is identical, but the choice of industrial grade tape for use within a coil could be disastrous. The basic difference between the two lies in the control that is exercised during the manufacturing processes and in the raw materials that are used. Electrical-grade tapes are carefully made under controlled conditions from selected materials known to be as free from ionizable material as is possible. In other words, the chief concern of the manufacturer of electrical-grade tapes is to turn out a product which will not induce corrosion under any condition of humidity and voltage.

At this point it should be clearly understood that it is entirely possible to purchase industrial-grade tape which will pass every corrosion test that may be given it, but under no circumstances should this fact be taken as evidence that this grade of material can be used with confidence in electronic applications. It is entirely possible that conditions during the manufacture of this particular lot of tape were such that no ionizable material was introduced, thus making the final product the exact equivalent

of electrical-grade tape. There is no assurance whatsoever that the next lot will come even close to having the same quality; hence the use of this type of tape is unfair both to the manufacturer of the tape and to the purchaser of the electronic components in which it is used.

ADHESIVES

A second means of grouping tapes might be in accordance with the type of adhesive used. Except in very rare instances, tapes used in electronic applications are of the *pressure-sensitive* type, as distinguished from the common varieties of gummed tapes whose adhesives must be softened with water before application. Gummed tapes are entirely unsatisfactory for electrical applications because of their tendency to corrosion, particularly in the case of small wires operating in the presence of humidity and d-c voltage.

Pressure-sensitive tapes require no wetting or other treatment prior to application and need only to be placed in contact with the surface to which they are to adhere. Many different formulations for this general type of adhesive are now in use with nearly all having as the base, pure, virgin rubber. For many years there was only one type of adhesive used on the available pressure-sensitive tapes, and it was universally recognized that such tapes were subject to one serious fault – a tendency to loosen in the presence of solvents or heat. As a result of the recognized need for an adhesive which would not fail under these conditions, tape manufacturers set up a research program which resulted in the development of a new adhesive, still pressure-sensitive and possessed of all the good points of the previous formulations, but having a new property – that of being thermosetting.

It was originally felt that if a pressure-sensitive adhesive that would be low in solvent retention could be developed, such an adhesive would prevent loosening or unsticking in the presence of varnishes and other treatment materials. Thermosetting adhesives have this property and, in addition, cure – polymerize – under heat to form a bond which is but slightly affected by normal solvent action and which will remain solid up to about 200 C.

Thermosetting adhesives must be cured before attaining their maximum solvent and heat resistance. Curing is accomplished by heating the tapes at 120 C for two hours, or 150 C for one hour, after which they may safely be used at temperatures up to 150 C. Properly cured, these films will remain firm indefinitely and will exhibit a high degree of solvent resistance.

BACKING MATERIALS

The backing materials used on electrical-grade tapes fall roughly into four classes, paper, cloth, plastics, and glass. Selection of the proper type of backing material should be based upon the specific requirements of the application with due regard for the conditions under which the end product will operate.

Paper-backed tapes are characterized by moderate tensile strength, low insulation resistance, low dielectric strength, but only moderately good resistance to corrosion. In high frequency transformers, the principal use of this group of tapes is to hold wires against the form at the start of a winding or to hold a number of leads together prior to final assembly. For those cases where the tape is applied merely as a temporary holding device to be removed before the unit is completed, paper-backed tapes are perfectly satisfactory since it is the paper and not the adhesive that becomes the source of corrosion under conditions of humidity and voltage. If, therefore, the paper is removed, it is unimportant if a bit of the adhesive remains since it is impossible for acids or other ionizable material to develop from the non-cellulosic adhesive.

Cloth-backed tapes must be divided into two groups – those using cotton cloth and those made from acetate cloth. The first group, because of the presence of cellulose in the cotton, is poor with respect to corrosion. The acetate cloth tapes, on the other hand, show excellent resistance to corrosion – a characteristic which, in view of their relatively high tensile strength, good stretch and adhesion, very high insulation resistance, and good dielectric strength, makes them a wise choice for electronic applications. Cotton-backed tapes, like paper-backed tapes, should be used with caution, if at all, as a permanent part of r-f components. The chief uses of cloth-backed tapes are as insulation within windings, as a means of terminating windings where good tensile strength is a requisite, and as a cover for windings which must serve as the base for another winding.

Tapes with inorganic cloth backings, such as glass, are used in high temperature applications and also in those cases where exceptional strength is required. Used in combination with some of the newer high temperature impregnation materials glass tapes provide a means of protecting units against high operating temperatures. Surprisingly enough, glass-cloth tapes are not particularly good with respect to corrosion and therefore should be used only when absolutely necessary.

The type of backing material most commonly specified in electronics applications is one form or other of plastic material. Tapes of this sort are available in a wide variety of materials and thicknesses and are generally high in tensile strength, relatively high in adhesion, and possessed of exceptionally high dielectric strength and insulation resistance. No other type of backing material affords so high a degree of insurance against electrolytic corrosion which, in itself, is sufficient reason for the almost universal adoption of plastic tapes in the manufacture of high frequency transformers and coils and other components using the smaller sizes of magnet wire.

The plastic material most commonly used in the manufacture of tapes is cellulose acetate. Polyester films[1] as well as films of polyethylene, Teflon, and the vinyls, are also available for those applications where special properties not possessed by cellulose acetate are required.

ELECTROLYTIC CORROSION

Electrolytic corrosion is a major problem in the manufacture of radio frequency transformers. In the early days of the industry when tapes were first introduced as holding devices, it was noticed that the surgical and gummed paper tapes which were then the only types available induced corrosion to a serious degree, particularly in the case of small copper wires.

It was first believed that if steps were taken to produce a tape in which both the backing and adhesive were held to a pH of 7, it would be impossible for the tape to cause corrosion, even in the presence of high humidity and a d-c voltage. This experiment was tried, but it was found that chemically neutral tapes still corroded small copper wires. The next step in the development was an attempt to purify the tapes by leaching out with distilled water all extractable material. Tapes produced in this manner lessened the tendency to corrosion but did not entirely eliminate the trouble.

Since in studies of the corrosion problem it had been noted that the most serious corrosive effects always developed in the presence of a d-c potential, tests were next made on the various materials used in the manufacture of tapes to determine the effect upon them of the continued application of a d-c voltage in the presence of moisture. The checks were made upon materials which were in the most highly purified forms obtainable. In a vast majority of the cases, it was found that exposure to humidity and a constant d-c voltage resulted in the formation of organic acids, even though no indication of acidity had been present prior to the test. This finding led to the natural conclusion that these acids had been formed from the original organic material by electro-chemical action. It was further noted that all cellulosic materials such as paper, cotton, cellophane, etc., were especially subject to this type of reaction, while materials like cellulose acetate apparently had the cellulose molecule so thoroughly enclosed chemically that no acids were formed by the electrolytic action.

Years filled with continued experiments along these lines conducted by tape manufacturers and their customers in the electrical field have borne out the truth of these findings. There now seems to be little doubt that the presence of cellulose, no matter how well it may be concealed in a physical manner, in the presence of humidity and d-c voltage will decompose into organic acids of a highly corrosive nature. It is for this reason that paper-backed and cotton-backed tapes must be viewed in the same light as wood-filled phenolics and paper coil forms—as potential sources of corrosion to be avoided in critical military applications.

It is important for a design engineer to understand the basic principles involved in electrolytic corrosion since failure to view this problem from all angles can easily lead to unwise decisions adversely affecting quality of an end item. In its simplest concept, this type of corrosion may be considered as a form of electroplating. The requirements are identical in that there must be a source of d-c voltage, and there must be some form of electrolyte between two points which are at opposite potentials and which may be considered as an anode and a cathode. The general effect, as in plating, is a transfer of metallic particles from the anode (positive pole) toward the cathode (negative pole). Fig. 8-4 offers a quick review of the fundamentals of electroplating; hence, of electrolytic corrosion. In this drawing, a jar filled with an electrolyte such as copper sulphate, is shown containing two electrodes with the anode represented as a bar and the cathode as a rod. If the anode bar consists of pure copper and the cathode is of any conductive material, when a source of d-c potential is connected as shown in the diagram, copper ions will be dislodged from the anode and will make their way through the electrolyte to the cathode where they will be deposited. This is the identical process which, on an infinitely smaller scale, takes place whenever there is electrolytic corrosion within an electronic component. The anode is invariably of copper and may be a wire or a winding, usually connected to B+, while the cathode may be another wire or winding, a mounting bracket, or other conductive substance which is at negative potential and in relatively close proximity to the anode. Instead of an electrolyte such as copper sulphate, a leakage path made up of moisture and ionizable material connects the anode and cathode, thus setting the stage for the same sort of transfer of copper ions that takes place in electroplating.

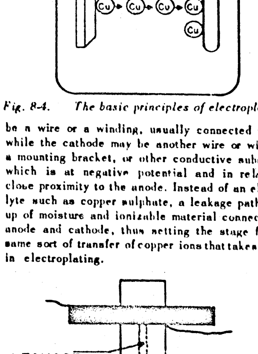

Fig. 8-4. The basic principles of electroplating.

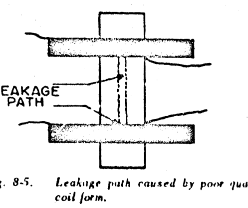

Fig. 8-5. Leakage path caused by poor quality coil form.

Fig. 8-5 shows a typical i-f transformer primary and secondary on a coil form whose quality is such that a leakage path exists between the two windings. Corrosion can therefore be expected along the lower surface of the top winding where it is in contact with the coil form. If, as in Fig. 8-3, a layer or two of electrical-grade cellulose acetate tape of a width substantially greater than the width of the winding is wrapped about the coil form beneath the plate coil, thereby insulating it from the form, a substantial degree of protection against corrosion will be introduced into the design. Another instance of electrolytic corrosion is illustrated in Fig. 8-6 where a leakage path is formed, not between windings, but between the plate coil and a mounting stud which is, of course, at ground potential. As in the other drawings, the point at which corrosion may be be expected is indicated by an arrow.

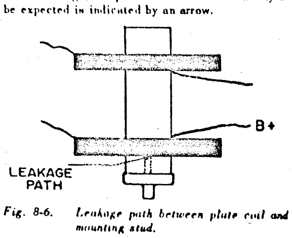

Fig. 8-6. Leakage path between plate coil and mounting stud.

This type of corrosion usually is accompanied by the presence of a green discoloration at the point where the action is centered. The exact nature of this deposit varies since it consists of copper salts formed as a result of electrolytic action between the copper and the ionizable material with which it is in contact. In this connection, it should be remembered that corrosion cannot take place in complete absence of moisture. The reason for this is simply that ionization is necessary before electrolytic action can begin, and ionization cannot occur except in the presence of moisture. This fact points out the necessity for thorough impregnation of all windings and assemblies, not only to insure satisfactory performance of the coils under humid conditions but also as a means of retarding electrolytic corrosion. If it were possible to seal a transformer so thoroughly as to prevent the entrance into any of its component parts of even the smallest amount of water vapor, the cheapest materials, even including paper and cellophane, could be used with complete

[1]Mylar manufactured by E.I. du Pont de Nemours & Company Wilmington, Delaware

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safety. Unfortunately, such perfect sealing cannot be readily accomplished, even with the aid of encapsulation — hence the need for careful selection of materials which are not subject to ionization.

A point of importance in this discussion is the fact that the general problem of corrosion increases as wire size decreases. This should be obvious since the process is one of erosion, and the less there is of material to erode, the faster the point of failure will be reached. It therefore follows that the present trend toward miniaturization of components, coupled with the expansion in environmental range, in which these units must work, places heavy demands upon those who must design components incorporating a high degree of reliability.

CORROSION TESTS

For the most part, it will be found that the recommendations and specifications of reputable tape manufacturers can be accepted with safety, especially if the tapes in question have been granted JAN approval. In the course of normal development work there may, however, come times when new or untried combinations and/or arrangements of materials make it advisable to conduct corrosion tests before proceeding with the work.

The requirements for determining resistance to corrosion are relatively simple since they center around a container so constructed as to be able to hold a specified degree of relative humidity and temperature. Successful tests have been run using a glass jar fitted with an insulated top and kept in an oven equipped with temperature controls and a circulating fan. The container must have two terminals to serve as connections to a source of constant d-c voltage of a magnitude which is usually somewhere between 100 and 250 volts. Test conditions will vary with circumstances, but 95 ±1 percent relative humidity and 120 ±1 F are most often specified.

The exact method of connection must be determined by the design of the component under test. It is customary to duplicate as nearly as possible the conditions under which the unit will operate, which is to say that the positive lead should go to the plate terminal and all other parts of the unit including its shield, mounting brackets, and the grid coil should be tied together and connected to negative. In those cases where only a coil form with its two windings is under test, one coil should be connected to negative and the other to positive.

It will be apparent that a unit connected as described above will not, under normal circumstances, draw any current since the only connection between positive and negative is that resulting from leakage. If there is no leakage, no current will flow, there will be no migration of copper ions, and there will be no corrosion. Let, however, ionization take place and leakage begins at once, bringing with it the movement of copper away from the positive and making this transfer evident by growth of the characteristic green deposit at the point where electrolytic action starts.

In setting up a test of this sort, absolute cleanliness is a requirement that cannot be disregarded. It must be remembered that the entire procedure is one aimed primarily at the detection of ionizable matter. If, therefore, the wire, the insulation material under test, or any part of the holding devices or the container are handled with bare hands, it is entirely possible that enough contamination will be introduced to render the whole test worthless. Clean rubber gloves should be used at all times, and a constant guard maintained against the introduction of any foreign matter. It is only in this manner that one can be certain of the results obtained from an electrolytic corrosion test.

Corrosion tests may or may not continue until failure occurs. Generally speaking, the average duration of test is 160 to 200 hours, with visible evidence of corrosion — presence of a green deposit adjacent to the positive wire — considered as evidence of failure even though the circuit may not be open. A system for evaluating the results of test not ending in failure is described in detail in Appendix 1 of ASTM Bulletin D 1000-48T. The basis for this evaluation is a comparison in the tensile strength of the positive wire and either the negative wire or a piece of identical wire not subjected to test. The tensile strength of the positive wire divided by the tensile strength of the negative wire gives a term known as the *corrosion factor*, which is used as a means of quantitatively comparing various materials in their resistance to corrosion.

DIELECTRIC STRENGTH

Dielectric strength is a property of electrical-grade tapes which is of extreme importance in those applications where tape must provide insurance against voltage breakdown. By definition, dielectric strength is the lowest voltage at which the insulation will break down, and the common unit of measurement is volts per mil thickness. Determination of dielectric strength is usually made in accordance with the requirements of MIL-I-7798 and ASTM Specifications D 1000-48T and D 149. Tests obtained by methods other than those outlined in the above-referenced specifications are difficult to evaluate since both the length of time that the test voltage is applied and the shape and size of the electrodes used in applying it have an effect upon the outcome of the test.

The manner in which the dielectric strength of an organic insulating material may be related to time is shown in the following table:[2]

1/1,000,000	second	105,000 volts
1/10,000	second	92,000 volts
1/100	second	78,000 volts
1	second	65,000 volts
100	second	51,000 volts
10,000	seconds (3.78 hours) ...	37,000 volts

Thus it appears that the dielectric strength of a typical organic insulating material decreases as the time interval during which the voltage is applied increases. Experimental work performed on insulating materials indicates that dielectric strength will continue to fall off as the time increases until that voltage is reached where the insulation will resist breakdown indefinitely.

Among electrical-grade tapes, the highest dielectric strengths will be found in those having plastic backings. One prominent manufacturer[3] lists a choice of acetate or polyester backings with dielectric values as high as 5500 volts. While it is true that high frequency coils seldom require protection against excessively high voltages, it is reassuring to know that the tapes best suited for use in such components are more than adequate in this respect.

STORAGE AND HANDLING

For best results, pressure-sensitive tapes should be handled properly. One important point concerns the storage of unused tapes since this type of material has a very definite storage life. Care should be taken not to buy in quantities so large that tape must remain in stock for more than three to six months. Storage areas should be kept below approximately 85 F and should not be exposed to direct sunlight. Rotation of stock is important because through its continued practice the oldest material will be used first. For the convenience of users, most tape manufacturers date their containers.

Most electrical tapes are supplied in the form of rolls wound on fibre or paper cores with diameters of between 1 and 3 inches. The length of tape per roll varies, but most commonly is either 36 or 72 yards. The most commonly available widths fall between 1/4 inch and 5 inches with the tape usually cut to an accuracy of at least 1/32 inch. Some users prefer to buy their tape in wider widths and do their own slitting as a means of reducing cost and inventory.

Tape dispensing has much to do with the efficiency of tape as a tool. Used without some sort of special dispensing equipment, pressure-sensitive tape is often neither easy nor economical to apply. Most tapes resist tearing to an extent which interferes seriously with smooth application unless some cutting device is used. Many different forms of dispensers are available with probably the simplest being slotted boards or drums on which tape is wound and then cut into desired lengths by running a knife through the slots. After being cut, the individual pieces of tape can be picked off, either by tweezers or by the fingers.

Several types of tape dispensers, now commercially available, will take one or more rolls of any pressure-sensitive tape and by one simple lever motion deliver accurately cut pieces of uniform length. Use of this sort of equipment insures that the same amount of tape will be used each time — a means of forced economy as well as a point of importance in certain electronic applications, such as those where thickness of tape has a bearing on the coupling between two windings. A further point to be considered as favoring the use of dispensing equipment is the assembly time saved by having a pre-cut tape available at the time and place that is required with the tape in full possession of its maximum adhesive power as a result of having been handled only as it was removed from the dispenser.

SELECTION OF PROPER TAPE

As in the case with nearly every component part of a high frequency transformer, the choice of the best tape for a particular application is dependent upon many factors. The dangers of electrolytic corrosion were pointed out earlier in this discussion and were intended to stress the importance of specifying only electrical-grade tapes. A good rule to follow would seem to be to avoid the use of tapes with paper, cellophane, or cotton backings and to consider only those backings which offer maximum resistance to corrosion.

Polyester and acetate films, acetate cloth, and acetate film cloth are unquestionably the best all-round tapes for radio frequency components. In the case of units which must operate at ambient temperatures substantially in excess of 100 C, the use

[2] Based on information provided by T. J. Skotnicki of Minnesota Mining and Manufacturing Company, Saint Paul, Minnesota.

[3] Minnesota Mining and Manufacturing Company

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Fig. 8-7. CHARACTERISTICS OF COMMON ELECTRICAL – GRADE TAPES***

Tape No.	Varnish Resistance	A.I.E.E. Temp. Classification	1 Mc. at 23°C and 50% R.H. Dielectric Constant	Dissipation Factor
1	Poor	A	2.6	.025
3	Poor	A	3.5	.025
5	Poor	B*	3.1	.02
6	?	A		
7	Poor	A	3.5	.03
8	?	A	4.3	.02
9	Poor	A	3.2	.025
10	Poor	O**	2.4	.013
11	?	O**	2.5	.015
26	Excellent	0**	3.4	.020
27	Good	B	2.9	.006
28	Fair	0**	2.4	.014
38	Good	A	2.4	.039
39	Good	A	3.7	.037
45	?	A	3.4	.025
56	Good	B*	2.5	.013
PTF-1S	?	H	2.4	.0014

* Adhesive is Class B but backing may be deteriorated by certain chemicals in ambient Atmosphere at Class B Temperatures.

** Without varnish impregnation. Class A if impregnated.

*** Information supplied by Minnesota Mining & Manufacturing Company and the numbers refer to their particular products.

of polyester (Mylar) backings will be found safe up to 125 to 135 C, while higher temperatures indicate a need for tapes with a backing of Teflon film.

In general, the use of tapes having thermosetting adhesives is recommended for use in electronic application. Their increased resistance to solvent action and their lessened tendency to loosen under elevated temperatures are properties which characterize these tapes, thus giving them, in most instances, a distinct advantage over normal pressure-sensitive tapes.

FILM INSULATIONS

Of lesser interest to the average designer of high frequency transformers but nevertheless of value in certain applications are film insulations having no adhesive of any kind. Films of this sort are usually around 0.001 to 0.003 inches in thickness, although they may be obtained over a thickness range of from 0.00025 inch up to several thousandths of an inch. In the design of radio frequency transformers, their principal use is to furnish insulation between an assembly and its shield. Space requirements often make it necessary to provide insulation for the full length of a shield can, something that would not be entirely practical to do with an adhesive tape, but which can be done with a minimum of trouble by the use of a flexible film of high dielectric strength.

Among the materials used in the form of thin, unsupported films is the familiar cellulose acetate which is most satisfactory as a backing film in tapes. Comparatively free from solvent attack, acetate films are dissolved only by the more active solvents such as acetone, carbon tetrachloride, perchloroethylene, and other esters and ketones. Acetate films may be operated with safety up to about 100 C, and if protected with silicone impregnants, at even higher temperatures. Available in a wide range of thicknesses and possessing good

dielectric strength, cellulose acetate films have a dielectric constant usually somewhere between 4.0 and 5.0.

A relative newcomer to the field of film insulations, but one which threatens to replace acetate films in a majority of electronic applications, is the polyester film marketed under the trade name of Mylar.

Mylar is an extremely tough film which is hard to tear or crease and which has lower moisture absorption and higher heat resistance than conventional acetate films. The fact that it is considered safe to operate Mylar continuously at 135 C makes it of special interest to designers faced with constantly rising ambient temperatures.

Polyester film is presently available in thicknesses between 0.00025 and 0.0075 inches. At least two companies[4] are using Mylar film for tape backing, while its special properties, particularly its strength and moisture resistance, have led other companies[5] to employ this film in combination with paper, glass, asbestos, vulcanized fibre, and many other materials. Many of these Mylar composites have interesting properties, and it is reasonable to assume that future coil developments may include materials of this sort.

Of some interest in the high frequency transformer field are the insulating films made up from mica splittings held together by an adhesive or bonding agent. Mica papers, as they are called, are available in thicknesses beginning with 0.005 inch and extending upwards to 1/8 inch. The thinner sheets have a high degree of flexibility—a characteristic which has encouraged use of this material in the form of rolled tubes impregnated with silicones for the coil forms in transformers which must operate at high temperatures.

Generally speaking, the type of bonding material used in making these flexible sheets has as its base, natural resins or gums. As a result, these materials are called Class B insulations and are therefore recommended for use only up to 130C. Since about 80 percent of the film consists of mica which is capable of withstanding temperatures up to almost 1000 F, it can be seen that the substitution of a silicone or other binder less affected by temperature would greatly increase the safe operating range of mica papers.

Also of interest among high temperature insulations are the films made from polytetrafluoroethylene. The excellent electrical properties and extremely high resistance to moisture and temperature have created much interest in Teflon—the name by which polytetrafluoroethylene is commonly known.

Primarily a molded or fabricated material, Teflon has proved to be difficult to produce in the form of thin films. For some time, the only method by which Teflon films could be made was by shaving or skiving blocks of the material. Films thinner than 0.002 inch are extremely difficult to produce by this method and have the basic defect of containing large numbers of pinholes.

Considerable work has been done in an attempt to produce Teflon films by casting and or extrusion. Processes have been developed by which films of from 0.00025 to 0.002 inch can be cast from an aqueous dispersion, dried, and finally sintered into a continuous strip. Films made in this manner have far fewer pinholes than do skived films. Extrusion methods have also been successful in that two different types of film — one fused and one fusible — are being produced in commercial quantities. Thus far, applications are few in high frequency transformers, but most engineers are showing a decided amount of interest in both forms.

In general, Teflon films are characterized by a low dielectric constant (approximately 2.0), low power factor (average 0.005), no measurable water absorption, high dielectric strength, and the ability to operate continuously at 150 C. In addition, they are exceptionally high in solvent resistance and are nonflammable.

Many other film insulations are available and, while not of general interest to designers of radio-frequency coils and transformers, are deserving of mention in this discussion. Among these materials may be listed Nylon, polystyrene, polyvinyl chloride, vinylidene chloride, and certain of the epoxy resins. Each of these insulations has its good features as well as those in which it is inferior. For the most part, it seems that Mylar or the well-known cellulose acetate film will most often be specified in the design of a high frequency transformer. This statement is not meant to dismiss the possibility of successfully using one of the less common films but rather to point up the wisdom of working with materials known to be good, while, at the same time, becoming familiar with those which may eventually prove to be better.

[4]Minnesota Mining and Manufacturing Company and Permacel Tape Corporation, New Brunswick, New Jersey

[5]Irvington Varnish Division, Minnesota Mining and Manufacturing Company, Irvington, New Jersey, Natvar Corporation Woodbridge, New Jersey; John Roebling's Sons Company, Trenton, New Jersey.

BIBLIOGRAPHY

Griffeth, R.L. and Younglove, E.R.
"Processing Mica Paper for Electrical Insulation,"
Electrical Engineering, May, 1952

Javitz, Alex, E.
"New Nonrigid Materials for the Functional Design of Electrical Insulating Systems,"
Electrical Manufacturing, September, 1953.

"Research Progress in Dielectrics"– 1951
Electrical Manufacturing, January, 1952

"Research Progress in Dielectrics"– 1952
Electrical Manufacturing, December, 1952

Moses, Graham Lee
Electrical Insulation, First Edition
McGraw Hill Publishing Company, New York, 1951

CATALOGS AND TECHNICAL INFORMATION OF:

The Acme Wire Company
New Haven, Connecticut

Bauer & Black
Division of The Kendall Company
2500 S. Dearborn Street
Chicago 16, Illinois

The William Brand & Company, Inc.
Willimantic, Connecticut

The Connecticut Hard Rubber Company
407 East Street
New Haven 9, Connecticut

Industrial Tape Corporation
New Brunswick, New Jersey

The M.W. Kellogg Company
P.O. Box 469
Jersey City 3, New Jersey

The Rex Corporation
Hayward Road
West Acton, Massachusetts

John Roebling's Sons Company
Trenton, New Jersey

Mica Insulator Company
Schenectady 1, New York

Minnesota Mining and Manufacturing Company
900 Fauquier Avenue
Saint Paul 6, Minnesota

Natvar Corporation
Woodbridge, New Jersey

Owens-Corning Fiberglass Corporation
Electrical Sales Division
16 East 56th Street
New York 22, New York

The Polymer Corporation of Pennsylvania
Reading, Pennsylvania

Resistoflex Corporation
Belleville 9, New Jersey

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SPECIFICATIONS

HH-I-538
"Insulation, Electrical, Pasted-Mica"

JAN-T-638
"Tape, Insulating (Electrical), Linen-Finish, Plain"

MIL-I-631A
"Insulation, Electrical, Synthetic-Resin Composition, Nonrigid"

MIL-I-1140A(1)
"Insulation, Electrical, Glass-Fiber, Untreated"

MIL-I-3158(2)
"Insulation, Electrical, Glass-Fiber, treated, Cloth, Tape, and Cordage (Resin Filled)"

MIL-I-3389A
"Insulation Cloth and Tape, Electrical, (Rayon, Untreated)"

MIL-I-3393
"Insulation, Cloth and Tape, Electrical, (Nylon-Fiber Varnished)"

MIL-I-7798
"Insulation Tape; Electrical, Adhesive, Plastic"

MIL-T-15126
"Tape, Insulating, Electrical, Pressure-Sensitive, Adhesive"

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POOR ORIGINAL

Section 9

FINISHES AND MARKING

HISTORICAL BACKGROUND

Prior to World War II it was not customary to attach too much importance to protective finishes for r-f transformers and their components other than to those materials used in the impregnation of the coils. To be sure, it was recognized that brass under certain conditions was subject to corrosion, and it was therefore assumed that brass parts should be electroplated or tinned, and iron and steel were known to require some sort of protective coating if serious rusting was to be avoided. It was, however, just about at this point that interest in and concern for other protective measures ceased to exist.

It was largely as a result of this situation that the Armed Forces, in the early days of the War, were confronted with a series of what at first were felt to be unexplainable equipment failures. Most, but not all, of these breakdowns seemed to originate under conditions of high humidity and elevated temperatures, and it was noted that inoperative equipments usually were accompanied by an accompanied by an accumulation of corrosion products or crystalline formations. Other instances of equipment failure were traced directly to the action of certain fungus growths which, in turn, were noted as being associated only with certain specific materials of construction.

It is not surprising that out of the many new developments resulting jointly from the War Effort and the varied environmental conditions under which equipments were used, that a new concept was developed regarding those protective measures necessary to deal with these new forms of failure-inducing organic and inorganic growths.

Today it is generally accepted that protective finishes are a necessary consideration in the design of high frequency transformers and among the reasons may be listed:

1. To prevent corrosion
2. To improve appearance
3. To provide (or maintain) good electrical contact
4. To reduce r-f losses or to improve Q
5. To facilitate soldering
6. To prevent growth of fungi
7. To prevent degradation of insulating material by moisture

The common types of finishes used (and it should be understood that this discussion will not include coil impregnation materials which were covered in Section 7) include electroplated and chemical-film coatings and paint finishes. A large number of military and government specifications cover the specific features of the various finishes. Since these specifications are frequently amended and sometimes may even appear to be conflicting, the reader is cautioned to check the current amendment of all applicable specifications whenever starting a new project.

Fortunately for the designers of most electronic inductive components, the degree of protection required inside of end equipments is not great. For this reason, electroplated and chemical-film finishes are generally found satisfactory. Certain special cases or a definite requirement for improved appearance sometimes creates the need for a painted surface, but it must be admitted that these cases are relatively uncommon.

CHEMICAL AND ELECTROPLATED FINISHES

The importance of the general subject of chemical and electroplated finishes is indicated by the fact that only rarely are metallic components used without having received some sort of chemical or electrochemical treatment prior to the completion of the unit of which they are a part.

Probably the most common protective process

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is one type or other of *electroplating*. Almost any metal can be plated under the influence of a direct current following the basic principles set forth in Section 8 of this manual and illustrated in Fig. 8-4. Particularly in civilian applications, the plating most often specified is cadmium, with zinc and silver appearing in special cases.

It is not intended here to cover the subject of electroplating in any great detail since only rarely need a coil engineer be directly concerned with the plating process. However, as has been indicated in other sections of this manual, it is a primary purpose of this work to supply general background information that may prove helpful in transformer design. For this reason, attention is called to the fact that electroplating is a process whereby metallic, or in certain specific instances, nonmetallic parts are coated with a metal which is deposited from an electrolyte as a result of ionization. It is important to remember that electroplating is accompanied by an increase in the size of the piece, and therefore plating thickness must be taken into consideration when designing mated or threaded parts. To be sure, the thickness of plating seldom exceeds 0.0005 inch and is more often of the order of 0.0002 or 0.0003 inch, but the fact remains that there always is a dimensional change as a result of electroplating.

In plating spring-steel parts with cadmium, a trouble known as *hydrogen embrittlement* is frequently encountered. This can be most serious as springs so affected will fail (snap) without warning. Usually apparent immediately following the plating process, hydrogen embrittlement occasionally will not show up until a considerable interval of time has elapsed. To avoid trouble, it should be clearly specified that all spring-steel parts which must be cadmium plated are to be placed in a 250 F to 275 F oven for at least 30 minutes immediately following removal of the part from the plating bath. In this manner, the serious effects of hydrogen embrittlement may be largely avoided although this condition can never be completely disregarded as a potential source of trouble when spring-steel parts must be electroplated.

An operation which is frequently made a part of electroplating procedure is that which involves the use of an acid solution known as *bright-dip*. As the name suggests, this treatment increases the luster of the surface, leaving it bright in appearance as well as chemically clean. A feature of the bright-dip process which is important to remember is that it is essentially an etching process which actually removes metal from the parts being treated. It therefore follows that closely fitted or threaded portions must be carefully designed and carefully treated[1] if uniform fits are to be maintained.

CHEMICAL FINISHES

Among the many chemical finishes which are used on electronic components may be listed *anodizing* – a finish sometimes applied to aluminum. The process of anodizing is an electrolytic one in which a protective coating or film of aluminum oxide is formed on the surface of the metal. Because this oxide film is an electrical insulator which seriously interferes with the grounding of treated parts, anodizing is seldom applied to the components of radio frequency transformer.

A second form of chemical treatment for aluminum is much more common, particularly in the case of shield cans. Known as *caustic etch*, this process is one involving an alkaline solution which attacks the surface of the aluminum leaving it clean and with a matte surface. All shield cans made from 2S, 3S, or 52S aluminum should receive a caustic etch treatment since its use provides a surface which is highly conductive as well as one which takes ink well, thus opening a way to legible marking.

A coating which is of considerable importance, particularly in the case of brass solder lugs and terminals, *is hot tin dip*. While a comparatively expensive coating, hot tin dip has the advantage of forming an easily solderable surface which is not subject to oxidation. The process whereby the coating is applied is relatively simple, consisting primarily of dipping cleaned, fluxed parts in a bath of either molten tin or a tin-lead (solder) alloy.

While in general ferrous parts are not particularly important in high frequency transformers, it does seem wise to include a reference to *phosphate coatings* for iron and steel. Essentially, these processes, of which *Parkerizing* and *Bonderizing* are examples, develop surface coatings of insoluable, nonconductive phosphates. Both of the above mentioned coatings are applied by dipping the parts in a hot solution for time intervals varying from 3 to 45 minutes. Bonderizing is widely used as a base for paints and lacquers since it facilitates adherence and also tends to retard rusting in the event of deep scratches which cut through to the bare metal. Parkerizing provides a somewhat heavier, more corrosion-resistant coating than does Bonderizing, but neither method is recommended as the sole protective finish for any component. A feature of phosphate coating is the high resistance surface film which results from the process.

In any general listing of chemical finishes, mention probably should be made of *black oxide coatings* and *passivation*, although admittedly, neither is of major importance in coil design.

Black oxide coatings consist of a very thin film of black iron oxide formed by chemical action on the surface of ferrous parts. While affording only limited protection against corrosion, such films form an excellent base for rust inhibiting oils and also give a very satisfactory black color to treated parts.

Passivation, as is suggested by the name, is a chemical process involving the use of nitric acid which renders the surface of certain steels, including those of the stainless type, inactive (passive) chemically. The degree of protection against corrosion is, in itself, not great, but in combination with that which is inherent in stainless steels, it has been found sufficient for certain limited applications.

By far the most important of the chemical dip treatments from a military point of view are the *chromate treatments*. Used on zinc and cadmium surfaces, chromate treatments result in the formation of a chemically complex, chromate type, corrosion resistant film. This film is very thin, being approximately 0.00002 inch in thickness, and thus has no effect on the fit of mating or threaded parts of close tolerance. A further advantage of this type of film coating is found in the gel-like structure of the film which makes it particularly resistant to cracking or separation from the metal.

It is possible to have chromate films which are olive drab, bronze, or transparent in appearance. An interesting feature of this particular type of finish is the manner in which the film duplicates the original surface lustre of the metal. In other words, if a bright plate or a polished surface is treated with a chromate finish, the resulting surface will be highly lustrous, while the same film applied to a dull surface will produce a dull finish. In addition to having color inherent in the process, chromate films may be dyed by the use of Alizarine or Diago series dyes, thus opening a way to the formation of almost any desired color.

Further features of chromate films include low electrical resistance – a property allowing this type of finish to be applied to parts which are used as electrical contacts. Certain types of chromate films make soldering somewhat more difficult, but a special type of treatment is available which when used on cadmium plated parts, permits easy soldering with resin type fluxes. Chromate films show a high degree of adhesion to the metal and will withstand drawing or forming processes to the same extent as will the metal surface to which the film is applied.

Corrosion resistance is generally improved by the use of chromate coatings, but by no means are these finishes to be considered as completely effective in this regard. The best surface protection is afforded by those films which are bronze or olive drab in color, with those which are either dyed or transparent being of lower protective value. As a basis for paint finishes, chromate films are especially good. Not only do they permit excellent adhesion, but in combination with the paint they provide a high degree of protection for the surface of metals to which they are applied.

It should be noted that chromate films do not provide protection against the corrosion of cadmium or cadmium plate which is packaged or confined in a damp atmosphere without air circulation in the presence of acidic organic vapors.

Metallic single crystals (trichites) may grow unassisted at room temperature from solid metal or plated surfaces such as tin, zinc, and cadmium. Complete failure of an equipment can result because of a low impedance or short circuit caused by trichites at a critical point in a circuit.

Frequently it will be found helpful to have knowledge of simple tests by which to identify electroplated films, base metals, and the presence of chemical type protective films. This subject has been thoroughly covered in Technical Manual No. M-1379 entitled *Inspection Kit for Finishes* written by Aubrey J. Raffalovich and available from the Signal Corps Engineering Laboratories at Fort Monmouth, New Jersey. This publication lists the reagents, the equipment, and the procedures – all simple and easy to understand – by which the various finishes and base metals can be positively identified.

[1] This subject is well covered in Information Bulletin No. 180, written by Mr. H. M. Gade. This booklet is available from the Components and Materials Branch, Signal Corps Engineering Laboratories, Fort Monmouth, New Jersey.

PAINT AND ENAMEL FINISHES

Paints are organic surface coatings made up primarily from drying oils and pigments, while *enamels* usually consist of a synthetic resin in combination with appropriate solvents, drying oils, and pigments. The general composition of these and other finishes appears in Fig. 7-2 of this manual.

In military equipment, organic finishes may be specified as a means of providing metallic surfaces with improved corrosion or weather resistance as well as with improved appearance. Since, however, this discussion is limited to high frequency transformer design techniques, and since components of this general type rarely if ever are located in other than enclosed areas such as the interiors of radio or radar sets, it is very seldom that painted finishes will be specified for use on these units.

For those occasions on which shield cans, for example, must receive a coating of paint on the surfaces, the procedures outlined in the chart appearing as Fig. 9-2 will usually be found satisfactory. Because a film of paint or enamel can be effective only if it adheres firmly to the surface being covered, it is customary first to treat the base metal with some sort of a chemical treatment and then to apply a zinc chromate primer before attempting application of the paint itself.

Two more factors are of importance in those cases where paints are used on the components of r-f coils. A film of paint or enamel is a good insulator which means that provisions must be included to allow for low resistance electrical connections whenever required. These may involve masking, the attachment of special ground straps, or the use of washers which pierce the outer insulating film and give contact with the base metal.

Color is often a problem in the use of paint. Federal specification TT-C-595, *Colors for Ready-Mixed Paints* offers a convenient and generally satisfactory means of specifying desired colors.

MARKING

Part numbers, ratings, terminal identifications, etc. are most commonly marked on coils and high frequency transformers by one or more of the following methods:

1. Rubber stamping
2. Silk screening
3. Die stamped or molded markings
4. Decalcomanias
5. Paper labels
6. Color code dots for terminals

Specification No. MIL-M-13231 entitled *Marking of Electronic Items* specifies certain tests for the permanence of markings which are required to withstand thermal shock, the solvent action of carbon tetrachloride, gasoline, and soap and water, to resist corrosion and abrasion, and not to exceed certain limits on flammability. For the exact methods of test, the reader is referred to the specification.

For all markings involving the use of ink, paint, cement, or other mediums which are based on adhesion, the surface to which they are applied must be free from wax, grease, dirt, or other foreign matter. Prior to marking, materials must be cleaned when necessary by wiping with a solvent such as alcohol, by vapor degreasing, or by the use of detergent cleaners. Because many types of markings when applied to insulating materials tend to degrade the insulating properties, it is usually preferable to mark on adjacent surfaces. If the marking must be placed on the insulation, it is desirable to locate the marking away from surface leakage paths across the insulated parts. Die stamped or molded markings on insulations are, of course, the exception to the above rule since no ink or other material is commonly used with this type of marking. It is customary on military components to protect markings against abrasion and the effects of fungus by coating them with clear fungicidal varnish.

RUBBER STAMPING

Rubber stamping is the simplest and probably the most widely used method of marking electronic components. Legibility is somewhat related to type size with 1/8 inch or larger characters giving best results although 1/16 inch characters can be successfully rubber stamped if necessary. Small quantity runs of components are most often hand stamped, while larger production quantities can be handled more efficiently with a jig which allows quick and accurate location of the stamp against the work. At least one American manufacturer[2] supplies automatic marking machines which operate on the rubber stamp principle. Almost any size or shape of article can be successfully stamped by machine if production entails sufficient quantities to make the process worthwhile.

Rubber stamps are sometimes used with liquid

[2]The Markem Machine Company, Keene, New Hampshire.

inks applied from a stamp pad. The more common practice involves a paste ink – usually termed printer's ink – rolled on a slab from which the stamp is inked. The advantage of paste ink is two-fold, it being more permanent than liquid ink and also providing better legibility to the stamping. Rubber stamps will, if kept clean, usually last for more than five thousand impressions before replacement becomes necessary. Considerable work is being done on the development of vinyl stamps as a replacement for rubber. Advantages claimed for this material include easier cleaning and much longer stamp life.

The markings applied by rubber stamps may be allowed to air dry, or if time is a factor, the markings may be force dried at temperatures up to 200 degrees F. Both ovens and infrared lamps will be found satisfactory sources of heat for forced drying.

SILK SCREENING

Silk screening is that process whereby a special enamel-like ink[3] is applied to a surface through a fine-mesh cloth or metallic stencil. The stencil or "screen", as it is termed, may be prepared photographically by filling all areas of the screen except where the markings are located, or it may be prepared by cutting away with a knife those portions of the screen which form the desired symbols. The part which is to be marked is then jigged in position below the screen – usually at a distance which just clears the under surface of the screen – and the rather thick ink is then forced through the screen with a rubber squeegee blade. This method can reproduce with accuracy a far more complex marking than is possible with rubber stamping, but it has the disadvantage of involving a relativly high cost for the screens themselves as well as the necessary lead time involved in making the stencil. Once the markings have been applied, they can be dried in the same manner as those applied by rubber stamps.

DIE STAMPED OR MOLDED MARKINGS

Die stamped or molded markings are economical only in large quantities where the markings can be incorporated in the tool producing the parts. (See Section 6) It is obvious that this system of marking does not lend itself to changes since any change in marking is necessarily accompanied by a tool change. Die stamping produces depressed characters and may be used on metal parts or on plastic parts punched from sheet stock. Molded markings may be either depressed or raised – in other words, can appear above or below the surface of the material – and are applicable to molded plastic or ceramic parts. Raised characters are more satisfactory from the standpoint of mold life. In cases where improved legibility is required, depressed lettering can be filled with a thick paint and the excess wiped off.

DECALCOMANIA

Decalcomania markings provide excellent legibility and appearance. The application of a "decal" usually involves the application of a coat of special cement to the surface where the decal is to be located. The decal is then moistened, freed from its backing paper, and slipped into place where it must dry for a matter of several hours. It is customary for the manufacturers[4] of decalcomania markings to specify the exact process of application and to supply the cements required.

PAPER LABELS

Printed paper labels cemented in place provide another means of marking that is occasionally used, particularly for identification during the manufacturing or development process. Paper labels with pressure sensitive adhesive are commercially available, but the adhesion provided by these labels is inclined to be inadequate for military applications where permanence of markings is a requisite. In general, the use of paper labels is not looked upon with favor by military designers.

COLOR CODE

A convenient means of identifying terminals which is widely used in commercial practice consists of the application of small dots of *color coding* lacquer adjacent to each terminal. Colors used correspond to the colors of the hook-up wires that will connect to the particular terminals with the standard color code being as follows:

1. *Green* for grid terminals
2. *Yellow* for grid return terminals
3. *Blue* for plate terminals
4. *Red* for B+ terminals

Color code dots can be applied either by machine or by hand and constitute a fast and positive

[3]Such as "S. S. Unitrade Silk Screening Paint" supplied by Union Ink Company, Ridgefield, New Jersey.

[4]Such as Meyercord Company, Chicago 44, Illinois, and Palm Brothers Decalcomania Company of New York.

method of identifying terminals. Suitable lacquers for this purpose are available from a number of American manufacturers*.

CONCLUSION

While admittedly there are many other methods of markings, the above discussion has centered around those methods considered most applicable to electronic components, particularly high frequency transformers. Probably close to 90 per cent of all such components are marked either with rubber stamped part numbers, color coded terminals, or die stamped or molded terminal numbers. The information which is to be stamped upon the shield can or otherwise included in the finished unit constitutes a definite part of the specification. For reasons of legibility, markings should be kept as simple as possible – standard commercial practice generally consisting of inclusion of the part number, the RETMA* identification code number, and possibly the name or trade mark of the manufacturer.

*Examples are "WEQ Lacquer" manufactured by Lacquer and Chemical Company, Brooklyn 32, New York, and "Dyecart Lacquers" manufactured by The Mass and Waldstein Company, Newark, New Jersey.

*Radio Electronics Television Manufacturers Association

Fig 9-1 TABLE 1 ELECTROPLATED AND CHEMICAL-FILM FINISHES FOR METAL PARTS

Base Metal	Application	Finish	Remarks
Aluminum	Shield cans, hardware, and small parts	Caustic etch for shields made of 2S, 3S, or 52S aluminum – especially where electrical contact is of importance. See Specification 72-53 Amendment 5.	Most often specified
		Chemical dip treatment per Specification MIL-C-5541 such as Iridite No. 14 (1)	Better protection than caustic etch
		Anodize per Specification MIL-A-8625. *Not recommended*	This finish is an insulating film which prevents or complicates electrical grounding
Steel	Shield cans, unthreaded hardware, and small parts	Cadmium (preferred) per QQ-P-416 or Zinc plate per QQ-Z-325 0.0003" thick minimum followed by a chromate chemical dip treatment such as Iridite No. 8-P (1)	Most generally specified
	Threaded parts, nuts, and screws	Same as above except use 0.0002" thickness of plating to prevent thread interference.	
	Shield cans	Silver plate 0.0003" thick minimum over 0.0002" thick copper plated undercoat	Corrosion protection poorer, but may reduce r-f losses in the can
Brass, Copper, Phosphor-Bronze	Shield cans, small parts, nuts, and screws	None except use "bright-dip" when parts are initially dull	Gives satisfactory corrosion resistance without plating
	Shield cans, unthreaded hardware, small parts	Cadmium (preferred) or Zinc plate 0.0003" thick minimum followed by a chromate chemical dip treatment such as Iridite No. 8-P (1)	Better protection than unplated. Used primarily to avoid galvanic couples. (Note galvanic corrosion chart in Specification 72-53 Amendment 5)
	Threaded parts, nuts, and screws	Same as above except use 0.0002" thickness of plating to prevent thread interference	
	Solder lugs and parts to be soldered	Hot tin dip	Best finish for solderability

4

Fig. 9-1 TABLE 1 ELECTROPLATED AND CHEMICAL-FILM FINISHES FOR METAL PARTS (cont)

Base Metal	Application	Finish	Remarks
Brass Copper, Phosphor-Bronze	Solder lugs and parts to be soldered	Electro-tin plate, 0.0002" thick minimum and fuse coating	
		Silver plate 0.0002" thick minimum	Used for lower r-f losses.
Zinc	Shield cans, small die-cast parts	Chromate chemical dip treatment such as Iridite No. 8-P (1)	Zinc cans were once generally used but are now displaced by aluminum
Stainless Steel	Small parts, hardware	Passivation	

(1) Iridite products are manufactured by Allied Research Products, Inc. of Baltimore 5, Maryland, American Chemical Paint Company of Ambler, Pennsylvania, supplies similar products.

Fig. 9-2 TABLE 2 PAINT FINISHES FOR SHIELD CANS

Base Metal	Surface Treatment	Primer	Finish
Aluminum	Chemical dip treatment per Specification MIL-C-5541 such as Iridite No. 14 (1)	Zinc chromate primer per Specification MIL-P-6889	Color optional – Either lusterless enamel per Specification TT-E-527 or Semi-gloss enamel, grade I, per U. S. Army Specification TT-E-529
Brass, Copper or Bronze	Treat with Phosphoric Acid Metal Conditioner per Specification MIL-C-10578		
Steel	Cadmium or Zinc plate 0.003" thick minimum followed by chromate chemical dip treatment such as Iridite No. 8-P (1)		
Zinc	Chromate chemical dip treat such as Iridite No. 8-P (1)		NOTE: Wrinkle finish is now considered undesirable on military applications

(1) Iridite materials are manufactured by Allied Research Products, Inc. of Baltimore 5, Maryland, American Chemical Paint Company of Ambler, Pennsylvania, supplied similar products.

BIBLIOGRAPHY

Moses, Graham Lee
Electrical Insulation, First Edition,
McGraw Hill Publishing Company, New York, 1951.

Raffalovich, Aubrey J.
"Fungicidal and Solvent Retention Problems in Organic Film Coatings"
Paper presented at New York University College of Engineering
Symposium on Varnish and Paint Chemistry.

Young, James F.
Materials and Processes, Eighth Printing,
John Wiley & Sons, Inc., New York, 1949.

Information Bulletin No. 180 "Precautions for Design Engineers"
Components and Materials Branch, Signal Corps Engineering Laboratories,
Fort Monmouth, New Jersey.

Information Bulletin No. 212 "Silver Migration in Electronic Equipment"
Components and Materials Branch, Signal Corps Engineering Laboratories,
Fort Monmouth, New Jersey.

Information Bulletin 213 "Studies of Galvanic Corrosion in Communication Equipments" Components and Materials Branch, Signal Corps Engineering Laboratories, Fort Monmouth, New Jersey.

SPECIFICATIONS

TT-E-527, Federal Specification
"Enamel; Synthetic, Lustreless"

TT-E-529, Federal Specification
"Enamel, Synthetic, Semi-gloss"

TT-C-595(2), Federal Specification
"Colors; (for) Ready-Mixed Paints"

QQ-Z-325, Federal Specification
"Zinc Plating (electrodeposited)"

QQ-P-416(1), Federal Specification
"Plating, Cadmium (electrodeposited)"

72-53 U. S. Army Specification
"Finishes (for ground signal equipment)"

MIL-T-152(1)
"Treatment, Moisture and Fungus Resistant, of Communications Electronic, and Associated Electrical Equipment: General Process for"

MIL-V-173A
"Varnish, Moisture and Fungus Resistant for the Treatment of Communications, Electronic and Associated Electrical Equipment"

MIL-C-5541(1)
"Chemical Films for Aluminum and Aluminum Alloys"

MIL-P-6889A(4)
"Primer; Zinc Chromate, for Aircraft Use"

MIL-A-8625
"Anodic Coatings, for Aluminum and Aluminum Alloys"

MIL-D-8635
"Decalcomanias, for Use on Internal Surfaces of Aircraft"

MIL-T-12879
"Treatments, Chemical, Prepaint and Corrosion Inhibitive, for Zinc Surfaces"

MIL-M-13231
"Marking of Electronic Items"

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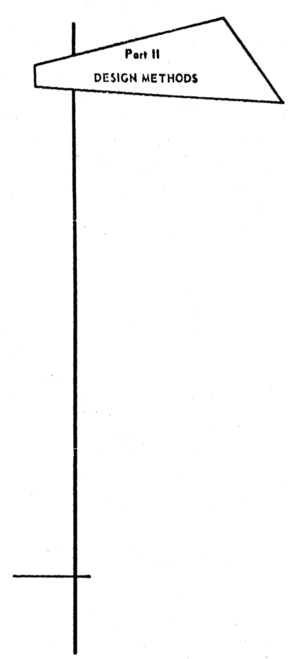

POOR ORIGINAL

INTRODUCTION

For many years the electronic industry has operated with hybrid methods of r-f transformer design made up of excellent but highly complex theory generously aided by cut-and-try procedures. This has obviously led to many cycles of making and testing samples, each cycle being a slight improvement over the preceding one, until finally, an established goal was attained. This laborious cut-and-try method was necessary since an adequate bridge between the complex theory and practical design was not available.

This empirical type of design operation was tolerated only because the electronic art was in its infancy and the necessary manpower was available to implement these cut-and-try procedures. Electronics has expanded tremendously since the beginning of World War II and most engineers have been occupied with systems and equipment designs; a comparatively small percentage has entered the inductive components field, which continues to operate with too thin a distribution of engineering manpower. Another state of National Emergency would seriously complicate this situation. One recognized expedient for coping with this current dearth of skilled manpower in the inductive field is to reduce or eliminate the time consuming cut-and-try design operations (which inevitably takes on the aspects of "art") in favor of a more straight-forward and simple analytical approach.

Part II presents two design procedures — one graphical and one analytical. Both approaches are especially directed to the engineer who has not had extensive experience in this specialized r-f coil field. The graphical method is the simpler, but necessitates that this manual or at least the necessary charts be available. A thorough study, leading to an understanding of the analytical method, is recommended since it can be carried out at any time or place without reference to special charts or graphs. Either method will establish values for all parameters, so that an experimental unit can be quickly produced without any extensive empirical operations.

The sections on *Windings*, *Types of Construction* and *Measurements* provide the necessary know-how to translate the computed electrical parameters into a physical unit and to finally establish, by suitable measurement and test, that the transformer will comply with the original requirements. It is recommended, therefore, that the *Measurement* section, particularly, be carefully studied so that new designs can be properly evaluated. Radio-frequency measurements are subject to considerable variation between equipment setups because of differences in instrument calibration, stray effects and differences in test jigs and fixtures. Such variations can be sources of *apparently* serious deviations from design values. The *Measurement* section emphasizes the necessity for standardization of measurement procedures in r-f coil testing.

It should not be assumed that even with simplified design procedures that it will always be possible to design r-f transformers capable of meeting all requirements without some modification. It is reasonable to expect, however, that a systematic following of the procedures as outlined will produce an experimental unit that will require a minimum of subsequent revision by cut-and-try techniques.

It is recommended for purposes of comparison that typical design problems be attempted by both methods. The results should be compared and judged upon the basis of the effort and time expended to attain the desired results. As the designer becomes more familiar with his field, experience will enable him to further reduce design time. The method that is finally chosen for future problems of r-f transformer design will be determined by personal preference.

Section 10
WINDING – EQUIPMENT AND TECHNIQUES

INTRODUCTION

Most inductive radio-frequency components are basically structures comprising one or more coiled windings of insulated wire. In this section are presented descriptions of types of windings, their characteristics, limitations, and applications, and discussion of the machinery and techniques employed in the fabrication of commonly-used types. Included also are formulae, charts, and tabulated data useful to the designer.

GENERAL

The characteristics and utility of a winding are governed by its dimensions, the disposition of wire within those dimensions, and the quantity and kind of wire that is used. Although other shapes may be used for special purposes, most practical radio-frequency windings are essentially cylindrical in form, with proportions varying over a wide range. In the typical case, the shape is that of a ring whose mean diameter is somewhat larger than either its length or its height, these being broadly of the same order (Figure 10-1).

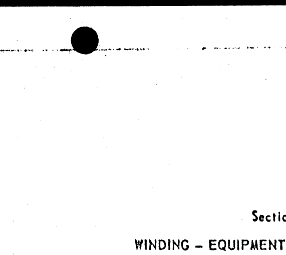

Figure 10-1. Basic Winding.

The winding may be elongated either axially (Figure 10-1a) or radially (Figure 10-1b) as the ratio between the number of layers of wire and the number of turns in each layer is varied. Thus, when the number of layers and turns-per-layer are about equal, the cross-section may be square (Figure 10-2). In one extreme instance of a winding all of whose turns are in a single layer, the structure resembles a helical spring (Figure 10-3) while a multiple-layer winding with one turn in each layer has the appearance of a spiral spring (Figure 10-4).

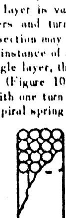

Figure 10-2. Layer winding, square cross section.

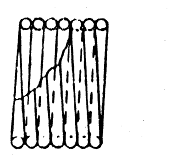

Figure 10-3. Solenoid Winding.

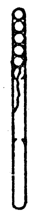

Figure 10-4. Spiral Winding.

An infinite variety of turns distributions can be conceived and wound between these limits. Within the selected winding outline, the spatial relations among the individual turns may be arranged in many ways. The proportion between diameter and cross-sectional dimensions is restricted by inherent structural limitations of some windings. The adoption of a specific winding configuration will be dictated for each application by economic considerations, as well as by the physical and electrical requirements.

Industrial machinery is available for the manufacture of all useful winding types and specialized handling techniques have been evolved. Familiarity with machines and methods is of great value to the coil designer. While secondary experimental work is often necessary and desirable, appropriate basic design procedures should be understood and adhered to. Decisions based from the chance interplay of unknown or abnormal materials and processes must be avoided for optimum results and if engineering prototype models are to be duplicated successfully in large-scale production.

WINDING TYPES, GENERAL

The simplest winding of more than one turn is that which has all of its turns sequentially and uniformly disposed in a single layer (Figure 10-3). Although the usage is ambiguous, such windings are conventionally called *solenoids*. The spacing between adjacent turn-centers may be equal to the effective wire diameter or greater. Solenoids are sometimes modified by a systematic variation in turn spacing, thereby becoming *variable-pitch solenoids*. A winding may be wound in successive layers of alternating direction, each essentially a simple solenoid; these are frequently called *multi-layer solenoids* or, simply, *multi-layer windings* (Figure 10-5).

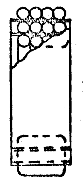

Figure 10-5. Multi-layer Winding.

If the wire in a multi-layer winding cyclically traverses the winding surface at an angle greater than the helical angle corresponding to the turns spacing, a *lattice* or so-called *universal winding* results. Portions of the wire in opposing half-cycles of axial movement then mesh with one another to form an interlaced pattern (Figure 10-6).

Figure 10-6. Universal Winding.

An important development from the universal winding is the *progressive-universal winding* in which gradual axial displacement, superimposed upon the cyclic excursion of the wire, retards layering by elongation of the winding in the direction of displacement. A solenoid-like winding thus can be produced in which the length-per-turn may be considerably less than the effective wire diameter (Figure 10-7). As in simple solenoids, variable-pitch windings can be made.

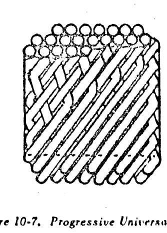

Figure 10-7. Progressive Universal Winding.

A flat spiral winding is one in which each turn lies directly over its predecessor. While it has some aspects of a single-layer winding, it is properly regarded as a one turn-per-layer structure (Figure 10-4). The banked winding is essentially a multiplicity of successive spirals laid down at a backward-leaning angle so that each turn in the upper layers is supported in the depression between two turns in the next lower layer in preceding spiral groups (Figure 10-8).

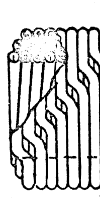

Figure 10-8. Bank Winding.

Duo-lateral, honeycomb, and *diamond-weave winding* are obsolete ancestors of the universal winding. The extinct *spider-web winding* is a spiral whose turns oscillate cyclically about their mean position as the analogous *basket-weave winding* is a solenoid in which the oscillation is about the mean circumference. Mention of these disused types is made only to clarify occasional textual references, particularly in older literature.

SOLENOID WINDING

The solenoid winding has the obvious superficial advantage of being a "natural" structure; it can be wound fairly easily by hand or on almost any rotating machinery, such as a lathe, in lieu of a specific winding machine. Since inter-turn capacitances are almost solely between sequential turns, total distributed capacitance tends to be low as does the accompanying dielectric loss. There is the incidental benefit that the distributed capacitance is not likely to constitute an inordinately large portion of the circuit capacitance with which the inductor is to be associated. Also because only sequential turns are proximate, the potential differences among them are a proportionately small part of the total potential appearing across the winding, making the ability to withstand high potentials inherently good.

The solenoid suffers from poor space utilization resulting from the high ratio of length to height (Figure 10-3). Additionally, due to the remoteness between turns near the extremities, the efficiency in respect to inductance yielded by a given linear quantity of wire is relatively low. Accordingly, it is difficult to keep conductive (ohmic copper) loss within reasonable bounds for high values of inductance to be used in radio receiving and other equipment where minimal size is desired. Where close spacing between windings is necessary, as in tightly-coupled transformers, construction can become clumsy or even physically unrealizable.

Suitable applications for solenoid windings are in inductors of the order of fifty microhenries or less corresponding very roughly to an operating frequency of two megacycles; for inductances of not more than about ten microhenries, there is usually no alternative. Where bulkiness is not proscribed and low distributed capacitance or high potential (voltage) capability are of paramount importance, as in larger radio transmitting equipment, or where poor conductance is tolerable, as in many radio-frequency choke coils, solenoids of several hundred microhenries may be utilized.

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The usefulness of solenoids may be extended notably by recourse to ferromagnetic cores. Aside from increases in practical inductance values they make possible, cores may be used for inductance adjustment; continuously variable cored inductors are used in variable reluctance or so-called permeability tuning systems (see Section 3).

VARIABLE-PITCH SOLENOID WINDING

In permeability tuning, the rate of inductance change must sometimes be related uniquely to the degree of core insertion. This need exists, for example, in the tuner for a superheterodyne radio receiver; in response to synchronized core movement, the inductance of the oscillator coil must change in a manner so related to the simultaneously changing inductances of the signal-circuit coils that the resultant resonant frequencies of the respective circuits are separated by a constant numerical difference, equal to the operating frequency of the receiver's intermediate-frequency amplifier section.

In other applications, it may be desired that the inductance of a coil vary with core position according to a logarithmic or exponential law. As an instance, inductance change proportionate to the square of core-travel is of interest when a linear display of frequency is wanted on a frequency-control indicator-scale. A variable-pitch solenoid winding offers means of at least partly fulfilling the foregoing requirements. Such a winding also may be resorted to when a coil is to be used as an artificial transmission line with dissimilar terminal impedances or wherever non-uniformly distributed capacitance is of value.

MULTI-LAYER WINDING

A layered winding of solenoidal form has high conductive efficiency. This can be explained in the following way: The total inductance of a winding is the sum of the inductances of each of its turns plus all of the mutual inductances among the turns. Mutual inductance is a function of electromagnetic linkage and, therefore, of spacing between turns. It follows that a compact winding, all of whose turns are close together, will exhibit high inter-turn mutual inductances and, thereby, high inductance per unit length of wire. As a corollary, a square cross-section will yield the highest conductive efficiency of any rectangular-section winding; the square is a rectangle of least dimensions defining a given area, hence, it is the most compact outline to contain a given number of turns.

Unfortunately, the very high distributed capacitance and dielectric loss of simple multi-layer windings render them unfitted to most services at frequencies exceeding a few tens of kilocycles. In the worst case of a two-layer winding, the first and last turns are in juxtaposition, introducing the capacitance between them across the entire winding. Paper or synthetic resin sheets are frequently inserted between layers to reduce capacitance, to furnish a stable base for each layer, and to increase the ability to withstand operating potential (Figure 10-5). Such windings are generally used with ferromagnetic cores. While it is primarily for sub-radio-frequency applications, not within the scope of this manual, discussion of the simple multi-layer winding helps to lay a background for some of the more complex and useful windings. Paper-section multi-layer windings are used in some inexpensive pulse transformers for television high-voltage power supplies.

UNIVERSAL WINDING

The universal winding (Figure 10-6) retains most of the advantages of the simple multi-layer winding while the faults are largely eliminated. It is virtually as compact and efficient in wire utilization but distributed capacitance and dielectric loss are greatly reduced. Were it not for the residual capacitance, which is still markedly higher than that of a solenoid, the square section would be preferred. Since the greater component of the capacitance may be regarded as residing between the layers rather than between adjacent turns, radial elongation to a degree dependent upon inductance and frequency will tend to reduce dielectric loss more rapidly than conductive loss is increased. A rectangular section whose height is, say, five times the length is a practical limit usually dictated by physical winding limitations.

Another method for distributed capacitance reduction, which is not as wasteful of space in the radial plane, is sectionalization (Figure 10-9). The relation between sections is sometimes manipulated for inductance adjustment. The nominal interface spacing of adjacent sections is not very critical. The advantage of using more than three sections is negligible. When ferromagnetic cores are employed, either for inductance adjustment or for the improvement of efficiency, the gains from sectionalization are much reduced, since a core will itself increase the distributed capacitance of a sectional winding more than it will that of a solid one.

Best applications for universal windings range from the highest inductances for radio frequencies

down to the vicinity of two-hundred microhenries, at frequencies up to about one-half Megacycle. As a general rule, universal windings should not be

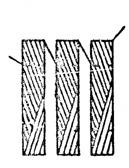

Figure 10-9. Sectionalized Universal Winding.

considered above two Megacycles unless distributed capacitance is unimportant or extreme space-saving is mandatory. Universal windings are extensively used in 0.455 Megacycle and lower frequency intermediate-frequency transformers.

PROGRESSIVE-UNIVERSAL WINDING

To the design problems posed by inductances ranging from fifty to two-hundred microhenries for use between one-half and two Megacycles, approximately the medium-wave radio-broadcast spectrum, where neither solenoids nor universal windings are entirely satisfactory, the progressive-universal winding offers an ideal solution (Figure 10-7). Low distributed capacitance, approaching that of solenoids, and reasonable conductive efficiency and compactness are characteristic. Inductances as high as two-thousand microhenries are sometimes wound over ferromagnetic cores.

Progressive-universal windings are well adapted to utilization in permeability-tuning systems requiring higher inductances than can be obtained properly with solenoidal construction. Where appropriate, the variable-pitch solenoid has its counterpart in the variable-pitch progressive-universal winding. The most common application of the progressive-universal winding is in the signal-circuit coils of radio-broadcast receivers.

FLAT SPIRAL AND BANKED WINDINGS

The flat spiral winding is difficult to wind and is extremely wasteful of space although it can have a fairly low distributed capacitance (Figure 10-4). For the latter reason and because the shape lends itself, the flat spiral is sometimes used for radio receiving loop antennas. Its large outline is consistent with this usage but it is usually deformed into ovoid shape to fit the oblong receiver housing (Figure 10-10).

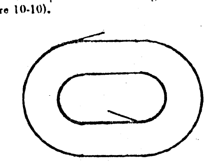

Figure 10-10. Spiral-wound Loop Antenna.

Banked windings share with progressive-universal windings many of their features and formerly were used widely in similar circumstances. Manufacturing difficulties have relegated them to obscurity except for occasional large, many-layered windings of large wire which are frequently hand-wound.

WINDING MACHINES

A winding machine must provide means of supporting and rotating the base upon which the winding is to be applied and a device for guiding the wire of which the winding is made. In simplest form the machine comprises a unifying frame; a shaft fitted with a power receptor, such as a pulley, gear, or crank; a mandrel for holding the work; and a wire-guide, idiomatically called button, finger, needle, etc., secured so as to retain limited freedom of radial movement. There is usually an auxiliary shaft which serves to transmit power to the button moving mechanism at the proper rate relative to work-shaft rotation. The necessary space and velocity relations between the shaft and the button are positively established, in all practical machines, by means of gear trains. To extend the utility of available gear sets, compounding facilities are usually provided. Sometimes compounding is limited to fixed idlers of simple fractional ratios.

The proportions of a machine are governed mainly by the size of the gears. The development of winding machinery has proceeded along two quite divergent lines; most machines are designed around spur gears of either 20 or 48 diametral pitch. Both styles can be adapted to most winding tasks but the smaller 48-pitch machines are somewhat better suited to small precisely-controlled work while the large 20-pitch equipment can handle work beyond the capacity of the more compact machines. 20-pitch change-gears are the smallest commercially available in consecutive sets; this accounts, in part, for their popularity.

Motivating power for most machines is derived from electric motors and is usually transmitted through belt-and-pulley drives. A few machines used for limited experimental work or for winding coils of very few turns are driven by manual crank. In practice, the work-shaft, or spindle, is normally discontinuous, its components being connected by a clutch mechanism. The clutch is conveniently joined with a braking member, the combination making rapid starting and stopping of the machine possible. Alternatively, the power may be applied to and the clutch and brake may be installed in the auxiliary shaft.

Presettable automatic stopping devices, both electrical and mechanical, are widely used. The latter tend to be more positive in action and are, therefore, preferable for highly accurate work. Preliminary slow-down prior to full stoppage has been incorporated in some machines in attempts to achieve precise control. Modern automatic-stopping machines do not require integral turns counters such as are built into manual or pedal controlled machines. In typical shop practice, a separate counter, with fitting for temporary attachment to each machine, is used for stop setup only.

The mandrel is essentially a removable extension of the spindle which is designed to fit the form upon which the coil is wound. Spring fingers, knurled surfaces, clamping screws, and the like are employed to secure the form and wire, yet to allow speedy removal of the completed coil. Tandem mandrel arrangements are sometimes used to increase machine output in relation to running time. In other systems, long mandrels supporting a number of coil forms end-to-end are used for the same purpose. The value of increased effective winding speed by such devices is self-limited beyond a point, related to the number of winding turns, since handling time becomes an increasingly larger proportionate part of the total cycle. Multiple mandrels are usually made quickly detachable from the machine spindle so as to remove a large part of loading and unloading time from the total. Such mandrels or any extraordinarily long, thin mandrels may require additional support from a tail-stock on the machine.

Winding speeds are limited by the inertia of moving machine elements, wear on cams and other elements moving at high peripheral speeds, wire strength according to its size, and the performance of the wire holder and tension device which is used. Since each of the foregoing factors influence machine speed to varying degrees it is difficult to suggest exact limits. A good average starting speed is about 150 ft. per minute. It is suggested that the recommendations of the machine manufacturer be carefully studied and applied as necessary.

The best types of holders are those in which the wire is removed from a stationary spool with little inertia. Tension should be capable of smooth and dependable adjustment. Rotating spool-holders are particularly poor since the inertia contributed by the spool is high and variable with wire content. Auxiliary wire guides, suitably supported, are used to direct the wire from the tension device to the button; abrupt changes in direction can cause work-hardening of the wire.

The button is a slotted or grooved member through which the wire passes as it is laid down upon the winding surface. It is preferably supported and shaped so that the normal plane of the wire, as it is released, is tangent to the surface of the winding so as to minimize frictional losses and other disturbing forces. A moderate amount of radial pressure is usually applied to the button to prevent erratic movement away from the winding surface. This is best obtained by adjustable spring-loading but, in some machines, is supplied by force arising from wire tension.

The specific configuration of the button is dependent upon the idiosyncracies of the winding and upon the skill and inclinations of personnel; an enormous variety of shapes and many materials have been successfully used. A hardened and polished steel block with a V-shaped groove emerging to the surface adjacent to the winding is generally useful. The contour of the button can be modified, as required, to fit various winding needs.

SOLENOID WINDING MACHINES[1]

As might be expected, solenoid winding machines are simple in form and in operation. An auxiliary shaft geared to the spindle actuates the button along the winding axis through a suitable mechanism. This may be a worm-and-nut, gear-and-

rack or cam-and-pushrod movement. In a typical worm-drive arrangement, a half-nut-like rider is held in place upon the worm by gravity and it is hinged so that it can be disengaged readily for return to the start. This movement is positive but there is an uncertainty in starting location due to the necessity for reengaging the worm. Obviously, the lead-per-turn of the winding will equal that of the worm for unity gearing of the spindle and worm-shaft. (Figure 10-11).

For rack movement, a pinion shaft is transversely located with respect to the auxiliary shaft and is coupled through reduction-gearing, usually 1/100 in ratio. A 16-tooth, 48-pitch pinion on this shaft engages the rack, giving $\pi/300$ or 0.010472 inch lead-per-turn when the spindle to auxiliary shaft ratio is 1/1. A friction clutch is included in the pinion shaft to enable manual return to the starting point (Figure 10-12).

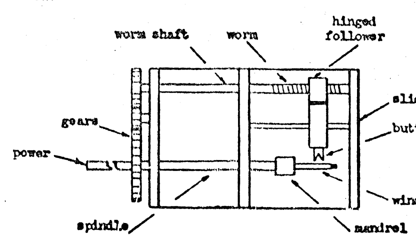

Figure 10-11. Solenoid Winding Machine.

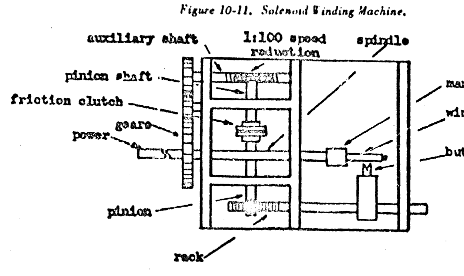

Figure 10-12. Solenoid Winding Machine with Friction Clutch.

[1] Photos of typical winding machines are included in the catalogues of most winding machine manufacturers such as George Stevens Manufacturing Co., Inc., Chicago, Ill., and Universal Winding Company, Providence, R.I.

In a somewhat similar arrangement, the pinion is replaced by a flat cam and the rack by a push-rod terminated in a cam-follower; this rod must move laterally in line with the cam center. For simple solenoids, the cam has linear rise but the contour can be shaped to produce virtually any variable pitch. This unique feature renders the cam-driven machine most useful of all solenoid winders. Spindle to cam-shaft gearing is determined by the relation between the rate of cam rise and the winding pitch; 1/100 gear reduction is usually introduced. Precise cam forming is essential since only part of one cam cycle is used for winding, the remainder being reserved for the return stroke. No other means of returning to the start is needed. The follower is spring-loaded to maintain contact with the cam, as are the analogous parts of the other machines to remove lash in their respective movements (Figure 10-13).

The button on a solenoid machine is usually spaced away from the winding surface slightly more than the wire diameter, by means of an adjustable stop, to avoid disruption of already wound turns. For the sake of simplicity, clutches, brakes, guides, etc. are omitted in machine figures.

UNIVERSAL WINDING MACHINES[1]

One variety of universal winding machine is quite similar to the cam-operated solenoid machine except that the cam-shaft is geared for much higher rotational speed. Instead of less than one cam revolution for the entire winding, there may be from a fraction of one to several cam cycles per turn. The specific number depends upon the number of traverses or crossovers required, each cycle covering an even number of crossovers; two are almost invariably used (Figure 10-14).

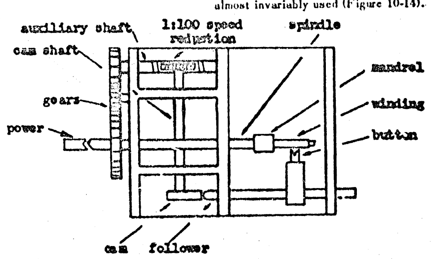

Figure 10-13. Solenoid Winding Machine with Spring-loaded Cam Follower.

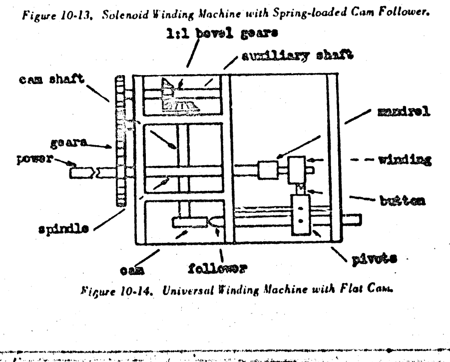

Figure 10-14. Universal Winding Machine with Flat Cam.

By reorientation of the cam-shaft parallel to the spindle, it can be geared directly to the spindle, a cylindrical cam being used instead of a flat cam. This construction is favored in heavy-duty 20-pitch machinery (Figure 10-15). In either machine type, gearing determines the relative shaft speeds, hence, the number of crossovers-per-turn. The length of the winding between wire-centers at the extremes of excursion is equal to the cam-throw; the overall length is one wire diameter greater in a perfect winding. All practical cams are linear and every effort is made to effect instantaneous return at the ends of the throw.

PROGRESSIVE-UNIVERSAL WINDING MACHINES

The progressive-universal winding combines characteristics of solenoid and universal windings and, similarly, a machine for winding them is a hybrid combining the functions of both machine types. A basic solenoid machine can be adapted to progressive-universal operation by the addition of a second gear-train, auxiliary shaft, and an oscillatory or shuttle cam to provide the cycle traversing movement (Figure 10-16). Alternatively, a universal machine is convertible by the addition of the components for lateral movement (Figure 10-17). The difference in approach is significant only insofar as whichever basis affords clearer understanding of the composite nature of the machine. As in solenoid machines, the cam-driven lateral drive makes it possible to wind other than constant-pitch windings. Buttons must be shaped to avoid previous turns.

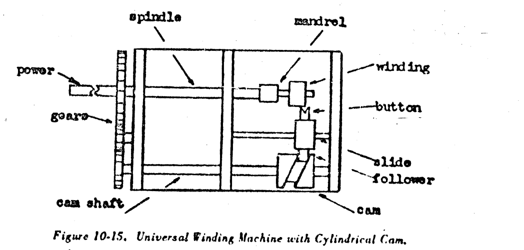

Figure 10-15. Universal Winding Machine with Cylindrical Cam.

The button is, of course, actuated by the cam-follower. For simplicity, the button is usually attached to a pivoted arm. Since the resulting arcuate button travel deviates from the radial path necessary to maintain tangency between the released wire and the winding surface, as the winding diameter increases, the advantage of using a long arm to minimize the deviation is apparent. Truly radial button travel can be obtained only with undesirable complications. The oscillating follower, push-rod, and button assembly are designed for light weight, consistent with rigidity and durability, to prevent disruptive mechanical resonances at high speeds. Buttons are often ganged to wind similar windings on the same coil form simultaneously or for use with multiple mandrels.

FLAT SPIRAL AND BANKED WINDING MACHINES

Flat spiral winding machines are basically simple inasmuch as no axial or oscillating movement takes place. Such complications as exist are entirely in the mandrel which assumes also the function of wire-guide. It is essentially a pair of demountable plates separated by a spacer as thick as the wire diameter. The shape of the spacer may be other than circular in order to develop the desired shape of the antenna loop, for the usual application. Corrugations are sometimes pressed into the winding after removal from the mandrel to impart stiffness and as a means of inductance adjustment; shape-retaining supports may then be dispensed with.

[1] Photos of typical winding machines are included in the catalogues of most winding-machine manufacturers such as George Stevens Manufacturing Co., Inc., Chicago, Ill., and Universal Winding Company, Providence, R.I.

Intricate machines were formerly used to wind banked windings. In view of the present existence of progressive-universal equipment, the simplest procedure in the rare event of necessity is to use such equipment with stepped cams of special design. Because of their virtual obsolescence further discussion or study of banked windings is unwarranted.

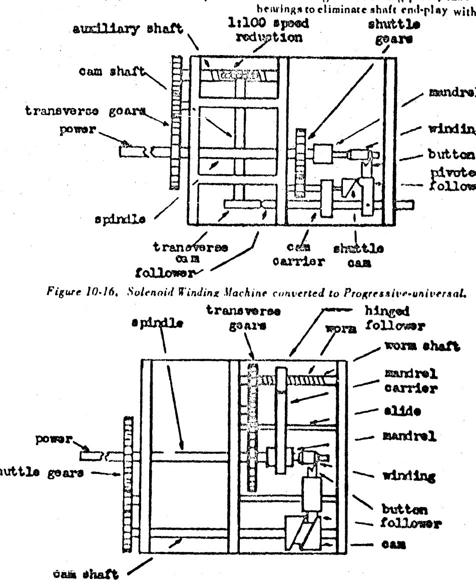

Figure 10-16. Solenoid Winding Machine converted to Progressive-universal.

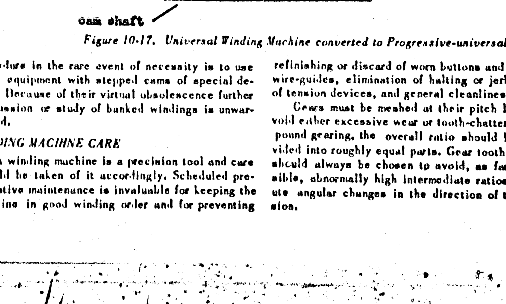

Figure 10-17. Universal Winding Machine converted to Progressive-universal.

WINDING MACIHNE CARE

A winding machine is a precision tool and care should be taken of it accordingly. Scheduled preventative maintenance is invaluable for keeping the machine in good winding order and for preventing work spoilage. Among items of maintenance common to all machines are periodic lubrication of shaft bearings and rubbing parts, take-up of thrust bearings to eliminate shaft end-play without binding, refinishing or discard of worn buttons and auxiliary wire-guides, elimination of halting or jerky action of tension devices, and general cleanliness.

Gears must be meshed at their pitch line to avoid either excessive wear or tooth-chatter. In compound gearing, the overall ratio should be subdivided into roughly equal parts. Gear tooth numbers should always be chosen to avoid, as far as possible, abnormally high intermediate ratios and acute angular changes in the direction of transmission.

Proper tools should be used to prevent such damage as burring of shafts and stripping of threads. Excessive force should never be applied to adjust or tighten machine members. Mandrels must be perfectly concentric and capable of holding securely coil-forms within their specified dimensional tolerances. Excessively worn parts that cannot be compensated should be replaced without delay. Clutch and brake facings must be kept free of grease and dirt. In cam-drive machines, special attention should be devoted to the cam contour and to the condition of the follower mechanism. Some machines utilize a lever action to effect a variable throw from a single cam; the elements of this action must be carefully scrutinized to prevent non-linearity and deviation of throw.

WINDING MACHINE SETUP

The detailed operation of setting up a winding machine will depend upon its application and construction. However, the general objective is to make preparation for winding coils, of desired dimensions, exactly located with respect to stated reference points on their forms. Means are provided in the button supporting system for properly positioning the start of the winding. The internal winding dimensions will lie within predictable ranges if these have been correctly set.

In the event that a winding does not meet the computed physical specifications or does not wind uniformly, under no circumstances should the setup be subjected to random experimentation. Rather, the winding table should be scrutinized for fundamental errors of judgement and execution and should be corrected or confirmed, as the case may be. Attempted compensation of design error or machine flaws by means of arbitrary modifications can be expected to make consistency and precise manufacturing specification impossible. Changes, if indicated, should be made in a systematic manner, based upon observed results.

Micrometer calipers, steel rules used with low-power magnifiers, and special gages to fit unusual situations are valuable tools for winding measurements. Initial samples made of controlled materials and, where feasible, held to closer than allowable dimensional standards should be submitted to electrical test, before considerable quantities are wound, to establish whether or not nominally correct winding data have been evolved to meet electrical requirements. Minimal wire tension to prevent the escape of wire from its spool faster than it is consumed in the winding and button pressure just sufficient to maintain contact are proper.

A natural inclination to force the winding into submission, by applying excesses, is to be suppressed. As a generality, it is axiomatic that a good winding is a passive winding which is neat and uniform in appearance, with its pattern clearly defined and without voids or erratic migrations of wire.

The most critical adjustment in machine setup is the button setting. Aside from proper shape and tangential relation of the groove to the winding surface, a series of minute secondary adjustments will usually be necessary for the achievement of best results. Positioning of automatic stops, on machines so equipped, may require painstaking manipulation. Variable-throw cam devices introduce an additional detail of careful adjustment.

WINDING HANDLING

A satisfactory winding having been designed and a stable machine setup suited to its production having been made operative, it is in order to consider methods for handling the coil, while on the machine and preparatory to its removal. Except in layered windings, in which the start is bound by the winding itself, the start and finish ends of the wire must be secured. Flowed-on molten sealing wax, fast-drying thermosetting cement, adhesive tapes, or bits of cellulosic film material softened in solvent may be used for this purpose. Care must be taken that the binding material does not interfere with other coil elements, or protrude beyond a specified outline dimension.

Integral or securely attached terminal lugs on some coil-forms may serve exclusively as means of ticing down the winding ends. When cement-coated wire is used, it can be made self-adherent by the application of a suitable solvent. A practice to be avoided, unless strictly necessary, is to wind over a coating of cement or layer of adhesive tape to secure a winding to its base. An insulating tape may sometimes be introduced when one winding is wound directly over another. Occasionally, windings are placed over removable sleeves or tapes to make subsequent positional adjustment possible.

It is sometimes expedient to run the wire through a liquid bath before reaching the button as a means of removing oily deposits and to promote temporary adhesion between turns in the course of winding. Any non-corrosive material, which will not attack wire coverings and which can be driven off readily by evaporation leaving no residue, is usable. The bath should not be used as a means of inducing an unsound winding to cohere. A completed winding is by no means a homogenous mass and must be handled with care, when removed from the mandrel and

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thereafter, to prevent deformation or disturbance of position.

WINDING DESIGN

The selection of an optimum winding design constitutes a complex problem whose solution is influenced by many interrelated considerations. Aside from the choice of a generic type, dictated by inductance and operating frequency, judgements must be made as to winding proportions, wire size, and relations to associated windings, electrostatic shields, ferromagnetic cores, and other elements of the coil assembly, which will assure compatible conformity with physical and electrical requirements.

When a coil is to be one of a series, it is usually desirable to combine forms and accessories into one or a few types; some degree of compromise to make this possible is necessary. It is frequently expedient to adopt available approximately correct parts rather than to search for theoretically optimum material.

The chief parameters that define a single winding are its inductance and its figure-of-merit or Q; secondary requirements to be satisfied may be distributed capacitance, direct-current resistance, voltage breakdown ability, and the environmental stability of one or more characteristics. If, for the purpose of simple analysis, the effects of ferromagnetic cores are neglected, inductance is simply a function of dimensions and the number of turns. Q depends upon volume, materials, and their distribution. Components of dissipation, which is the reciprocal expression of Q, are simple conductive loss (direct-current resistance), increase of conductive loss through skin-effect (the tendency of high-frequency alternating current to flow near the surface or by the path of least impedance in conductors), distributed capacitance loss (the unfavorable redistribution of potentials in a winding resulting from internal capacitances), dielectric loss (effective shunt resistances originating in imperfect insulators), and induced loss (power consumed by the induction of currents in nearby objects).

For practical design purposes, there is little value in attempting analytical treatment of each loss component; it is sufficient to be aware of its existence and to design so as to minimize all losses in the aggregate. For the sake of Q alone, the largest winding is best for a coil isolated in space. Practical size is limited by other members of the coil assembly and extraneous objects. The usual limitation is imposed by a can placed around the coil for electrostatic shielding and physical protection.

The effect of a shield upon Q (See Section 2) depends upon its shape, proximity, surface resistivity, and electromagnetic permeability. For closed cylindrical shields of aluminum, a nearly optimum compromise between maximum winding size and separation from the shield is given by a two-to-one ratio between the inside shield diameter and the outside winding diameter and by end-spacing of the winding from the shield by its own diameter. Copper, having lower resistivity, may be brought somewhat closer to a winding and commonly-used zinc should be spaced about 25 percent further for best results. Electromagnetic metals, such as steel and nickel-bearing alloys, or high-resistance materials, such as interior cadmium plating, preferably should not be used in coil shields. A shield should be seamless in its circumference to avoid high-resistance joints. A square can may be considered the equivalent of a round can of equal internal periphery.

The selection of wire size partly involves elementary understanding of skin-effect. The density of a high-frequency current in a wire is greatest at the surface, falling inwardly at a rate varying in inverse proportion to the square root of the frequency. Although the transition is not truly abrupt but gradual, it may be assumed for practical evaluation that all of the current flows uniformly in a tubular outer section of the wire, within which there is no current. In copper, the depth of this penetration is taken as 2.5 milinches, at one Megacycle. A straight wire five milinches in diameter can, therefore, be stated to have no skin-effect at that frequency. The resistance of wire up to this diameter varies inversely with the area, or square of the diameter, as it does for direct-current. In wire larger than the critical diameter, the current boundaries are defined by two radii whose difference is constantly equal to the depth of penetration. Therefore, the resistance in such wire varies inversely only with the first power, or directly, with the diameter. Thus, for any frequency, a wire size can be established beyond which further resistance reduction is greatly diminished. When a wire is formed into a ring, as in a turn of a winding, the current seeks the most direct path which is at and near the inside of the turn; the conductive area may be considered to terminate at the mid-section of the wire. From this somewhat oversimplified analysis, it is apparent that the largest wire without substantial skin-effect in a winding is 2.5 milinches thick, or AWG42, at one Megacycle.

Now, while it is true that such a small wire is more efficient for its size than a larger wire, it may still produce much more resistance than is tolerable in the completed winding. To decrease the resistance, it is customary to use cabled wire of a number of insulated strands of the desirable small size; this is commonly known as Litzendraht (or Litz) due to the German origin of the product. At frequencies above about three Megacycles, the use of Litzendraht is inhibited by the tendency of current to pass intermittently among the strands, thereby not making full utilization of the cable. Strands somewhat smaller than would be adopted only from skin-effect considerations may yield a small additional increase in Q because of a slight reduction in distributed capacitance attending its use. Early usage of very many strands in very large coils used at low radio frequencies dictated twisting of the strands in a rope-like lay, to make best use of each strand. Present-day coils, using a few strands, at medium frequencies, show no significant improvement resulting from special strand distributions (See Section 1, page 7).

Aside from conductance, modified by skin-effect, the number and size of strands that may be used, where Litzendraht is suitable, or the diameter of solid wire, in other applications, depends upon a balance between available space and distributed capacitance. No generally-applicable categorical rules can be laid down for the selection of wire. Better guidance is possible under a more detailed discussion of each winding type. In regard to adaptability from a Q standpoint, the application ranges, to which the three major types are best suited, are summarized, in review, in the table of Figure 10-18.

magnetic shells for electromagnetic shielding; a shell is usually a liner within an electrostatic shield or can. The can is then electromagnetically isolated from the winding, thereby removing its induced loss to the extent that magnetic shielding is complete. The shell itself will introduce loss in the same manner as does a core. This loss is often more than compensated when both a core and shell are used, a common practice; the reduced reluctance of the more nearly closed magnetic circuit effects a proportionately larger rise in inductance than in loss.

Further increase in efficiency may be possible by completion of the magnetic circuit through bridging members across one or both ends of the core and shell. Frequently, one of these members is integral with the shell or both shell and core. When an electrostatic shield is not electromagnetically isolated from a winding, a material effect upon inductance is observed. To a fair approximation, assuming end spacing equal at least to winding diameter, and low-resistance unity-permeability shield material, Equation 1 is useful.

$$L_o = \frac{L_s}{1 - \left(\frac{E}{S}\right)^3} \qquad (1)$$

L_s: shielded inductance
L_o: unshielded inductance (same units as L_s)
E : winding outside diameter
S : shield inside diameter (same units as E)

The equation fails for layered windings more than half as large as their shields, inadvisable in any case, and for very long or very short windings.

	Inductance Range (μH)		Frequency Range (Mc)	
Winding Type	Normal	Extended	Normal	Extended
Solenoid	under 50	under 200	over 2	over 0.5
Progressive-universal	50 to 200	50 to 2000	0.5 to 2	0.2 to 2
Universal	over 200	over 50	under 0.5	under 2

Figure 10-18. Recommended operating range of basic types of windings.

The effect of a ferromagnetic core upon a winding is to increase both the inductance, in accordance with its effective electromagnetic permeability, and the loss, due to eddy currents and its dielectric loss, as well as dielectric loss between coil and core. With proper usage, inductance is multiplied more than loss, yielding a net gain in Q (See Section 3). Coils may be surrounded by ferro-

SOLENOID WINDING DESIGN

Solenoid winding constants are easily computed. Having delineated the permissible outline dimensions, it is next in order to consider the best winding proportions. Since diameter is usually the limiting factor, it is convenient, first, to establish a diameter and, then, to determine the most favorable length. At very low frequencies, simple conductive loss is preponderant; the ideal length/diameter ratio, producing most inductance for the least resistance of wire, is 0.406/1. In radio-frequency service, where other losses assume greater importance with increasing frequency, a larger ratio creates a more favorable balance. For all normal radio-frequency applications of solenoidal windings, roughly 1/1 is entirely satisfactory, except when reduced length is dictated by the proximity of other windings or of a shield.

Once the dimensions have been established, it is possible to compute the number of turns required to produce the desired inductance. Over a wide range of length/diameter ratios, for windings of more than about seven turns, the solution can be obtained with good accuracy by Equation 2. For equal length and diameter, the simpler form of Equation 3 applies; Equation 4 covers the case of the 0.406/1 ratio.

$$N = \sqrt{\frac{(18D + 40W)L}{D}} \qquad (2)$$

$$N = \sqrt{\frac{58L}{D}} \qquad (3)$$

$$N = \sqrt{\frac{34L}{D}} \qquad (4)$$

N: number of turns
L: inductance (microhenries)
D: winding inside diameter (form o.d. - inches)
W: winding length between wire-centers (same units as D)

Corrections should be applied, as required, for shielding, the effective permeability of cores and/or shells, and for parasitic inductances external to the winding if they are of significant magnitude. If, for any reason, the first result is in error, correction is made easily by alteration of the number of turns in proportion to the square root of the required ratio of inductance adjustment, in the same winding length.

The most favorable wire size is smaller than the space alloted per turn in the winding length, roughly according to the empirical relation given by Equation 5.

$$d = \frac{s}{1 + \sqrt{\frac{FD}{30W}}} \qquad (5)$$

d: nominal wire diameter
s: lead-per-turn (same units as d)
D: winding inside diameter (form o.d.)
W: winding length between wire centers (same units as D)
F: frequency (Megacycles)

For use over a band of frequencies, the frequency where Q control is of greatest importance should be used; otherwise, the geometric mean frequency is a good choice. The theoretically best wire diameter must be rounded off to the nearest available size in the American Wire Gage standards table. Care must be taken that the maximum tolerable diameter over insulation, if any, plus a small allowance for the helical winding angle, say two percent, does not exceed the set lead-per-turn.

The choice of wire insulating material, if used, may depend upon allowable capacitance between superimposed or interwound windings, potential considerations, and climatic requirements. In a few applications, usually radio-frequency chokes for high-current circuits, direct-current resistance and current-carrying capacity may assume importance. Departures from optimum winding dimensions and wire size may be necessary to meet such an overriding requirement. Litzendraht is not of value at frequencies where solenoids are useful.

Due to the many factors involved, only the most general estimate of Q is possible. A formula, accurate to perhaps ±20%, is given by Equation 6. The simpler form of Equation 7 covers the case of equal length and diameter.

$$Q = \frac{300DW\sqrt{F}}{D + 2W} \qquad (6)$$

$$Q = 100D\sqrt{F} \qquad (7)$$

Q: figure-of-merit ($\omega L/R$)
D: winding inside diameter (inches)
W: winding length between wire-centers (inches)
F: frequency (Megacycles)

A final consideration in the completed winding design may be the necessity to adjust the turns to a particular fractional turn difference between start and finish, to fit a form configuration or terminal arrangement.

The order of the distributed capacitance of a winding is sometimes of interest. An approximation, valid for windings of equal length and diameter and not greatly affected by wire size, is expressed by Equation 8, when adjacent dielectric is essentially air.

$$C = 1.2D \qquad (8)$$

C: distributed capacitance (micromicrofarads)
D: winding inside diameter (inches)

The numerical factor increases to about 1.4 for length/diameter ratios of 0.4 and 2.3. Wiring, terminal, and other stray capacitances, as well as shielding and the presence of impregnants and coatings, will add substantially to the capacitance; coil grounding will roughly double it.

The calculation of winding machine gearing depends, of course, upon the type of machine used. In a worm-drive machine, the ratio of the rotational speeds of the spindle and the worm-shaft is expressed by Equation 9.

$$\frac{b}{a} = \frac{s}{w} \qquad (9)$$

a: spindle rotational speed
b: worm-shaft rotational speed (same units as a)
s: winding lead-per-turn
w: worm lead-per-turn

For rack-drive machines using the conventional 16-tooth, 48-pitch pinion and 1/100 reduction between the auxiliary and pinion shafts, the spindle to auxiliary shaft speed ratio is as shown by equation 10.

$$\frac{b}{a} = \frac{s}{0.010472} \qquad (10)$$

a: spindle rotational speed
b: auxiliary shaft rotational speed (same units as a)
s: winding lead-per-turn (inches)

Ratio calculation for cam-drive machines is based from the rate of cam rise. Although the complete rotation of a cam cannot be utilized for winding, a portion being set aside for the return stroke, it is a convenient concept to rate cams in projected rise per 360°. The return stroke may consume thirty or more degrees depending upon machine speed, total cam rise, and the number of wound turns. Equation 11 is applicable.

$$\frac{b}{a} = \frac{s}{rp} \qquad (11)$$

a: spindle rotational speed
b: auxiliary shaft rotational speed (same units as a)
s: winding lead-per-turn
r: cam rise per 360° (same units as s)
p: auxiliary shaft to cam-shaft reduction (usually 1/100)

Having determined the required speed ratio, it is necessary to convert this into a fraction whose numerator and denominator are both numerically equal to the tooth numbers of available gears which can be made to mesh when placed upon their respective shafts. The simple inverse relation is given by Equation 12. Naturally, simple gearing is unaffected by the inclusion of simple idler gears in the train. If compound gearing is used, the simple fraction expressing the required ratio is factored into two component fractions, preferably of comparable absolute value, each of which must be stated in numbers corresponding to practical gears. Equation 13 states this simple relation.

$$\frac{g}{f} = \frac{b}{a} \qquad (12)$$

$$\left(\frac{g}{f}\right)\frac{k}{j} = \frac{b}{a} \qquad (13)$$

a: spindle rotational speed
b: auxiliary shaft rotational speed
f: auxiliary shaft gear (teeth)
g: spindle gear (teeth)
j: gear of compound set meshing with spindle gear (teeth)
k: gear of compound set meshing with auxiliary shaft gear (teeth)

Frequently, it will not be possible to precisely fit actual gears to the needed ratio; the best practical compromise should be made. Compounding greatly improves the likelihood of obtaining very close fit. It will be quite apparent that the primer-like procedure, outlined above, can be reduced to a simple slide-rule or mental arithmetic operation, once the principles have been understood. One added consideration, peculiar to cam-drive machines, is that the chosen gearing must be capable of producing an integral number of turns in a projected complete cam rotation, corresponding to one complete cycle of winding and return; otherwise, the

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spindle (and mandrel) will not return to the same angular position for successive winding starts, as is frequently necessary.

In the special case of variable-pitch solenoids, worth-while analysis of general characteristics is impossible. The relation between instantaneous pitch and inductance slope is susceptible of mathematical treatment but this is tedious and not of general interest. It can be generalized that, in a variable-pitch winding, the inductance will be higher and the Q will be lower than in a constant-pitch winding of equal turns and dimensions. In respect to winding machine gearing, the method used for the cam-drive machine winding of simple solenoids is applicable, taking into account the variable rise of the specially-contoured cam.

UNIVERSAL WINDING DESIGN

In an air-cored multi-layer winding, the most inductance for a given wire resistance is produced when the length and height are both equal to 0.495 times the inside diameter; for all practical purposes, this may be taken as one-half. These proportions hold true for a universal winding at very low frequencies (under approximately 100 kc). At higher frequencies (above 100 kc) where losses other than conductive must be dealt with, reduction of length can improve Q substantially. A rough guide to the determination of optimal length is given by Equation 14, the height remaining one-half the inside diameter.

$$W = \frac{H}{1 + \sqrt{8F}} \qquad (14)$$

W: winding length between wire-centers
H: winding height (same units as W)
F: frequency (Megacycles)

The winding inside diameter should be, as nearly as possible, one-half the outside diameter of the coil permitted by shielding and other space considerations. The important operating frequency or the geometric mean of the frequency range should be considered.

When a winding is sectionalized, an overall square cross-section, with height again one-half the inside diameter, is satisfactory. The length of each section is then shown by Equation 15. A non-critical 1/16" interface spacing is customary.

$$V = \frac{H - u(q-1)}{q} \qquad (15)$$

V: winding section length between wire-centers
H: winding height (same units as V)
u: interface spacing (same units as V)
q: number of sections

If the use of a ferromagnetic core is contemplated, weight must be given to the importance of winding all turns as closely as possible to the core for its best utilization, i.e.: highest effective permeability, without too much departure from the dimensions dictated by other considerations (See Section 3). Accordingly, a shortened winding is not indicated; equal length and height, each not greater than half the inside diameter, produce best results in most cases. After a first estimate of dimensions has been made, modification may be necessary to suit available coil-forms and, when used, ferromagnetic cores. Additionally, since shuttle-cams are usually available in fixed steps, deviation from the ideal length may be required.

When the projected winding dimensions have been established, the number of turns can be computed by the use of Equation 16. For the preferred ½ ratio of height to diameter, Equation 17 applies. If the length is also one-half the diameter, the simple form of Equation 18 may be used.

$$N = \sqrt{\frac{(15.5D + 63H + 45W)\ L}{D + H}} \qquad (16)$$

$$N = \sqrt{\frac{(21D + 20W)\ L}{D}} \qquad (17)$$

$$N = \sqrt{\frac{31L}{D}} \qquad (18)$$

N: number of turns
D: winding inside diameter (form o.d.-inches)
H: winding height (same units as D)
W: winding length (between wire-centers; same units as D)
L: inductance (microhenries)

Account should be taken of shields, cores, shells, and other external influences upon inductance. Secondary corrections can be made, over a moderate range, by changing the number of turns in proportion to the square root of inductance change, if the length remains unaltered. It is difficult to lay down rules for the selection of wire. The largest nominal wire diameter, over insulation, that can be fitted into the calculated winding cross-section is shown closely by Equation 19.

$$d = 0.9 \sqrt{\frac{HW}{N}} \qquad (19)$$

d: nominal wire diameter
H: winding height (same units as d)
W: winding length (between wire-centers; same units as d)
N: number of turns.

It does not follow that the largest wire for which space exists is best, as would be true at low frequencies. Frequently, better Q can be obtained with a smaller wire, in a shorter winding, or with comparatively thick insulation for which room must be made at the expense of conductor size.

Litzendraht is well adapted to most applications of universal windings. Where its use is justified, a cable of appropriate size, containing the least number of largest strands that will give the desired Q, should be chosen. The cost of a many-stranded cable of fine wire may be much greater than that of one with fewer strands of somewhat larger wire. The smallest strand that may be useful is related to frequency, as previously shown in the discussion of skin-effect. The secondary bases influencing the choice of insulating material and of wire size have been mentioned in the consideration of solenoid windings.

The distributed capacitance of universal windings cannot be calculated with any accuracy; it can merely be stated that the capacitance varies with winding diameter and with winding length. Height is a small factor except in windings with very few, three or less, layers, where capacitance is apt to be very high; such windings are not recommended.

Since the universal winding has three-dimensional form, its analysis and design constitute problems in solid geometry. Separation of the axial and radial elements makes possible the evolution and application of simple working rules. Of first importance is appraisal of the permissible angular lay of the wire referred to a line perpendicular to the winding axis. The maximum usable angle is principally dictated by the frictional coefficients of materials; in the first layer, the coefficient between the wire and the form or other base and, in succeeding layers, the self-coefficient of the wire is involved. Arbitrary angular limits based upon group classification of materials have been satisfactorily confirmed in practice. The minimum allowable angle is governed by the tendency of turns in upper layers to wedge between turns in lower layers spreading and ultimately disrupting the winding and by the loss of mutual support among turns expected of the binding action characteristic of the interlaced pattern. The assignment of minimum limits is empirically based from observation of the effect of the angle upon the excellence of windings.

A plane projection of an individual turn shows that the winding angle is the inverse tangent of the ratio of the shuttle-cam throw to the circumferential length of one crossover (Figure 10-19). If the circumference (or diameter) at any layer and the cam throw are known, or postulated, the number of crossovers per turn for any angle can then be calculated. Since the basic design requirement, as to the number of crossovers, is to find a configuration which will have started to wind well at the start, or inside diameter, and which will not yet wind badly at the finish, or outside diameter, it can be seen that the greatest number of crossovers at the start (maximum winding angle) and the least number of crossovers at the finish (minimum winding angle) should be evaluated and set as limiting conditions. A little reflection will reveal that the process can also be reversed. For a given number of crossovers-per-turn, the smallest and largest diameters for

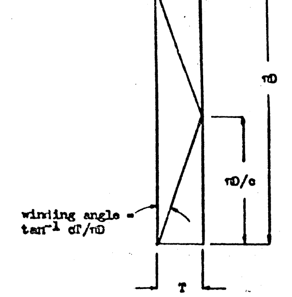

Figure 10-19. Plane projection of an individual turn

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POOR ORIGINAL

satisfactory winding can be established. As a corollary, the maximum winding height upon a given form diameter can be determined. Likewise, if the maximum number of starting crossovers is smaller than the minimum number of finishing crossovers, the solution is untenable; the ratio of diameters exceeds that which can be wound within the specified angular limits.

When the computed number of starting crossovers is greater than four, a secondary adjustment must be made. The need for this can be understood from the following analysis: Assume that a length of wire is grasped in both hands and looped over a cylindrical surface (i.e.: coil-form). Assume, further, that slight pulling forces are applied by the hands along parallel or convergent lines. If the wire on the cylinder's surface opposite the location of the hands is now arranged by an assistant in a manner similar to one cycle of two crossovers in a winding, between the points of tangency, the frictional coefficient between the wire and the cylinder will resist the pulling forces tending to realign the wire in a straight line. If the crossover angle is not excessive and the applied forces are nominal, the resistance will be successful. Now assume that the hands are spread progressively apart so that the forces are applied in increasingly divergent lines. The frictional resistance in the manipulated part of the wire will become progressively less effective until, at 180° divergence, it is completely ineffective.

Since the borderline beyond which divergence of forces takes place is at two crossovers in one-half the circumference, it is plain that windings with more than four crossovers per turn are subjected to divergent application of the tensional forces exerted in the wire. To relieve the increased possiblity of slippage unchecked by friction, it is, therefore necessary to reduce the maximum allowable winding angle and, accordingly, the maximum number of crossovers. Since the winding angle diminishes as the winding layers outward, the adjustment need only be made at the start of the winding.

The applications of selected winding angles and minimum frictional self-coefficients for no slippage, with roughly corresponding material groups, are shown in the table of Figure 10-20. The numbers of crossovers coinciding with these angles, expressed in terms of diameter and shuttle-cam throw, are given in the table of Figure 10-21. It should be remarked that, in a properly-formed winding, the length between wire centers is exactly equal to the throw; the symbol T is used instead of W for the sake of clarity. Corrections applicable to solutions giving more than four crossovers are tabulated in the table of Figure 10-22. The relation between the corrected number of crossovers and the diameter/throw ratio is graphically depicted in Figure 10-23.

Winding Angle (°)	Application	Minimum Slipless Coefficient	Material Groups
5	absolute minimum; requires care in handling	–	–
6	recommended minimum; generally satisfactory	–	–
7	desirable minimum; excellent winding quality	–	–
9	desirable maximum; suited to all materials	0.16	resinous coatings
12	recommended maximum; absolute for 0.16	0.21	synthetic fibers
15	absolute maximum; requires care in handling	0.27	natural fibers

Figure 10-20. Recommended winding angles.

Winding Angle (°)	Crossovers-per-turn Precise	Crossovers-per-turn Approximate
5	0.275 D/T	3 D/11 T
6	0.330 D/T	D/3 T
7	0.386 D/T	5 D/13 T
9	0.498 D/T	D/2 T
12	0.668 D/T	2 D/3 T
15	0.812 D/T	5 D/6 T

D: winding diameter
T: shuttle-cam throw (same units as D)

Figure 10-21. Winding angle vs. crossovers expressed in terms of diameter and cam throw.

Crossovers-per-turn Preliminary	Crossovers-per-turn Corrected
4.00	4.00
5.26	5.00
6.93	6.00
8.95	7.00
11.31	8.00
14.00	9.00
17.01	10.00
20.35	11.00
24.00	12.00
27.97	13.00
32.26	14.00
36.88	15.00

Figure 10-22. Corrections for crossovers-per-turn.

It will be apparent that more than fifteen crossovers-per-turn cannot be used in any normal winding. The computed number of crossovers will usually turn out to be a complex mixed number. A complex number of crossovers will create a confused pattern tending to interfere with orderly lay of the wire. Accordingly, the next step to be taken is rationalization of the number of crossovers to the nearest integer or simple fraction lying within the calculated allowable range.

In windings of moderate height, more than one choice may exist; the number nearest to the arithmatic mean of the calculated extremes is preferred. In the ideal case, electrically, of a ½ ratio between height and inside diameter, making the outside diameter twice the inside, the obvious choice is the 6-12° angular winding range which, fortunately, coincides with the recommended maxima. The nominal number of crossovers should be selected from the table of Figure 10-24.

Nominal crossovers: 2/5, 1/2, 2/3,
(1, 4/3, 2, 3, 4, 6, 8,) 10, 12, 14.

Figure 10-24. Nominal Crossovers.

Numbers within the parentheses will take care of the requirements for all properly proportioned windings. Were a winding to be attempted with a nominal number of crossovers, the repeated cyclic excursion of the wire would simply lay turns one on top of another instead of developing layers of adjacent turns. It is necessary to add an increment to the nominal number, in order to form a layer pattern. The increment, which may be of either sign, is a function of the intra-layer wire spacing, which is somewhat greater than the wire diameter.

The magnitude of the increase takes into account such matters as the effect of the helical winding angle upon the axial space occupied by the wire, departure from instantaneous return at the ends of the shuttle-cam throw, deformation of the wire at the crossover junction points, and the formation of radii at the junctions due to limited ductility of the wire. The increase that has been found necessary in practice is seven percent plus 0.6 milinch. Since virtually all universal winding machines utilize simple, one-cycle cams and straightforward transmission from spindle to cam-shaft, shuttle-cam gearing is independent of machine construction. One exception will be alluded to at the conclusion of the discussion of gearing.

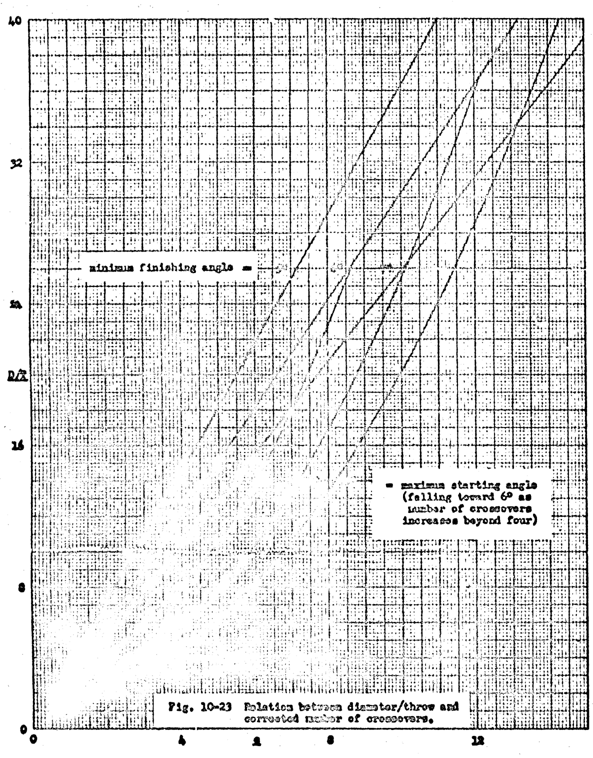

Fig. 10-23 Relation between diameter/throw and corrected number of crossovers.

Having tentatively adopted a set of winding dimensions, wire size, and a number of crossovers, the first step in gearing computation is to weight the wire size with the spacing factor, by Equation 20.

$$e = 1.07h + 0.0006'' \qquad (20)$$

e: weighted wire diameter (inches)
h: maximum specified wire diameter (inches)

Next, it is necessary to determine the nominal number of turns-per-layer, using Equation 21.

$$m = \frac{U}{e} \qquad (21)$$

m: nominal turns-per-layer
U: compensated shuttle-cam throw
e: weighted wire diameter

The compensation of the shuttle cam throw consists of an allowance of 0.005" negative tolerance, i. e.: $U = T - 0.005''$, T being the nominal throw. A winding with an integral number of turns-per-layer may tend to wind poorly because the crossover junctions lie directly one above another; wire deformation can cause some polygonal distortion of the normally round winding outline. A mixed number of which the fractional component is ¼ or ¾ produces a superior winding. Equation 22 indicates the desirable adjustment.

t = m reduced to next lower number having form, $n \div 4$ (22)

t: actual turns-per-layer
m: nominal turns-per-layer
n: any odd number

Now, inasmuch as each spindle rotation produces two crossovers, the nominal ratio of spindle and cam-shaft speeds is equivalent simply to one-half of the number of crossovers. To this ratio must be added an increment whose value depends upon the number of turns-per-layer and the number of turns by which adjacent turns in the layer are removed; the latter number is conveniently found by taking the denominator of the simplest fraction by which one-half the number of crossovers can be expressed. Equation 23 is used to determine the actual shaft speed relation.

$$\frac{b}{a} = \frac{c}{2} \pm \frac{1}{2tv} \qquad (23)$$

a: spindle rotational speed
b: cam-shaft rotational speed
c: crossovers-per-turn
t: turns-per-layer
v: denominator of simplest fractional equivalent of c/2

The simplified step-by-step procedure, outlined above, may be consolidated when an understanding of the entering factors has been gained. The computation can then be undertaken by means of composite Equation 24.

$$\frac{b}{a} = \frac{c}{2} \pm \frac{(0.535h + 0.0003'')}{v\,(T - 0.005'')} \qquad (24)$$

a: spindle rotational speed
b: cam-shaft rotational speed
c: crossovers-per-turn
h: maximum wire diameter (inches)
v: denominator of simplest fractional equivalent of c/2
T: shuttle-cam throw (inches)

In order to comply with the desirable condition that the number of turns-per-layer should be one-half more than a whole number, the incremental part of Equation 24, with the factor v omitted, should be reducible to a fraction whose numerator is two and whose denominator is any odd number.

The selection of gears is made according to Equation 12 or 13. Compound gearing is not always provided for in universal winding machines. In one type, a unique incremental movement is used as a substitute; the instructions of the manufacturer should be followed.

The choice of sign, in the addition of the incremental term of Equation 24, is not of first importance. It will be observed that, when the winding angle is large, particularly near the finish, a negative increment will tend to produce a better winding; the converse is true when the angle is small, the more usual case. For sectionalized windings, a negative increment is preferable, giving needed support to the wire as it leaves the top of one section to resume winding at the base of the next section; for best results, therefore, the widest practicable angle should be used in sectional windings.

Windings produced by gearing based upon positive incrementation are frequently called retrogressive windings, since each turn appears to fall behind the previous adjacent turn in the pattern. For the analogous reason, windings resulting from the use of a negative increment are known as progressive; this use of the word should not be confused with its meaning in respect to progressive-universal windings.

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It is sometimes thought that abnormally wide turn spacing, relative to wire size, can be used to reduce distributed capacitance, thereby increasing Q, in a long winding. This practice is not sound; a loose, spongy structure, which is hard to reproduce, is the result. Q improvement is illusory, since turns tend to fall into the large interstices of the winding, promoting proximity between turns in normally remote layers. A correct solution, when fewer turns-per-layer are indicated, is simply to use a shorter winding. A physically unsound winding would be unjustified, even if electrical performance were improved. Clear indications of a successful winding design are good self-supporting ability, clean rectangular outline, and well defined patterns at the surface and in the side-walls.

Since every change in winding and wire dimensions may affect the choice of the number of crossovers and is sure to alter the gearing, several revisions of the first computations may be necessary. This is not to be regarded as an indication of basic fallacy or as an unreasonable burden; it is merely the normal course of design necessitated by the many controlling requirements to be satisfied or compromised. Skill arising from experience and the accumulation of data from recurring similar tasks will enable rapid and sound judgements to be made by simple interpolations. The serious designer will prepare for himself his own pragmatical working rules as he progresses, thereby greatly reducing the labor involved in universal winding design as well as in other phases of coil engineering.

PROGRESSIVE-UNIVERSAL WINDING DESIGN

In its usual applications, the most favorable length/diameter ratio for the progressive-universal winding is, as for solenoids, nearly 1/1. The inside diameter is set so that the outside diameter, inside diameter plus twice the winding height; does not exceed the usual restrictions. The height cannot be more than about one-half the shuttle-cam throw; otherwise, the tendency for wire slippage toward the base becomes uncontrollable. The shuttle-cam throw should be no greater than necessary to support the required height.

The limited height dictates the number of effective layers that can be wound, as a function of wire size. The number of turns required is obtainable by Equation 25 with good accuracy: Equation 26 yields a fair approximation. For typical proportions in which length equals diameter, shuttle-cam throw is one-fifth the diameter, and height is one-half the shuttle-cam throw, the simple form of Equation 27 may be used; for other proportions, the accuracy is poor.

$$N = \sqrt{\frac{L}{(D+H)\left[\frac{D+H}{18(D+H)+40(W-T)} - \frac{H}{63(W-T)}\right]}} \qquad (25)$$

$$N = \sqrt{\frac{L\,[\,71D + 36H + 160\,(W-T)\,]}{2D+H}} \qquad (26)$$

$$N = \sqrt{\frac{46L}{D}} \qquad (27)$$

N: number of turns
L: inductance (microhenries)
D: winding inside diameter (form o.d. - inches)
H: winding height (same units as D)
W: winding length (same units as D)
T: shuttle-cam throw (same units as D)

The effects of other components, such as shields, ferromagnetic cores, and shells, should be taken into consideration. Since the height/diameter ratio is always small, the presence of a core does not noticeably influence the winding shape. From a first estimate of turns, corrections can be made by alteration proportionate to the square root of the required inductance change, provided length remains constant. A basis for wire choice is hard to find, as in the case of universal windings, but somewhat greater flexibility exists. If larger wire is needed than was first thought, the increased height, as such, does not represent a serious problem. However, a greater shuttle-cam throw may be required, introducing higher distributed capacitance and dielectric loss, eventually nullifying the reduced conductive loss of the larger wire.

The distributed capacitance of a well-proportioned winding made with minimum throw is very roughly that of a similar-sized solenoid times the number of effective layers; this number may be defined as the winding height divided by the nominal wire diameter, measured over insulation.

The use of Litzendraht is advisable in all typical progressive-universal winding applications. Insulation and wire-size requirements for reasons other than high Q will occasionally enter, as for solenoids. As a result of the composite movement of the wire in the formation of a progressive-universal winding, it travels up and down in an inclined plane in alternate excursions. This makes support much more difficult to obtain and, accordingly, slippage more likely. The choice of winding angle is, therefore considerably restricted; the desirable maximum angle, 9° (See Figure 10-20), should not be exceeded and a minimum of 6° should preferably be maintained. Since this permits a height-diameter ratio of ¼, it imposes no great hardship. Other considerations leading to the adoption of a satisfactory number of crossovers-per-turn are much like those applying to universal windings.

For the computation of gearing, universal winding procedure is employed, with one exception. On one or the other excursion of the wire, depending upon pattern progression or retrogression the lead-per-turn opposes the wire spacing provided by the incremental expression in Equation 24. The number of crossovers has no bearing on this loss of space. A compensating term, not divided by the factor v is added in compensation. The modified formulation is given by Equation 28.

$$\frac{a}{b} = \frac{c}{2} \pm \frac{(0.535h + 0.0003")/v + 0.5s}{T - 0.005"} \qquad (28)$$

a: spindle rotational speed
b: cam-shaft rotational speed
c: crossovers-per-turn
h: maximum wire diameter (inches)
v: denominator of simplest fractional equivalent of c/2
T. shuttle-cam throw (inches)
s: lead per turn (inches)

Adherence to a number of turns-per-layer equal to an integer plus ¼ or ¾ is not important, since crossover junctions are continually shifting axially. Shuttle-gear selection is exactly the same as for universal windings and the transverse movement gearing is calculated, according to lead-per-turn, as for a solenoid machine employing the same sort of mechanism. Variable-pitch progressive-universal windings are paradoxical in that proper compensation of the shuttle gearing for axial travel is impossible. Since the most critical portion of the winding is in its greatest height, where the lead-per-turn has the least spreading effect, gearing should be computed for this region and, if necessary, slightly modified in a loosening direction to better accomodate the balance of the winding.

TYPICAL DESIGN EXAMPLES

Example A

Required, a 4.2 microhenry inductor to operate at 10 megacycles, wound on a ¾ inch diameter form. The wiring inductance is .1 microhenry.

Steps

a. selection of winding type; solenoid indicated both by inductance and frequency (See Figure 10-18).

b. setting of dimensions; winding inside diameter equals form diameter; in absence of restrictions, length set equal to diameter ¾".

c. allowance made for wiring inductance;

$$4.2 - 0.1 = 4.1\ \mu h$$

d. computation of turns; by Equation 3,

$$\sqrt{\frac{58(4.1)}{0.75}} = 17.8 \text{ turns}$$

e. computation of lead-per-turn; length divided by turns, 0.75"/17.8 = 0.0421"

f. computation of wire diameter, by Equation 5,

$$\frac{0.0421"}{\sqrt{\frac{10\,(0.75)}{30\,(0.75)}}} = 0.0267"$$

g. selection of wire size; from wire table (appendix), 0.0267" = ABG 22 = 0.0253"

h. estimate of Q; by Equation 7,

$$(100)\,0.75\sqrt{10} = 237$$

i. estimate of distributed capacitance; by Equation 8, doubled for grounding,

$$1.2\,(0.75)\,2 = 1.8\ \mu\mu f$$

j. computation of shaft speeds, using rack-drive machine, by Equation 10

$$\frac{\text{auxiliary shaft speed}}{\text{spindle speed}} \text{ is } \frac{0.0421}{0.010472} = 4.0202$$

k. selection of gears; with 20 to 120 tooth gears available and no compounding, by Equation 12, with aid of slide-rule or gear table,

$$\frac{\text{spindle gear}}{\text{auxiliary shaft gear}} \text{ is } 4.0202 \approx \frac{117}{29}$$

Example B

Required, a 750 microhenry inductor for operation at 0.455 megacycles with a Q of 60 in an aluminum shield 1.250 inches square inside and having a .156 inch corner radii. Coil form diameter is available in steps of 1/8 inch.

Steps

a. selection of winding type; universal indicated by both inductance and frequency (See Figure 10-18).

b. setting of dimensions; by simple arithmetic, equivalent round shield has 1.506" inside diameter;[1] optimum outside winding diameter[1] is 1.506"/2 = 0.753"; optimum inside winding diameter[2] is

1. See article, "Winding Design", page
2. See article "Universal Winding Design", page

$0.753''/2 = 0.376 \approx 3/8''$ $(0.375'')$; height is $0.375''/2 = 0.187''$; length (throw), by Equation 14,

$$\frac{0.187''}{1 + \sqrt{8\,(0.455)}} = 0.062''$$

c. inductance correction made for shield; by Equation 1,

$$\frac{750\ \mu h}{1 - \left[\frac{750}{1506}\right]^3} = 855\ \mu h$$

d. computation of turns; by Equation 17,

$$\sqrt{\frac{\left[(21)(0.375) + (20)(0.062)\right]\ 843}{0.375}} = 234 \text{ turns}$$

e. selection of crossovers; by tables of Figures 10-20 and 10-21, inspection shows suitable maximum angle to be 12° and minimum angle to be 6°; from 12°, the crossovers per turn are

$$\frac{0.668\,(0.375)}{0.062} = 4.04 \text{ max; from } 6°,\ \frac{0.330\,(0.750)}{0.062} =$$

3.99 min; mean is 4.01 ≈ 4

f. estimate of wire diameter; by Equation 19,

$$0.9\ \sqrt{\frac{0.187\,(0.062)}{234}} = 0.0063''$$

(note that I.D. of winding is used with maximum angle and O.D. of winding with minimum angle).

g. selection of insulation; enameled, single-silk covered wire dictated by other considerations such as availability, experience and cost.

h. selection of wire size; from wire table, $0.0063'' \approx$ AWG 38 SSE $= 0.0060''$; maximum diameter is $0.0067''$.

i. weighting of wire diameter; by Equation 20, $(0.0067'')(1.07) + 0.0006'' = 0.00777''$.

j. computation of nominal turns-per-layer; by Equation 21,

$$\frac{0.062'' - 0.005''}{0.00777''} = 7.34 \text{ t-p-l}$$

k. adjustment of turns-per-layer to integer plus ¼ or ¾ Equation 22; 7.25 t-p-l.

l. computation of shaft speed ratio; by Equation 23, assuming retrogressive winding,

$$\frac{\text{cam shaft speed}}{\text{spindle speed}} \text{ is } \frac{4}{2} + \frac{1}{2\,(7.25)\,1} = \frac{60}{29}$$

m. selection of gears; with 20 to 120 tooth gears available and no compounding, by Equation 12,

$$\frac{\text{spindle gear}}{\text{cam shaft gear}} = \frac{60}{29}$$

alternatively, combining steps (i) to (m); by Equation 24,

$$\frac{\text{cam shaft speed}}{\text{spindle speed}} \text{ is}$$

$$\frac{4}{2} + \frac{\left[(0.535)(0.0067'') + 0.0003''\right]}{1\,(0.062'' - 0.005'')} = 2.06897$$

then, selection of gears; by Equation 12,

$$\frac{\text{spindle gear}}{\text{cam shaft gear}} \text{ is } 2.06897 \approx \frac{60}{29}$$

Conclusion

The Q of an experimental coil wound with the calculated gear ratio and turns was found to be 58 at 0.455 Mc. This was considered acceptable for the required Q of 60.

If the measured Q had been outside of the prescribed tolerance a recalculation based upon the following additional step would have been necessary.

For a high-Q case; of say 75 (step n_h):

Consult "Wire Size vs Effective Q" charts in Appendix; for this particular case use A-22. These curves are based upon various wire sizes and cam throws all with SSE wire. Since there is no curve for a cam throw of .187 inches, it is necessary to interpolate between the 1/8" and 1/4" cam curves. 38 SSE gives a Q of 68, whereas 40 SSE would give about 55. (These curves cannot be used for exact numerical data but only to indicate a trend, since they were not prepared to include identical parameters such as coil form diameter etc., used in this example). Using the ratio of 68/55 we can expect the Q of 75 to be reduced to near 60 if No.40 SSE wire is substituted for the No.38 SSE originally chosen.

For low-Q case, of say 45 (step n_l):

Consult Figure 1-3, page 1-5, noting that No.37 wire and 5/44 Litz have equivalent area in circular mils. Assuming they are served with silk in a similar fashion it can be expected that they will wind coils of approximately the same diameter, using a given machine set-up.

Consult wire table in Appendix and note that 4 strands of No.44 have the same cross-sectional area as one strand of No.38. Based upon the relative Q increase shown in Figure 1-3, it is probable that the required Q of 60 will be obtained. (Since 4 strands of 44 Litz is an uncommon size, it would be more practical to use 5 strands of 44 and make a redetermination of the gear ratio to accommodate 5/44 Litz).

Variation of Shield Size:

If space permits, it is possible to increase the Q of a shielded winding, to a certain extent, by increasing the cross sectional area of the shield.

Problem:

In this case determine to what extent the Q of 45, previously mentioned, can be made to approach the required 60 by increasing the size of the 1¼" square shield and what smaller shield is required to depress the Q of the winding having a Q of 75 to the required 60.

Solution, low Q case:

Reference to Figure 2-9, on page 14 of Section 2, indicates that the loss of Q due to the shield decreases as the shield size increases, i.e., the distance from the center of the winding to the inside of the shield increases. Choose winding "A" (curve A) as being the nearest to the winding previously developed in this example. Then the distance between the outside of the developed winding and its shield is

$$\frac{\text{Shield ID} - \text{Coil OD}}{2} = \frac{\frac{40''}{32} - \frac{24''}{32}}{2} = \frac{8''}{32}$$

The curves of Figure 2-9 are plotted against, "center of winding to inside of shield". To convert the distance between, "outside of winding to inside of shield", to "center of winding to inside of shield", it is necessary in this particular case to add half the diameter of coil "A" to the 8/32" dimension calculated above, i.e., 8/32" + 8/32" = 16/32".

Now, using 16/32" as the starting point on Curve A, we see that the loss of Q is 5½% from the Q of the same coil without a shield. Since complete removal of the shield would only produce a Q of approximately 48, [45/.945 = 47.6] which is far below the required Q of 60, it is evident that the only solution is to change the wire from solid to Litz, as was previously suggested.

Solution, high Q case:

$$\text{Required loss} \quad \frac{75 - 60}{75}\,(100) = 20\%$$

This can be attained if the distance between, "center of winding to inside of shield", is approximately 12/32" (curve A). Translating to, "outside of winding to inside of shield" for this problem, we see that the distance between the OD of coil A and the shield to produce a 20% loss of Q is 12/32" − 8/32" = 4/32". Adding this distance to the diameter of the winding developed in this example (24/32" + 4/32" = 28/32"), we see that a shield of 7/8" square would depress the Q of 75 to the required 60.

It was previously stated in this section that good practice dictates a shield diameter equal to twice the winding diameter. Depression of Q by the close proximity of the shield is not only uneconomical from a standpoint of wire usage but tends for lack of uniformity due to slight variations in winding and shield dimensions. This practice, even though workable, is not recommended.

[redacted] These examples [redacted] merely illustrate the application of systematic procedural rules in a consecutive manner. The individual designer will, in the course of time, reformulate and simplify many of the steps in accordance with his abilities and his usual approach to mathematical solutions.

MISCELLANEOUS INFORMATION

One of the most useful tools for coil winding is a spring balance with about a one pound range. This can be used to measure and regulate wire tension and button pressure in objective terms. Of some value is a stroboscope, with which it is possible to detect mechanical resonances in machines and the cause of flaws in windings.

Solenoids are sometimes wound with a wide space injected into the winding near one end; the spinning of turns in this area provides a simple means of accurate inductance adjustment. Special mandrels with spring-loaded inserts are used for this purpose while, in cam-drive machines, an abrupt break in the cam contour can accomplish the same discontinuity.

In setting the usual geared type of turns counter back to zero, it is best to turn the hands in a counterclockwise direction, in order to take up backlash which can cause an erroneous reading.

When wire kinks easily or springs uncontrollably off the spool during winding, reversal of the spool in the holder will usually correct the trouble.

Instead of passing wire through a liquid bath by means of guides, possibly causing work-hardening, the liquid may be applied by means of a wick kept saturated in the winding liquid.

All wire should be inspected for conformity with published standards. Wire which is outside specified diametral tolerance limits or which is not properly covered or annealed may cause an otherwise good winding to be condemned or an inferior wind-

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ing to appear acceptable. The tolerances on wire, as well as on forms and other assembly members, should be carefully considered when a winding must fit into a restricted space.

It is sometimes desired to estimate the quantity of wire used in a winding. The quantity, in length, is easily obtained by Equations 29 or 30, for solenoids or universal windings, respectively. A reasonable allowance should be made for wire used for anchoring the start during winding and for terminations. The quantity can be converted to weight through consultation of wire tables.

$$B = 0.265\,(D + d)\,n \qquad (29)$$

$$B = 0.265\,(D + H)\,n \qquad (30)$$

B: wire quantity (feet)
D: winding inside (form) diameter (inches)
H: winding height (inches)
d: nominal wire diameter (inches)
n: number of turns

NOTATION

Consistent notation has been used throughout this chapter. However, due to the large number of symbols required, no attempt has been made to follow conventions used elsewhere, except for the electrical symbols of C, F, L, and Q. Other than ω, the rather pretentious use of Greek letters has been avoided in consideration of those not scholastically prepared to use them. Possibly confusing subscripts have not been employed. For quick reference, all notation is recapitulated below.

a – spindle rotational speed
b – auxiliary, worm, or cam shaft rotational speed
c – crossovers-per-turn
d – nominal wire diameter
e – weighted maximum wire diameter
f – auxiliary, worm, or cam shaft gear
g – spindle gear
h – maximum wire diameter
j – gear of compound set meshing with spindle gear
k – gear of compound set meshing with auxiliary, worm, or cam shaft gear
m – nominal turns-per-layer
n – any whole number
p – auxiliary shaft to cam or pinion shaft reduction ratio
q – number of sections
r – cam rise in 360°
s – lead-per-turn
t – actual turns-per-layer
u – interface spacing
v – denominator of simplest fractional equivalent of c/2
w – worm lead-per-turn
B – wire quantity
C – capacitance
D – winding inside diameter
E – winding outside diameter
F – frequency
H – winding height
L – inductance
N – number of turns
Q – figure-of-merit ($\omega L/R$)
S – shield inside diameter
T – nominal shuttle-cam throw
U – compensated shuttle-cam throw
V – winding section length
W – winding length

Wherever possible, units of measurement have not been specified; it is frequently convenient, for example, to operate with milinches, or thousandths of an inch, instead of inches. No metric dimensions have been used. All equations are of proven accuracy sufficient for use in practical coil designing. Those who wish to inquire further into the more rigorous classical formulae from which they are mainly derived, should refer to the bibliography, which follows.

BIBLIOGRAPHY

Nagaoka, H., The Inductance Coefficient of Solenoids, Journal of the College of Science, Tokyo, August 15, 1909, page 1

Rosa, E.B. and Grover, F. W., Formulae and Tables for the Calculation of Mutual and Self Inductance, NBS Scientific Paper 169, January, 1912

Radio Instruments and Measurements, NBS Circular 74, 1924

Simon, A. W., Winding the Universal Coil, Electronics, October, 1936, page 22

Joyner, A.A. and Landon, V.D., Theory and Design of Progressive Universal Coils, Communications, September, 1938, page 5

Hershey, L.M., The Design of the Universal Winding, Proc. IRE, September 1941, page 442

Kantor, M., Theory and Design of Progressive and Ordinary Universal Windings, Proc. IRE, December 1947, page 1563

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Section 11

TYPES OF CONSTRUCTION

GENERAL

Reduced to its simplest state, a high frequency transformer need consist merely of a tapped inductance in parallel with sufficient capacitance to produce resonance at a desired frequency. Under certain circumstances this capacitance may come entirely from tubes or circuit strays, leaving the tapped inductance, a means of mounting, and provision for three connections as the only components of the transformer — assuming it to be untuned and unshielded.

Between this simplest of transformers and the complex, temperature-compensated, double-or triple-tuned, high-Q units used in certain equipments are to be found examples of many fundamental types of construction. Among these may be listed the patented constructions which represent the ideas of a particular individual or of a company. Also to be noted are those basic designs which have been in common use for years and which with slight modifications make up the bulk of the high frequency transformers being produced today. Any classification of transformer types must include shielded and unshielded, tuned and untuned, air core, iron core, cup core, fixed capacitance, variable capacitance, and bus-bar construction, as well as the various patented structures.

For the most part, radio frequency transformers are of the shielded variety. While there are exceptions, particularly in the case of untuned or single-tuned types, nearly all double-tuned transformers are enclosed in shield cans, the physical size of which have a pronounced influence not only upon the mechanical construction but also upon the electrical parameters of the units which are enclosed within the cans.

By far the majority of high frequency transformers have at least one winding capable of being tuned. The most commonly used varieties have two windings which are tuned. The tuning range will vary somewhat with the specific needs of each individual design, but in most cases provision must be included for varying frequency at least 10 per cent from the design center.

Since the resonant frequency of a transformer is a function of the product of the inductance and the capacitance (commonly called the LC ratio), it follows that tuning may be accomplished by varying the inductance, the capacitance, or both. Since normally only L or C are variable, it becomes possible to divide all common types of transformers into those which are permeability-tuned (variable inductance) and those which are trimmer-tuned (variable capacitance).

TRIMMER-TUNED TRANSFORMERS

For many years, trimmer-tuned i-f's made up the bulk of production of this general type of high frequency transformer. Fig. 11-1 shows a popular version of this design. The tuning capacitances were provided by a dual mica trimmer of the type shown in the illustration. The base (1) was usually made of ceramic (Steatite), although some manufacturers did at times use plastics of both thermosetting and thermoplastic types. This particular part was designed in such a way that it not only provided the base upon which to build the capacitors but in addition carried two bosses (2) which served to locate the assembly within the shield cans and another boss (7) over which could be fitted the coil forms. The capacitor elements were made up of interleaving plates (5) of metal between which were pieces of plain mica. Position of the plates was controlled by the screw (3) which was insulated from the plates by ceramic or mica washers (4). Tightening the screws would bring the plates into closer proximity, thereby increasing the capacitance of the trimmer.

Several variations of this trimmer have appeared at one time or other. The size of the base has

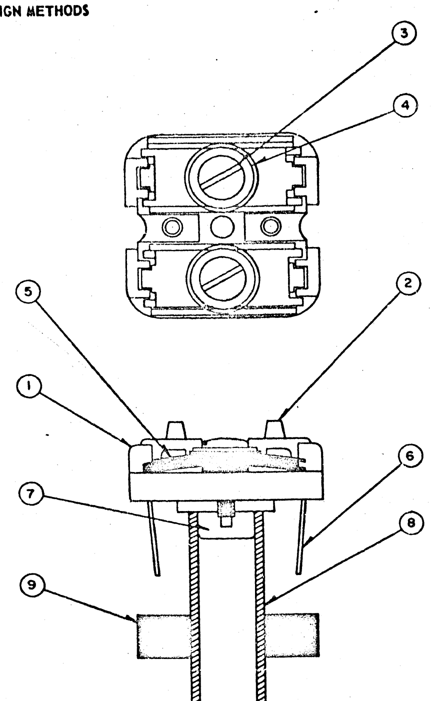


Fig. 11-1 Trimmer Tuned i-f

ranged between 3/4 inch square and 2 1/2 inches square. The number of plates has varied from 2 to as many as 10 or more. Since large numbers of plates complicated the assembly and often adversely affected the temperature stability of the capacitor, some manufacturers would build up part of a large capacitance value out of silvered mica pieces and then finish the stack in the conventional manner. This made a capacitor element having a certain fixed capacitance in parallel with an adjustable capacitance of a value sufficiently large to permit tuning over the desired frequency range.

Among the many modifications of this basic design was that which has a threaded stud extending upwards from the surface of the capacitor base by which the entire assembly could be firmly attached to the shield can. Those units designed to operate in the diode stage often would have the trimmer capacitors designed with the diode filter capacitors built in as separate fixed sections of one side of the tuning capacitor, thus reducing the number of parts required to complete the circuit.

DESIGN LIMITATIONS

Certain basic faults were inherent with this type of construction. The capacitor was an open type which could not be sealed against the entrance of moisture. Silicone applied to the base, plates, and mica improved the moisture resistance although nothing could render this type of capacitor immune against the effects of moisture. Lack of temperature stability was a persistent problem that was only partially solved by careful selection of plate material, design of plates, and cycling (heat treating) of the assembled capacitors. To the above troubles can be added the fact that assemblies of this type were unnecessarily expensive because of the large number of small, delicate parts which were involved. It is little wonder, then, that the industry was quick to overlook the advantages of this type of construction which included uniformity of coupling — that is, no change in mutual inductance due to tuning and the ability to tune both primary and secondary circuits from one end of the assembly — and to embrace the simpler and cheaper type of construction made possible by the introduction of permeability tuning.

PERMEABILITY-TUNED TRANSFORMERS

As was pointed out in Section 3, the idea of using small particles of iron rather than solid pieces of magnetic materials as cores within inductances has been known for many years. It was, however, not until the work of W. J. Polydoroff resulted in the production of low-loss, powdered-iron cores that important uses for this material developed in electronics. The adaptation of powdered-iron cores offered a new and less complicated method of tuning by varying the inductance rather than the capacitance. Tuning thus became a matter of varying the position of a core within a winding — a process requiring only that a screw be attached to the powdered iron and that some sort of threaded bushing provide a drive for the screw. To be sure, certain disadvantages were apparent. For example, unless a designer was willing to accept the problems that come with side-by-side mounting of coupled inductances, tuning could no longer conveniently be accomplished from one end of the assembly without recourse to complicated, concentric-screw, core-drive mechanisms.

Early design of permeability-tuned transformers were plagued by variations in the coefficient of coupling resulting from changes in core positions during tuning. However, when designers learned to take full advantage of the permeability of powdered iron, it was found possible to run the cores at positions within the windings where tuning would have little effect upon coupling.

One definite advantage of permeability-tuned coils appeared in the higher Q's obtainable with solid wires, thus often making the use of Litz wires unnecessary, and thereby opening a way to substantial savings in cost with no loss in quality of performance. The increase in inductance resulting from the presence of the iron also made smaller coils possible with consequent savings in material, winding time, and space. In fact, it was permeability tuning that actually opened the door to miniaturization, because through the use of powdered iron not only within the coils but also surrounding them, it became possible to duplicate in small spaces the performance of conventional larger units.

CUP-CORE DESIGNS

The powdered iron containers in which whole windings could be enclosed were called *cup cores*. The big advantage offered by these cup cores was to be found in the fact that the outer shell of iron provided a very satisfactory magnetic shield which minimized the effect of conventional shield cans, thus making high-Q, small transformers a practical reality.

The fact that the windings were completely enclosed within powdered iron greatly reduced the magnetic fields surrounding the inductances. Since in most cases magnetic coupling between windings is desirable, the use of cup cores made control of inductive coupling much more difficult. Instead of the coils being separated by a distance of one inch or more as had been common in earlier designs, to produce the required amount of coupling when using cup cores, it now became necessary to separate the coils by distances of only a few thousandths of an inch. Accurate control of this spacing was necessary to insure uniform performance and was usually obtained by the use of spacers made from bakelite or other insulating material. So critical was this spacing, that normal manufacturing tolerances in the thickness of these spacers could cause serious variations in the coupling and consequently in the performance of transformers of this type.

Another problem frequently encountered in the use of cup cores came about as a result of non-uniformity in the air gaps formed by the two parts of the cup cores. The simplest form of cup cores were made with their air gaps at right angles to the magnetic field where even a slight variation in spacing would be reflected in a change in both the Q and the inductance of the enclosed windings. Actually, the effect upon the inductance was less serious than that upon the Q, simply because a cup core accounts for only 15 to 20 per cent of the permeability increase resulting from the iron, while the balance of this increase comes from the center-slug portion of the core. Q is seriously affected by variations in the air gap.

Only a few years ago nearly all i-f transformers were made with flexible leads as the means of connecting into the amplifier circuit. It was standard practice for the customer to specify the length of each lead and also the length of the tinned portion so that the transformer could be connected directly into the circuit without the necessity of cutting and stripping leads. The very fact that the leads were long and flexible automatically made it difficult to control capacitive coupling. As a result, it was difficult to produce transformers having identical performance characteristics, and satisfactory testing methods constituted a serious manufacturing problem. As a means of minimizing variations due to lead lengths, cardboard or fibre lead guides were often provided. Slots cut in these pieces located the leads with a fair degree of accuracy but still did not provide enough support to permit uniform duplication of performance within satisfactory limits.

BUS-BAR CONSTRUCTION

The logical outgrowth of the situation described above was the development of a basic type of construction which eliminated use of flexible leads. Fig. 11-2 shows an example of a transformer of this type which will be seen to consist basically of a rigid framework made of hardened and tinned copper bus-bars (4) soldered into eyelets (2) inserted in two insulating plates (1) which form the ends of the transformer assembly. The windings (7) which are carried on the coil form (5) have their leads soldered (6) directly to the bus-bars to which are also soldered the capacitors (8) which are a part of the resonant circuits. The coil form is supported at its ends by bushings (3) which also serve to drive the tuning cores and in some instances act as a tension device as well. (See Section 4)

External connections to transformers following this basic design can be made either to an extension of the bus-bars (9) or to terminals riveted to the end plate in place of the eyelets. In this latter case, care must be used when soldering to the terminals as the application of too much heat may cause the solder to run out of the eyelet portion of the terminal, leaving the bus-bar without support at that end.

This general type of construction has appeared in a variety of sizes ranging from as small as 3/4 inch square up to some which are 2 or more inches square. Bus-bar construction is particularly well adapted to low frequency operation but has been used successfully at frequencies as high as 10.7 Mc. Beyond this point, difficulties may be expected in holding primary to secondary coupling to satisfactorily low levels because of the proximity of the bus-bars to the windings.

Because bus-bar construction is a fundamentally sound design which by virtue of the definite positioning of its components lends itself to successful duplication, it is but natural that this type of construction should have appeared in a number of variations. Among the more common modifications may be listed those units provided with:

1. One or more windings enclosed in cup cores.
2. Three or more insulated plates and several tuned circuits or other network configurations.
3. Universal windings.
4. Both universal and solenoid windings.
5. Two coil forms located side by side, thus permitting top tuning of variable-inductance-type transformers.

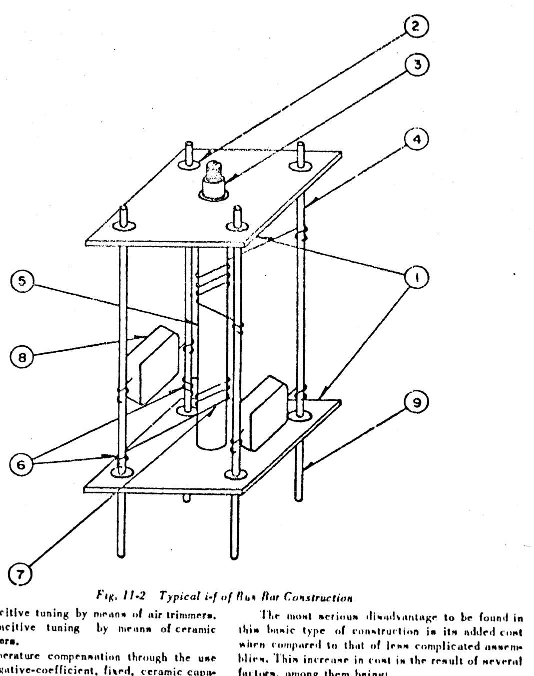


Fig. 11-2 Typical i-f of Bus Bar Construction

6. Capacitive tuning by means of air trimmers.
7. Capacitive tuning by means of ceramic trimmers.
8. Temperature compensation through the use of negative-coefficient, fixed, ceramic capacitors in parallel with silvered mica capacitors.
9. Temperature compensation through the use of special materials selected for their thermal coefficient of linear expansion and used in bus-bars, core screws, and/or core drive mechanisms.

The most serious disadvantage to be found in this basic type of construction is its added cost when compared to that of less complicated assemblies. This increase in cost is the result of several factors, among them being:

1. The large number of component parts required.
2. The large number of soldered connections.
3. The fact that many assembly operations must be conducted in a jig or fixture in order to assure uniformity of performance and proper fit within the shield can.

POOR ORIGINAL

An additional disadvantage – actually of little significance in view of the vastly improved uniformity of performance – is found in the absence of flexible leads which places upon the customer the full burden of connection into the circuit. What was previously a case of soldering four or more connections now becomes a matter of cutting, stripping, and inserting an equal number of leads. However, the advantages so far outweigh the disadvantages that in one form or other this general type of construction has become standard throughout the industry. Improved designs taking full advantage of basic principles have lowered the cost and reduced the size of high frequency transformers and at the same time have actually improved their performance.

PATENTED STRUCTURES

An illustration of a transformer design[2] which incorporates many of the advantages of bus-bar construction but which is admirably suited to large-scale mass production is shown in Fig. 11-4. It will be noted that in this design bus-bars have been replaced by plastic side frames which serve a variety of purposes including:

1. Wire guides to conduct coil leads to transformer terminals.
2. Support for coil forms.
3. Threaded supports providing drive and tension for tuning cores.
4. Location of assembly within shield cans.

The silvered-mica tuning capacitors are located

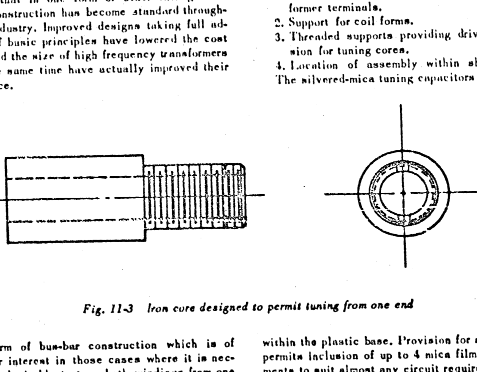


Fig. 11-3 Iron core designed to permit tuning from one end

A form of bus-bar construction which is of particular interest in those cases where it is necessary or desirable to tune both windings from one end is offered by one manufacturer.[1] The distinguishing feature of this design is the iron core used to tune the top windings. As is shown in Fig. 11-3, this core is made with a hole extending the full length of the threaded sleeve and the powdered iron as well. The sleeve is slotted on the top, thus affording a means of adjusting the top core, while the lower core can be tuned in the conventional manner from the bottom of the transformer or by a tuning tool inserted through the hollow top core and engaged in a slot on the upper end of the bottom core. The torque of both cores is controlled by D-spring type tension devices which work on the threaded sleeve of the top core and on the screw of the bottom core.

within the plastic base. Provision for six terminals permits inclusion of up to 4 mica films in arrangements to suit almost any circuit requirement. Capacitance values from as little as 3 upward to 1000 mmf are commercially available in the construction. Normally the capacitors are of the "open-end" type, which is to say that the silvered mica films are merely held between two pieces of plastic, and no attempt is made to seal the assembly against the entrance of moisture. For ordinary commercial use, this type of capacitor has proved perfectly satisfactory. The fact that moisture can enter the assembly with ease also means that it can leave with equal ease, so that while under conditions of high humidity there may be a substantial temporary decrease in the Q of the capacitor, only a few moments of drying are needed to restore the original Q. The addition of a silicone coating to

[1]Aladdin Radio Industries, Inc.

[2]K-Tran manufactured by Automatic Manufacturing Corporation, Newark, New Jersey.

the mica films and to the base assembly minimizes to a considerable extent the adverse effects of moisture.

This particular transformer is of the permeability-tuned type although differing from conventional designs in that it utilizes a threaded (on the outer periphery) form of open cup core which combines the advantages of the conventional cup core with those of the ordinary slug-type tuning cores. By having adjacent ends left open, this type of core avoids the close mechanical spacing and the resultant coupling sensitivity ordinarily associated with cup cores. The reason for this fortunate behaviour seems to be that substantial portions of the magnetic fields of the two windings are so located that the flux lines flow naturally through the iron and are little affected by the air gaps, thus providing good magnetic shielding and high effective permeability.

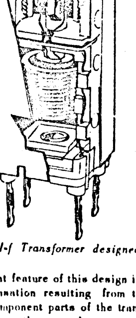


Fig. 11-4 I-f Transformer designed for mass production.

An important feature of this design is the temperature compensation resulting from the arrangement of the component parts of the transformer. In Fig. 11-5 is shown the general arrangement of parts with the planes of the side frames which support the threaded cores being indicated by lines BB' and CC'. AA' represents the axis of the coil form which is, of course, parallel to the planes of the side frames. The coil form (3) supports the two windings (4) and is in turn supported by the side frames at points 2a. XX' represents the plane in which the coil form is supported. It is with this plane as the starting point that temperature changes produce linear expansion within the structure as shown by the arrows parallel to the side frames and the coil form. An examination of this drawing will reveal what happens when the assembly is subjected to an increase in temperature. Both the side frames and the coil form will expand in the direction of the arrows. Because the side frames are longer, they expand more than the coil form with the result that the cores are moved slightly out of the windings, thus counteracting to some degree the increases in the inductances of the coils caused by rising temperatures.

Since the side frames and the coil forms are not commonly of the same material, a further opportunity for self-compensation is offered by this type of assembly. The selection of a material having a low, or more preferably a negative, thermal coefficient of linear expansion for the coil forms, and of another material with a high expansive rate for the side frames will produce greater relative motion between the cores and the windings, thereby providing a higher degree of temperature compensation. Proof of this reasoning appears in the graph shown as Fig. 11-6.

Under the sponsorship of the Signal Corps,[3] a ruggedized, high-temperature transformer was developed for use in military applications. This unit has the same construction as is shown in Fig. 11-4 with these exceptions:

1. The thermoplastic side frames were replaced by frames molded from thermosetting, mineral-filled melamine.
2. Two steel springs instead of one supplied tension to the tuning cores.
3. The open capacitor base was replaced by a mica-filled phenolic molding in which the silvered mica tuning capacitors were completely encapsulated.

By these comparatively simple structural changes accompanied by a new and improved impregnation technique for the windings[4] an inexpensive transformer suitable for operation only up to 85 to 90 C was changed into a version which, while only slightly more expensive, is suited for operation up to 125 C and in addition is capable of withstanding rigorous requirements of shock, vibration, and humidity.

[3]Research and development contract DA-36-039 sc-15321

[4]See Dual-Varnish Treatment – Section 9.

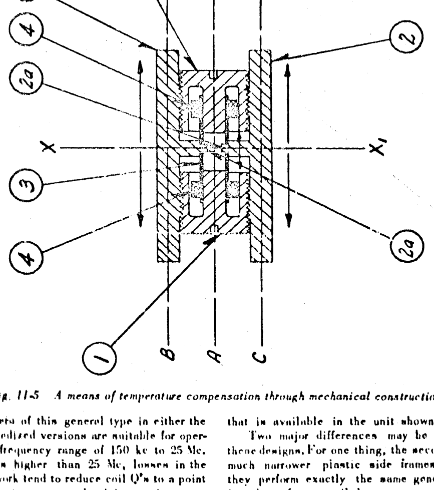


Fig. 11-5 A means of temperature compensation through mechanical construction

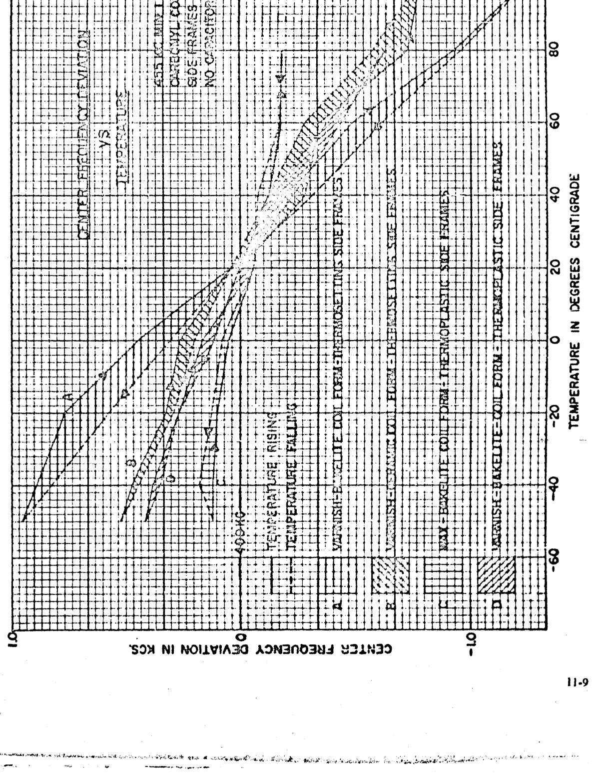
Transformers of this generel type in either the regular or ruggedized versions are suitable for operation in the frequency range of 150 kc to 25 Mc. At frequencies higher than 25 Mc, losses in the plastic framework tend to reduce coil Q's to a point too low to meet average selectivity requirements.

In Fig. 11-7 appears another patented structure[8] possessing many of the characteristics of the above-mentioned design. It will be seen that the coil form (7) is supported at its center by the plastic side frames (3) which support and drive the iron cores (9) by means of molded threads (1) which are held under tension by the steel spring (8). It is obvious that this particular assembly has the same sort of built-in temperature compensation that is available in the unit shown in Fig. 11-4.

Two major differences may be noted between these designs. For one thing, the second design has much narrower plastic side frames which, while they perform exactly the same general functions, introduce fewer coil losses as a result of their smaller mass, thus permitting transformers of this design to operate satisfactorily at considerably higher frequencies. The second major difference appears in the type of iron cores used to tune the inductances. The second design utilizes simple, threaded, slug-type iron cores instead of open cup cores. This, of course, means that windings are not magnetically shielded and are therefore more subject to "shield effects" than are those which are enclosed in cup cores. Naturally this effect is more pronounced in the case of the high-inductance,

[8] J-Tran manufactured by Automatic Manufacturing Corporation, Newark, New Jersey.

Fig. 11-6 Effect of materials of construction upon frequency drift



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large OD windings used at the lower frequencies and is of much less importance in the case of the solenoid windings intended for use at frequencies of 10 or more megacycles.

One point which is sometimes of great importance is the fact that both windings in transformers of the type shown in Fig. 11-7 can be tuned from either end of the assembly. This is possible because the tuning cores have a hexagonal-shaped hole extending throughout their length which permits a special plastic tuning tool to be inserted either into the first core, or through the first core and into the second. It is therefore possible to tune from either or both ends -- whichever is better suited to the particular installation in question.

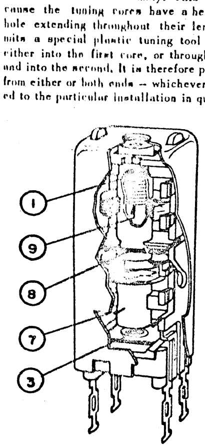


Fig. 11-7 A patented transformer construction

OTHER DESIGNS

One relatively new design intended for operation in the 44 Mc range is interesting in that it employs the "loss-tuning" principle. In this particular design, tuning is accomplished by positioning what is effectively a shorted turn about the windings. In the commercial version, this shorted turn takes the form of an eyelet which is moved by means of an arm attached to a plastic screw.

While offering the advantage of single-ended turning, this design would seem to require major modifications in the matter of supporting and driving the tuning mechanisms before it could be considered for applications where shock and vibration would be important factors.

Another design which enjoyed considerable success a few years ago but which is no longer in production is interesting because of the unique type of iron cores which were used. Made in the popular "3/4 inch" size, this unit was trimmer-tuned and had its inductances random wound directly over the iron in the slot of a dumbbell-shaped powdered-iron core. Using cores with a maximum OD of 0.250 inch and a slot 0.125 inch wide, it was possible to obtain inductances of nearly 2.5 mh with No. 40 HF wire.

A GENERAL DESIGN

A general type of construction which has become common in the coil industry in recent years is represented by the sectional drawing Fig. 11-8. Since nearly every major manufacturer of high frequency transformers has produced his own version of this general design, Fig. 11-8 is meant only to be representative and to show the general principles that go to make up this basic type of high frequency transformer.

Primarily, this general design is made up of three fundamental parts, consisting of a base assembly (1), a winding assembly (2), and a shield (3). Almost without exception, this type of transformer is made only as a permeability-tuned model, and its acceptance has been wide in high-frequency applications where absence of supporting frameworks has helped to eliminate losses and thereby to improve performance. As will be pointed out later in this discussion, the extreme simplicity of this design with its lack of built-in support for the coil leads tends to make it somewhat difficult to maintain uniform coupling between windings.

As shown in Fig. 11-8, the base assembly consists of two plastic plates which enclose one or more sheets of silvered mica which make up the fixed capacitors. The lower of these two plastic plates contains the terminals, while the top plate serves as a support for the coil form assembly.

In other versions of this same design, the base assembly consists simply of a laminated phenolic plate of a size and shape such as to fit the shield can. In the center of this plate is inserted either a combination coil form holder and core-screw tension device of the type described in Section 4 of this manual or a stud over which the coil form is pressed. The fixed capacitors may be molded silver micas, open silver micas, glass, or ceramic. Bases

of this type are most generally supplied with solder lugs for connection into the amplifier circuit, but many similar units have been made using regulation terminals and even some with leads soldered to eyelets mounted in the terminal board (base).

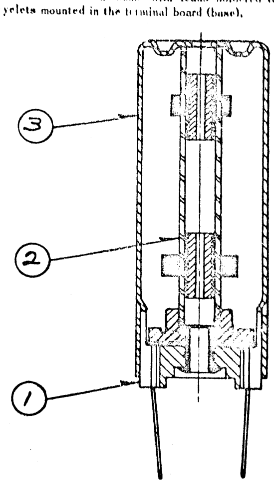


Fig. 11-8 A general transformer construction

Winding assemblies have fallen generally into two classes (1) those in which tuning is accomplished with powdered iron cores attached to metal screws and (2) those tuned by powdered iron cores which are themselves threaded on their outer peripheries.

If the first of these two general classifications is employed, a slight problem is involved in adjusting the upper core. This operation may be accomplished through a threaded stud attached to the top of the shield or by the use of a combination coil form holder and tension device placed over the end of the tubing used as the coil form and locked into a suitable opening in the shield to prevent rotation. This feature is sometimes the cause of trouble since unless the assembly is firmly positioned with respect to the core drive mechanism, detuning will result from any displacement of these two parts.

The second basic type of winding assembly which utilizes threaded cores is also made in two distinct versions. The more common of these depends upon partial (or, in one version, full) threads impressed or cut into the tubing used as the coil form. The major problem connected with this design is found in the matter of tension applied to the cores. Because most of the materials commonly used in coil forms are somewhat hygroscopic, and because changes in moisture content are accompanied by dimensional changes in the tubing, it has proved to be all but impossible to maintain uniform core torque during the course of normal variations in weather.

A considerable amount of work has been devoted to this problem of torque control by manufacturers of the coil forms and by members of the coil industry. Maximum and minimum torque requirements* have been specified by a number of sources, but up to the time of this writing no completely satisfactory solution to the problem of core torque in applications of this sort has been reached. Costwise, there is much of interest in this basic design, but in view of its recognized shortcomings it is not recommended for use in military equipments.

In an attempt to employ this general design without encountering the torque problems discussed above, certain manufacturers have taken the basic principle of torque control incorporated in the D-spring tension device (Section 4) and adapted it to the drive and tension of the tuning cores in this general type of transformer construction. To carry this out, slots are punched in the tubing so as to resemble the slots in the D-spring tension device referred to above, and either regulation D-springs of suitable size, or what might best be described as plastic C-washers, (Fig. 11-9) are inserted around the tubing and through the slots so as to engage the threads of the tuning cores.

From the foregoing description, it will be seen that both drive and tension or torque control are imparted to the cores through the medium of the D-springs or C-washers. No threads are required in the ID of the tubing, and as long as the slots are

*Representative values are 3/4 inch ounce for a minimum and 7 inch ounces for a maximum.

kept clear – thus permitting free action of the spring – tension will be fully controlled by spring action. A major weakness of this design is lack of resistance to "push through" – a condition likely to be encountered in tuning unless particular care is exercised to avoid the application of more force than is actually necessary to engage the tuning tools in the slots of the cores. Another criticism often directed at this design is that the slots for the springs must be located close to the windings if unnecessarily long cores are not to be used. This requirement, particularly in the case of metal D-springs, causes a reduction in the Q of the windings because of the close proximity of even this small mass of metal.

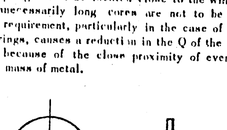


Fig. 11-9 Plastic C-washer used to supply core drive and tension.

Shield assemblies of this general design are fairly uniform in type except for minor features. Included among these differences may be listed the means provided for holding the top of the coil form in place within the shield. This feature is of the greatest importance in those units having threaded cores, since the use of a tuning core with a screw automatically provides a means for locating the top of the coil assembly.

In the case of those units using threaded cores, two general methods of securing the tops of the coil forms are employed. The cheaper of the two makes use of an extruded shape in the shield of a size which will fit either the ID or the OD of the coil form (see Fig. 11-10a & 11-10b) and which will, therefore, conduct the tuning tool to the proper location within the coil form. The second, and probably more common method, uses a stamped coil form holder of the general type described in Section 4, except that the portion intended to engage the screw threads is omitted and replaced with a hole large enough to admit the tuning tool.

A major problem in this type of construction has proved to be the establishment of a satisfactory means of holding the assembly within the shield. Some of the first commercial versions used a combination coil form holder and tension device at the top of the coil form as described above and depended entirely upon this device to hold the

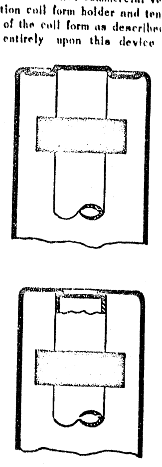


Fig. 11-10a & Fig. 11-10b Common methods of supporting top of coil form.

complete assembly within the shield (see Fig. 11-11). It was reasoned that the main need for this holding action would be prior to installation of the transformer in the chassis and that it would not be necessary to provide a completely rigid assembly. For equipments which will not be subject to shock or vibration, such an arrangement could possibly be considered satisfactory, especially if the length of the coil form and the length of the shield bore a relationship such that the lower surface of the base carrying the winding assembly and the bottom of the shield were in exact alignment. Because of normal manufacturing tolerances, this ideal condition was rarely realized with transformer instability a natural result of shock and/or vibration.

As an alternative to the above listed method, some designers turned to a base plate in which were located holes through which spade bolts in-

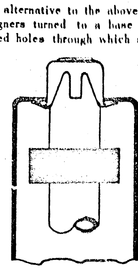


Fig. 11-11 Coil form holder used to support top of coil assembly.

tended for mounting the shield to the chassis could be used to hold the transformer assembly in the shield. From the standpoint of results obtained, this system was satisfactory, but it was expensive both from the viewpoint of materials required and from the extra operations involved.

The generally accepted method in use at this time consists of locating the top of the base assembly against shelf-like projections which are bumped out from the ID of the shield as shown in Fig. 11-8. In this manner, the assembly is definitely located without respect to the bottom of the shield where it may be held by means of a simple crimping operation. A little thought on the part of the designer, particularly with respect to reasonable and practical tolerances of all dimensions, should result in assemblies which are satisfactory in this regard.

As was intimated at the beginning of this discussion, a serious problem with this type of construction stems from the lack of support for the coil leads . Since these leads are surrounded by both electrostatic and electro-magnetic fields, it is obvious that their position with respect to one another will affect the coupling present between windings. Variations in lead position will, therefore, be reflected in variations in performance (response), thus making duplication of units a difficult matter. The extent to which lead position may influence response is probably best illustrated by stating that it is not an unknown practice in the coil industry to adjust gain and band width simply by positioning coil leads. This very fact points out the danger of using this type of construction in any application where it may be expected to encounter shock or vibration – a limitation which, for all practical purposes, eliminates this basic design from all military consideration.

TEMPERATURE COMPENSATION

Increased requirements for uniform gain and band width over a wide range of ambient temperatures make some form of temperature compensation a requirement in many transformer designs. It was pointed out earlier in this discussion that certain patented structures have a degree of temperature compensation inherent within themselves as a result of basic construction principles. This very fact accounts for the exceptional temperature stability manifested by some civilian models – a stability which often exceeds that of more expensive military components.

From the standpoint of a transformer designer, it is unfortunate that all the common and basic components of high frequency transformers tend to move in the same direction when exposed to temperature changes. This general change or drift is in a positive direction, which is to say that as the temperature goes up, so also does the inductance of the windings and the capacitance value of most of the capacitors used for producing resonance.

It has been shown in various places throughout this manual that such factors as wire insulations, coil form materials, and coil impregnation materials have a definite effect upon the temperature coefficients of universal coils. The same general principles apply, although to a somewhat lesser degree, in the case of solenoid windings.

Silvered mica capacitors also have a positive temperature coefficient. This means that the changes in the coil and in the capacitors of those units tuned with silvered mica capacitors are additive, and both being positive must result in a lowering of the resonant frequency as the temperature increases.

Fortunate indeed is the fact that ceramic capacitors are available in a wide range of temperature coefficients, including many which are highly negative. It therefore follows that a proper combination of negative-coefficient, ceramic capacitors with other capacitors possessed of positive characteristics can produce a unit which will be relatively stable over a limited temperature range. The effect of the addition of negative capacitors to a conventional transformer is clearly shown in the graph presented as Fig. 11-12.

As was intimated earlier, proper selection of materials of construction or the utilization of specific mechanical configurations may also offer a means toward temperature compensation. Generally speaking, however, such ideas are expensive to put into operation, and most often the designer will solve his temperature problems through selection of the proper combination of the proper capacitors.

SUMMARY

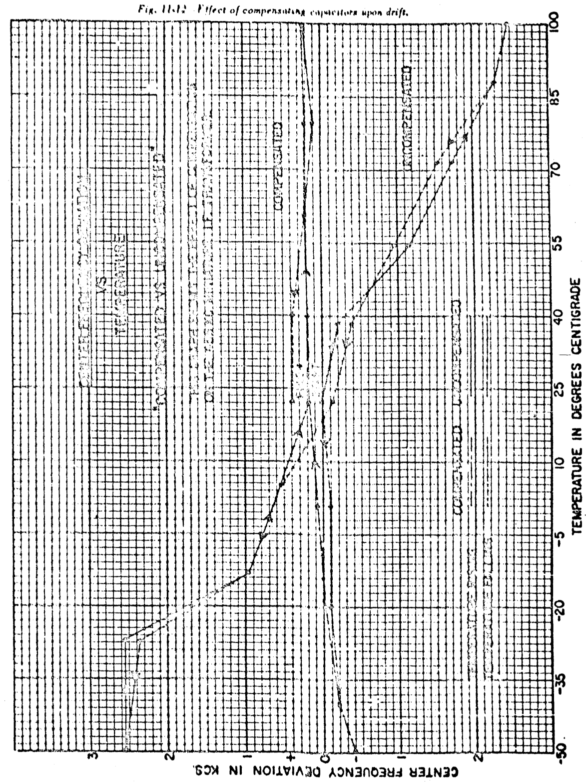
Since the primary purpose of this manual is to provide information leading to the design of high frequency transformers for military use, it is at once apparent that many of the designs discussed above cannot be considered adequate, even though they may give entirely satisfactory performance in civilian applications. Readers who are unfamiliar with the rigorous requirements of the Armed Forces are referred to MIL-C-15305A – a careful reading of which will show why transformer designs which are completely satisfactory for civilian radios or television sets are entirely unsuited for military use.

In view of the conditions under which end equipments must operate, it is not surprising that a majority of those transformers intended for such use are of bus-bar construction. Admittedly expensive, this basic design is versatile enough to allow modification to a degree permitting its consideration in many military applications. Bus-bar construction is sturdy, and in the larger shield sizes has ample room for compensating capacitors or for other network components. For those cases where size is a definite factor, it would appear that the miniature transformer developed under contract DA-36-039 sc-15321 has much to offer. In combination with its special mounting clip described in Section 2, this unit has much to recommend it for general military use – especially in view of the present trend toward miniaturization of equipments.

SPECIFICATIONS

MIL-C-15305A
"Coils, radio frequency; and transformers, intermediate and radio frequency"

Fig. 11-12 Effect of compensating capacitors upon drift.



Section 12

MEASUREMENTS – THEORY AND PRACTICE

GENERAL

Electrical measurements of coils and transformers are of extreme importance to a design engineer. In addition to providing a measure of performance, electrical measurements furnish a means of specifying fundamental parameters, thus opening a way to the successful duplication of coils and transformers.

If electrical measurements could be made with the same degree of ease and accuracy that mechanical measurements can be taken, specification and subsequent duplication of electronic components would be vastly simplified. Unfortunately, however, it is not an easy task to measure in an accurate manner the performance of a high frequency transformer. The seriousness of the problem stems largely from the difficulties involved in eliminating from consideration all the other components in the test circuit, thus insuring readings which relate only to the unit under test.

In general, r-f transformers can be considered as four-terminal passive networks. To conduct transformer tests in line with this concept means that the units must be driven with a known value of signal input while the resultant output is being measured.

From a purely practical standpoint, measurements made in the foregoing manner have comparatively little value because of the fact that almost every r-f transformer is designed to operate in conjunction with a vacuum (electron) tube or transistor. Since the type of tube and the voltages supplied to its various elements along with the general features of the related circuitry have a decided influence upon the performance of a transformer, it follows that measurements taken under other than actual operating conditions may be of questionable value. This unfortunate situation is aggravated as the frequency of operation increases. At low frequencies – and this can mean frequencies approaching even as much as 5 or 10 Mc – measurements under operating conditions can be carried out with a minimum of attention to circuitry and layout. Above this point, particular attention must be given to such details as connections to the tube sockets, arrangements of ground returns, shielding, by-passing of filaments and cathode returns – just to name a few of the critical points. It should be noted in this connection that at the higher frequencies all vacuum tubes will exhibit some degree of input loading in their grid circuits as a result of the Miller Effect.[1] Even with a maximum of care and attention to detail, at frequencies above 100 Mc it is extremely difficult to duplicate measurements. At such frequencies, it is common practice to make all final adjustments to coils and transformers in the actual equipments of which they are a part, rather than to attempt to depend upon the results of tests conducted outside of the actual circuit.

Another situation in which it is generally difficult to duplicate measurements involves the use of high impedances at relatively low frequencies. It therefore becomes apparent that accurate testing and measurement of high-frequency transformers can be complicated by the presence of high impedances and/or high frequencies.

In the early days of radio, it was customary to make all coil and transformer measurements either at d-c or at some very low frequency such as 1000 cycles. To be sure, it was recognized even then that such readings were of necessity somewhat inaccurate, but in view of the relatively low frequencies in use in those days as well as the fact that most inductances were of simple design not involving magnetic cores, little effort was devoted to the development of better methods of measurements.

[1] See Section 14 for more detailed discussion of the Miller Effect.

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The rapid growth of electronics during the past fifteen or twenty years and the changes accompanying this growth have focused attention on the inadequacies of direct-current or very-low-frequency measurements. Higher operating frequencies, reduction in size of component parts, demands for more selective circuits, increased requirements for duplication of inductive components – all these, and many more, are reasons why growing importance is attached to those measurements which are made as nearly as possible under actual working conditions. Because both types of measurements have their place in the coil industry, this discussion will include both low frequency and high frequency equipments and techniques.

LOW FREQUENCY MEASUREMENTS

D-c measurements – particularly of resistance – are often of value in coil design. If only moderately accurate resistance readings are required, an ohmmeter[1] will suffice as the measuring instrument. For accurate determinations of resistance, the standard instrument used is the *Wheatstone Bridge* – a typical circuit diagram for which is shown in Fig. 12-1.

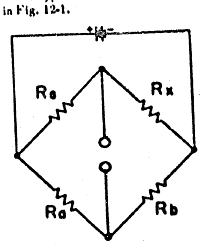


Fig. 12-1 Wheatstone Bridge

The fundamental Wheatstone Bridge circuit is representative of "null" methods of measurement. A network capable of adjustments such as to give zero transmission in the detector circuit is the fundamental requirement of a bridge circuit. When used with a d-c source, either a galvanometer or vacuum-tube voltmeter can indicate the null point. A basic circuit for this type of bridge is shown in Fig. 12-1. In such a bridge, when the voltage across the detector terminals is zero, the relationship $R_a/R_b = R_s/R_x$ will exist within the circuit. R_a and R_b act in the capacity of ratio arms, with the actual bridge balancing for comparison of resistance being accomplished by varying the value of R_s. Adaptations of the basic circuit of Fig. 12-1 can be used not only at d-c but at audio frequencies (usually 1000 cycles) and even at radio frequencies as high as 100 Mc.

[1] Consisting of a series arrangement of battery, resistor, and d-c meter movement.

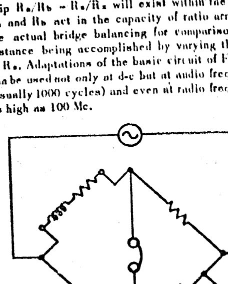


Fig. 12-2 Maxwell Bridge

Among the more common low frequency bridges may be listed the *Maxwell Bridge*, a circuit diagram of which is shown in Fig. 12-2. An important advantage of this particular bridge is the fact that the standard of comparison is a capacitor which, of course, has no stray fields associated with it, thus preventing coupling problems from developing in the measurement of unshielded coils. For components having moderate values of Q (less than 10), the Maxwell Bridge will be found generally satisfactory.

Differing from the Maxwell Bridge in that the standard capacitor is part of a series circuit rather than of a parallel circuit, the *Hay Bridge* (Fig. 12-3) will be found far more satisfactory for Q's with values of 10 or more. The use of a capacitor as the standard again helps to eliminate problems originating in stray fields.

The instrument most commonly used for inductance measurements on coils is probably the *Inductance Bridge*. This is of the general form of the Wheatstone Bridge as can be seen from the basic circuit shown in Fig. 12-4, and because of the fact that a coil is used as the standard, this bridge is particularly well adapted to inductance matching. The fact that this type of bridge is most often used at a frequency of 1000 cycles makes shielding something of a problem, particularly since

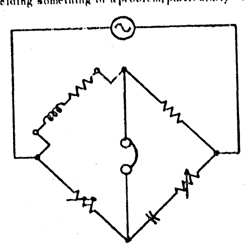


Fig. 12-3 Hay Bridge

the standard has a very definite stray field which, at this low frequency, is difficult to shield effectively.

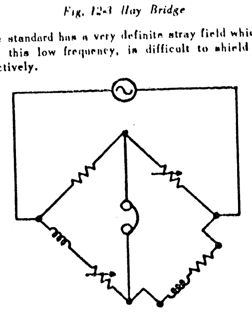


Fig. 12-4 Inductance Bridge

It must be recognized at this point that one of the major problems connected with the use of low frequency bridges is to be found in the successful grounding and shielding with respect to the bridge of the coil under test. This problem has its origin in the well-known *skin effect* which causes high frequency currents to flow on or near the surface of a conductor. It is therefore apparent that at high frequencies shielding will be somewhat easier to accomplish and will permit the use of thinner shield

materials than will be the case at lower frequencies.

A sample of a null-type network which is not based on the Wheatstone Bridge is the *Twin-T*, the basic circuit for which is shown in Fig. 12-5. Beyond the fact that this instrument can be used over a wide frequency range, (500 kc to 40 Mc in one commercial version) there are certain other advantages including the fact that no isolation transformer is required, that the circuit arrangement itself tends to minimize the effects of troublesome residual capacitances, and that there is used a variable capacitance rather than inductance as the balancing component – a practice which is much more successful at high frequencies.

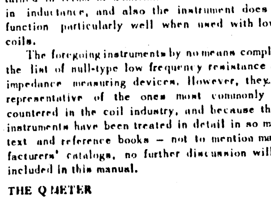


Fig. 12-5 Twin-T basic circuit.

Two possible disadvantages to the Twin-T should be listed. For one thing, readings are obtained in terms of admittance rather than directly in inductance, and also the instrument does not function particularly well when used with low-Q coils.

The foregoing instruments by no means complete the list of null-type low frequency resistance and impedance measuring devices. However, they are representative of the ones most commonly encountered in the coil industry, and because these instruments have been treated in detail in so many text and reference books – not to mention manufacturers' catalogs, no further discussion will be included in this manual.

THE Q METER

If a vote were to be taken among coil engineers for the purpose of selecting the most useful instrument for the determination of coil parameters at radio frequencies, it is certain that the *Q Meter* would win by a substantial margin. Developed about 1934, this versatile instrument which uses the resonance principle to indicate directly on a meter the Q of the coil under test, at the same time shows the frequency of resonance and the value of the capacitance required to produce it.

The basic circuit of the Q Meter appears as Fig. 12-6 and shows the instrument to include an oscillator which generates a small voltage across a small resistance — usually either 0.04 or 0.02 ohms. The coil under test is connected in series with the resistor, while the tuning capacitor and its vernier are in parallel with both the fixed resistor and the coil. Also in parallel with the capacitor is a specially calibrated, vacuum-tube voltmeter.

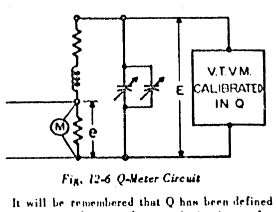


Fig. 12-6 Q-Meter Circuit

It will be remembered that Q has been defined as the figure of merit of a tuned circuit or of a coil or capacitor. It is equally true that Q may be described as a measure of the ability of a coil to store up energy and then to slowly dissipate this stored energy.

The Q Meter depends for its operation upon the fact that at resonance the ratio E/e is equal to the Q of the circuit. It is, therefore, necessary only to set the value of e to a convenient level (as measured by the insertion voltage meter shown in Fig. 12-6), whereupon Q may be read directly from the vacuum-tube voltmeter which is actually measuring the stepped-up voltage resulting from resonance of the tuned circuit.

There are many advantages to this method of measurement. Not only is Q read directly, but the frequency of resonance and the capacitance necessary to tune the coil are both indicated on calibrated dials, thus making it possible to calculate the inductance of the coil.

As useful and as widely accepted as the Q Meter has become, it is not without its limitations. Most of these stem from certain assumptions upon which are based the whole theory of the Q Meter. For example, both the internal resistance of the inserted voltage and the input resistance of the vacuum tube voltmeter are assumed to be a part of the resonant circuit. Because of the high quality of components used in the Q Meter, this assumption to the effect that the entire circuit loss may be found in the inductive portion of the tuned circuit is actually not too serious. It does, however, introduce a slight error, particularly at the higher frequencies.

Another limitation is based on the assumption that the coil which is made a part of the Q Meter circuit has no distributed capacitance. The importance of this assumption is apparent when one considers that only under such an ideal condition could the inserted voltage, e, be in series with the resonant circuit; and since all coils have some distributed capacitance, it follows that the true Q of a coil is actually higher than that which is indicated by a Q Meter reading. Since, however, the tuning capacitance is actually many times larger than the distributed capacitance, the difference between the $Q_{effective}$ (as read on the Q Meter) and Q_{true} will rarely exceed 10 per cent. To convert $Q_{effective}$ to Q_{true}, the formula

$$Q_{true} = Q_{effective} \times \left(1 + \frac{C_d}{C}\right)$$

may be used if C_d is taken as the distributed capacitance of the coil and C as the reading of the Q Meter capacitor dial.

Another source of comparatively small errors in Q Meter readings results from the harmonic content of the built-in oscillator. This effect is most noticeable when readings are being taken on two overlapping frequency ranges because the harmonic content of the oscillator is usually higher at the high frequency end of each band. Since the insertion voltage reads essentially the rms value of the oscillator output, the higher the harmonic content is, the higher the meter will read, resulting in a false Q indication because of too low an insertion voltage. This accounts, at least in part, for the variation in output levels as the Q Meter oscillator is tuned through a band and also indicates why Q's taken in the lower portion of the band are more accurate than those taken near the upper limits of the various frequency ranges.

Working with unshielded coils may introduce certain measurement problems. This is especially true in the case of large coils — particularly loop antennas — where stray fields or pick-up from nearby sources of r-f energy may produce false readings.

Under such circumstances it is advisable that all readings be taken in a well-shielded room. If this precaution is not possible or is considered unfeasible, it is often advantageous to take a series of readings with the coil in a different position each time, and from the effect of these different orientations, to arrive at an average which will largely eliminate the stray field effect.

At frequencies above 50 Mc, it becomes increasingly difficult to distinguish between the coil and the self-inductance of the tuning capacitor and associated wiring of the Q Meter. It is for this reason that a special model of the Q Meter has been developed for use at very high frequencies in which are incorporated circuit and layout changes tending to improve performance at frequencies up to 260 Mc.

OTHER METHODS OF DETERMINING Q

As an engineer engaged in the design and development of high frequency coils and transformers is, of necessity, interested in absolute as well as in comparative measurements. Since absolute measurements at high frequencies involve many problems, it follows that every important parameter should, wherever possible, be determined by at least two completely independent methods.

Before the introduction of the Q Meter it was customary to use either the *frequency-variation method* or the *reactance-variation method* when it was necessary to measure Q.

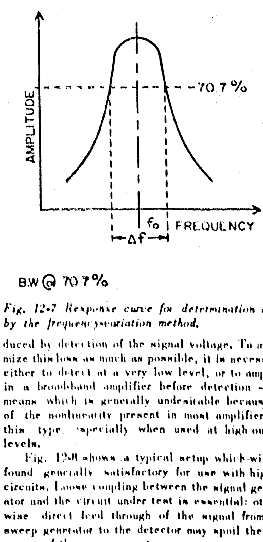
The first of these methods, known as the *frequency-variation method*, is based upon the fact that the bandwidth of a tuned circuit at 70.7 per cent of its response at resonance when divided into the resonant frequency is equal to Q. Expressed as an equation this becomes

$$Q = f_o/\Delta f\ (70.7\%)$$

and is based on the type of curve shown in Fig. 12-7.

The most common means of carrying out the frequency-variation method for the determination of Q involves the use of a sweep generator with an oscilloscope on which to view the selectivity curve of the tuned circuit. This method suffers somewhat in accuracy because of the difficulties surrounding a true linear detection of the sweep wave, thus making accurate calibration of the oscilloscope of vital importance since it offers the only practical means of overcoming the nonlinearity introduced by the detector.

A second problem associated with this system of measurement is the change in the Q of the tuned circuit which results from the loss of energy introduced by detection of the signal voltage. To minimize this loss as much as possible, it is necessary either to detect at a very low level, or to amplify in a broadband amplifier before detection — a means which is generally undesirable because of of the nonlinearity present in most amplifiers of this type, especially when used at high output levels.



B.W @ 70.7%

Fig. 12-7 Response curve for determination of Q by the frequency-variation method.

Fig. 12-8 shows a typical setup which will be found generally satisfactory for use with high Q circuits. Loose coupling between the signal generator and the circuit under test is essential: otherwise direct feed through of the signal from the sweep generator to the detector may spoil the accuracy of the measurements.

Fig. 12-9 shows a similar setup where inductive coupling rather than electrostatic (capacitance) coupling is used between the sweep generator and the tuned circuit. For lower Q circuits, this system usually will give more dependable results than the setup outlined in Fig. 12-8.

A convenient check of the validity of measurements made by either of the above setups consists of detuning the coils without touching the generator or the detector. If, under these circumstances, the response falls off substantially to zero, the degree of coupling is such as to prevent direct feed through, and readings may be assumed to be correct.

Fig. 12-8 Typical setup for determination of Q by the frequency-variation method. (High Q circuits.)

A somewhat similar method of determining Q which is slightly more time consuming but very much more accurate is that known as the *point-by-point method.* The required equipment for this system includes a standard signal generator and some sort of output indicator which most commonly is either a detector-amplifier or a vacuum-tube voltmeter. The point-by-point method consists of adjusting a tuned circuit to resonate at a desired frequency, and then of maintaining a constant output level while varying both the frequency of the signal generator and the input level (output of the signal generator). From information thus obtained, a response curve can be plotted or the Q calculated from the response at resonance and the 70.7 per cent points. A caution to be observed at all times in connection with the point-by-point system of measurement is the avoidance of tight couplings between the signal generator and the tuned circuit. For best results, coupling should not exceed the minimum required for transfer of a sufficient signal to carry out the desired measurements.

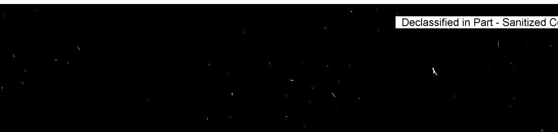


Fig. 12-9 Typical setup for determination of Q by the frequency-variation method. (Low Q circuits.)

For use at higher frequencies, a somewhat more fundamental method of measuring Q — *the decremental method*[1] — is sometimes employed. In the years prior to the invention of the Q meter — in fact even before the introduction of the term Q — the decremental system was used as a means of checking circuit efficiency. Based on a power concept, this system, as the name suggests, actually measures the amount of time required for the current in a shock-excited, tuned circuit to die down to $1/e$ of its original value. In Fig. 12-10 appears a basic layout for this system of measurement which will be seen to consist of a pulse-modulated signal generator loosely coupled to the tuned circuit which in turn is loosely coupled to a detector whose output is shown on an oscilloscope.

It is this general type of measuring system — one involving the use of decrement — that forms the basis of the "echo boxes" which were used as radar calibration aids during World War II. In this application, high-Q cavity resonators were excited by the carrier pulse from a radar transmitter and then were allowed to reradiate back into the radar receiver in a manner closely resembling reradiation from a target.

The great danger in using the decremental method in the laboratory lies in the use of too tight couplings between the generator and the tuned circuit and between the two circuits and detector. A practical check which may be used to determine optimum coupling value is to lessen the coupling value by degrees until that point is reached where two successive readings produce the same results. At this point it may be assumed that the loading is such as to reproduce no error, and it may be considered that the percent of energy loss per cycle is equal to 1/Q.

[1] Included in this discussion because of its historical value. A present-day coil engineer will find few practical applications for this method.

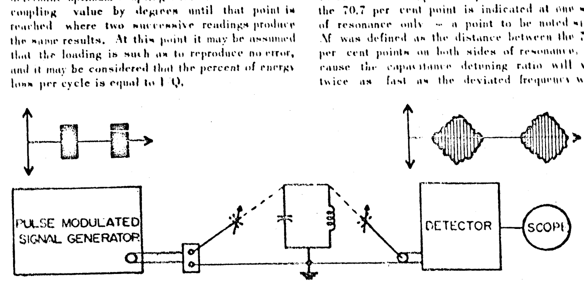


Fig. 12-10 Decremental system for determination of Q.

A second caution which might be listed concerns the need for a pulse length sufficient to allow for the formation of a definite and recognizable plateau from which to measure the decline of the circuit energy. The shape of the detected pulse is largely determined by the circuit Q, with those circuits which have the highest Q's requiring the longest time intervals to reach their peaks.

The decremental system is not recommended for use with unshielded coils since the normal radiation loss present in such coils is usually sufficient to make the results of this system of measurement of questionable value. Considered as a basic system of measurement, there is little to recommend the decremental system. It is, however, of sufficient historical value to be worthy of inclusion in this discussion, particularly as a means of confirming Q measurements.

Another method for determining Q is that which involves *reactance-variations.* This is essentially the same process as the frequency-variation method except that the tuned circuit is detuned by a small vernier capacitor connected in parallel with the tuning capacitor. By means of this vernier, the circuit may be detuned to one 70.7 per cent response point, whereupon Q may be defined as $Q = C/\Delta C$. The justification for this equation stems from the fact that Q is equal to $f_0/\Delta f(70.7\%)$. From the familiar relationship $2\pi f = 1/\sqrt{LC}$, it can therefore be said that Q is equal to $C/\Delta C$. Fig. 12-11 shows a typical response curve in which the 70.7 per cent point is indicated at one side of resonance only — a point to be noted since Δf was defined as the distance between the 70.7 per cent points on both sides of resonance. Because the capacitance detuning ratio will vary twice as fast as the deviated frequency when using the reactance-variation method of measuring, it is necessary merely to detune sufficiently to reduce the maximum voltage at resonance to 70.7 per cent of its value.

However true the foregoing statements may be, it is believed that when using the reactance-variation method it will be found far more satisfactory to include both 70.7 per cent points as was done in the frequency-variation method. This recommendation is based upon the established fact that the 70.7 per cent points can be located with a higher degree of accuracy than is the case with the resonant frequency. It is therefore apparent that the overall accuracy of the measurement will be increased by the use of two 70.7 per cent points rather than one. Under this system, ΔC assumes a value twice as great as when only one 70.7 per cent point is used, thus making it necessary to double the value of the capacitance required to produce resonance. In other words, while it is perfectly true that $Q = C/\Delta C$ as was first stated, it will be found possible to increase the accuracy of the method by using the formula $Q = 2C/\Delta C$ where ΔC represents the range of capacitance included between the two 70.7 per cent points rather than between resonance (f_0) and one half power point as was originally suggested.

It is often convenient to use this reactance-variation method of measurement as a crosscheck on the performance of the Q Meter since the necessary vernier and calibrated capacitance are a part of the conventional Q Meter circuit.

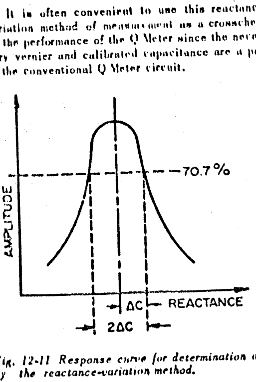


Fig. 12-11 Response curve for determination of Q by the reactance-variation method.

GRID-DIP METER

An instrument which is of considerable value in coil work, especially for the determination of resonant frequency, is the *grid-dip meter*. This device was originally developed by radio amateurs

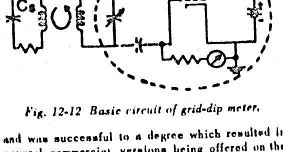


Fig. 12-12 Basic circuit of grid-dip meter.

and was successful to a degree which resulted in several commercial versions being offered on the market. A typical instrument of this type[4] covers, with the aid of plug-in coils, a tuning range of from 2.2 Mc to 400 Mc.

[4]The Megacycle Meter manufactured by the Measurements Corporation, Boonton, New Jersey.

Grid-dip meters are generally of the basic type represented by the simplified circuit diagram of Fig. 12-12. The principle of operation depends upon the fact that a tuned circuit, when coupled to the coil of a grid-dip meter, will absorb energy from the oscillator when both circuits are tuned to the same frequency. This loss of energy from the circuit of the grid-dip oscillator causes a lessening of feedback with a resultant decrease or "dip" in grid current. The sharpness of this dip is a function of the Q of the external circuit, with those of high Q's causing a sharp dip in grid current as measured by the milliammeter.

In these instruments only the coil is exposed, with the balance of the circuit including the calibrated capacitor enclosed within a case which serves as a shield. Used with reasonable care, grid-dip meters will measure frequencies to an accuracy of 2 per cent or better. Satisfactory operation of a grid-dip meter is largely dependent upon the degree of inductive coupling present between the tuned circuit and the exposed coil of the meter. If the coupling is too tight, two frequencies will be noted as the results of over-coupling. The optimum coupling varies, of course, with the LC ratio and the Q of the tuned circuit. Fortunately, it is a simple matter to learn to use these instruments, and only a little experience is necessary to produce dependable readings. When it is desired to measure the resonant frequency of shielded coils, a short twisted coupling loop can readily be made up and inserted inside the shield can near the low-potential point of the coil to provide coupling between the magnetic field of the coil under test and the external coil of the grid-dip meter. Fig. 12-13 shows a typical setup of this type. The coupling must be adjusted by trial and error to prevent "jumping" or "snapping" of the oscillator which will be noted if the coupling is too tight.

It is apparent that inductance can be measured by means of a grid-dip meter if the value of C_s (Fig. 12-12) is known. Preferably, this capacitor should be small in size and of a low-loss type such as silvered mica or ceramic. Instruction books accompanying most meters of this type give detailed instructions for measuring inductance values by merely attaching standard capacitors to the clips provided with the meters for this express purpose.

Another use for meters of the grid-dip type is in point-by-point frequency-variation measurements of Q. In many instances, the calibrated dial will not be sufficiently finely divided to provide the desired accuracy, thus making it necessary to

resort to beat-frequency techniques. It is this particular measurement technique that is most often used to evaluate relatively high Q coils such as trap circuits while they are a part of the complete circuit and are in position in the end equipment.

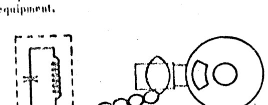


Fig. 12-13 Method of coupling grid dip meter to shielded coil.

TESTING R-F TRANSFORMERS

As used in this discussion, the term *r-f transformer* applies to that type of high frequency coupling device whose broad band characteristics make it suitable for use in the "front end" of a receiver. When such transformers are tested, the features of greatest interest are invariably the selectivity curve and the stage gain.

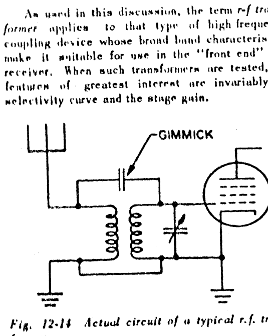


Fig. 12-14 Actual circuit of a typical r.f. transformer.

A typical circuit for such a transformer appears in Fig. 12-14. Since it follows conventional design practice, this unit will be seen to consist of an untuned primary and a tuned secondary with a coupling capacitor commonly called a *gimmick* connected across the high ends of the primary and secondary windings. The reason for the use of the gimmick is probably best shown by Fig. 12-16, 12-17 and 12-18 which represent typical r-f transformer response curves when primary to secondary coupling is entirely capacitative (Fig. 12-16), entirely mutual inductance (Fig. 12-17), or a combination of mutual inductance, and capacitive coupling – the latter as the result of a gimmick capacitor (Fig. 12-18). The obvious improvement in uniformity of gain over the band which results

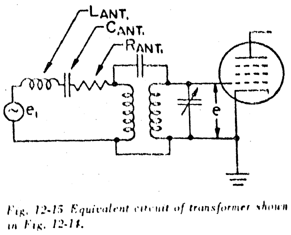


Fig. 12-15 Equivalent circuit of transformer shown in Fig. 12-14.

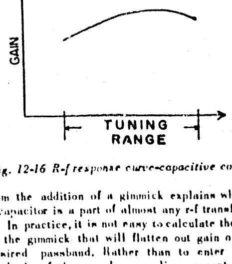


Fig. 12-16 R-f response curve-capacitive coupling.

from the addition of a gimmick explains why such a capacitor is a part of almost any r-f transformer.

In practice, it is not easy to calculate the value of the gimmick that will flatten out gain over the desired passband. Rather than to enter into an analysis of the complex coupling present in any such transformer, it will be found far simpler in most cases to determine the gimmick value experimentally, while measuring overall performance of the coil in a circuit similar to that shown in Fig. 12-20. Since the value of the capacitance will undoubtedly be small – something in the order of 0.5 to 5.0 mmf in most instances – a convenient means of finding the proper value may involve the use of short lengths of parallel line (two conductors imbedded in a plastic insulation –

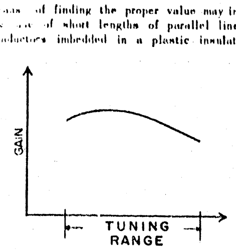


Fig. 12-17 R-f response curve-mutual inductance coupling.

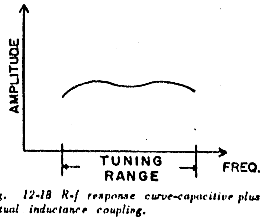


Fig. 12-18 R-f response curve-capacitive plus mutual inductance coupling.

usually polyethylene – as shown in Fig. 12-19) one end of which is soldered to each winding of the transformer, after which pieces are snipped from the other end until the desired coil performance is recorded. At this point, the gimmick may become a permanent part of the transformer or it may be removed and its capacitance measured after which a new capacitor of equal value can be added in its place.

Two points of interest relative to gimmicks might be worthy of mention at this point. For a comparatively accurate evaluation of the capacitance represented by a gimmick, it is necessary only to measure the capacitance of a rather long piece of parallel line – say twelve or more inches. Since the characteristics of this type of line will be relatively uniform, it then becomes possible to estimate with a high degree of accuracy the capacitance value of short lengths of gimmick, provided, however, that the conditions of measure-

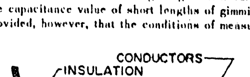


Fig. 12-19 Parallel line used for gimmick capacitors.

ment were held constant. Actually, in view of the difficulties encountered in the measurement of small values of capacitance – as for example, those of less than one mmf – it will frequently be found entirely satisfactory as well as much more convenient to determine the value of gimmick capacitance in terms of gimmick length.

The second point concerns the effect that the physical location of a gimmick may have upon the performance of a transformer. It must be remembered that the gimmick does not represent the total capacitive coupling present in the unit, since a certain amount of this coupling is the natural result of proximity of coils and leads. It therefore follows that the position of the gimmick with respect to either or both of these parts of the transformer must inevitably influence the primary to secondary capacitive coupling. So great may this effect be, that a relatively common procedure in the coil industry calls for the adjustment of gain and bandwidth by varying the physical location of the gimmick. Obviously, this very fact points out the need for keeping the gimmick in a position consistent with stability under shock and vibration. It is also apparent that performance may be affected if the gimmick is removed and replaced by a small, fixed capacitor.

R-f transformers are usually measured in a circuit similar to that shown in Fig. 12-20. If the circuit is broken at the grid of the r-f tube – the point indicated in the drawing by X – the signal generator may then be connected behind the transformer, thus eliminating it from the circuit, and a record made of the amount of signal (generator output voltage) required to produce a convenient reading on the vacuum tube voltmeter. This reading is usually referred to as the *standard output*, and the method by which it is determined as the *calibration of the output meter*. Once the output meter has been calibrated, the connections are

returned to those shown in Fig. 12-20, and the attenuator on the signal generator reduced (assuming that the design is such as to produce gain) until the VTVM indicates standard output. The ratio of the two attenuator settings then represents the stage gain of the r-f (antenna) transformer alone.

intermediate frequency amplifier, the image frequency would then be 1200 kc plus 455 kc plus 455 kc or 2110 kc. To determine image rejection under these conditions would simply become a matter of finding the generator output necessary to produce standard output at 1200 kc and again at 2110 kc, with the ratio between the two values

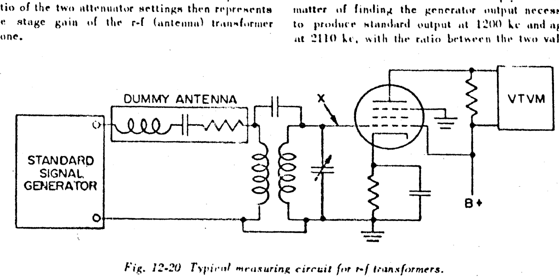


Fig. 12-20 Typical measuring circuit for r-f transformers.

Of importance in such a setup is the value of the dummy antenna. Naturally, the closer this value is to the actual value of the antenna that will be used, the more accurate the readings will be. In actual practice, the dummy antenna most often used has the form and the value shown in Fig. 12-21 – a combination which experience has shown to be representative of average antenna parameters.

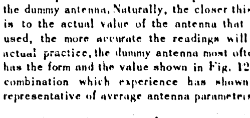


Fig. 12-21 Standard dummy antenna.

Besides the determination of the actual passband and the gain at various points within this band, *image rejection* is an r-f transformer characteristic which is usually of concern to the engineer. In a conventional receiver which has the local oscillator operating above carrier frequency, the image frequency will be that of the carrier plus twice the i-f frequency. For example, when a broadcast receiver is tuned to receive a carrier of 1200 kc and assuming a conventional 455 kc

being the image rejection of the particular transformer being tested. While this is a comparatively simple measurement to make, it may sometimes be found that a simple VTVM will not have the necessary sensitivity to work with the limited output of an average signal generator. In such instances, a more sensitive detector such as a crystal in combination with a d-c amplifier (chopper) may be used as may also a form of tuned detector which must, of course, be calibrated at the image frequency.

All performance characteristics of an r-f transformer as, for example, i-f rejection, harmonic responses, second images, or other spurious responses, may be checked by the same general measurement techniques.

It should probably be mentioned in passing that because of the Miller Effect, the input reactance of the first r-f tube will vary with the plate load impedance in both phase and magnitude. As is pointed out elsewhere in this manual, this condition is not particularly serious in well-shielded amplifiers using tubes of the pentode type. However, it is well to remember that in cascode or grounded-grid amplifiers conditions may develop such that individual stage gains cannot be measured directly. Under such circumstances, the best procedure seems to be to adjust the coils individually after they are in place in the equipment, tuning the

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individual circuits with a grid-dip meter and measuring the gain indirectly, much as conversion gain is measured when grid mixing is employed.

I-F TRANSFORMER MEASUREMENTS

As was said to be the case with r-f transformers, the characteristics of primary concern in the performance of i-f transformers are the shape and the size of the response curve – in other words, the selectivity – and the gain of the particular transformer.

Unlike r-f transformers, intermediate frequency transformers are designed for the express purpose of passing only a specific and usually comparatively narrow band of frequencies. The actual shape of i-f response curves will be found to vary greatly with some showing the high gain, sharply selective characteristics represented by Fig. 12-22, while others are wide, double peaked, and relatively low in gain as shown in Fig. 12-23.

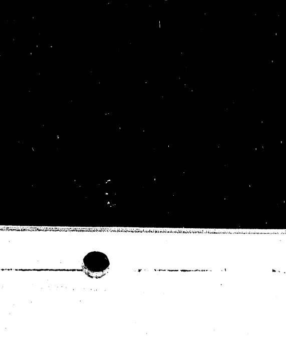

Fig. 12-22 Response curve of high-gain, narrow-band, r-f transformer.

To describe adequately the performance of an i-f transformer, it is necessary to know the bandwidth at several points off resonance. These points are usually expressed as so many *db down* or so many *times down*. The expression 2 times down – written 2X – means the point at which the output voltage is one half that at resonance, while 10 times down, 10X, is the point where the output voltage had dropped to one tenth that at resonance. As will be pointed out later, the conventional method of determining these points calls for maintenance of a uniform output from the generator. Therefore in practice, the 2X points are those at which generator settings of a magnitude twice that at resonance will produce standard output, with the 10X, 100X, and 1000X points requiring outputs of ten, one hundred, and one thousand times, respectively.

Fig. 12-23 Response curve of broad-band, double-peaked, moderate gain, r-f transformer.

In single stage measurements, it will usually be found sufficient to take readings at 2X, 10X, and 100X, while in complete amplifier measurements it is customary to include 1000X readings to provide a better idea of adjacent channel attenuation.

The circuit shown in Fig. 12-24 is representative of those used for measuring i-f stage gain and, by the point-by-point method, the response as well. Such arrangements of components as these are known in the coil industry as *test jigs* and represent by far the most common method of checking the performance characteristics of high frequency transformers in coil development laboratories.

To be completely effective, such test jigs should be constructed and operated with due consideration for at least the following:

1. The signal generator should be terminated in accordance with the recommendations of the manufacturer.
2. In so far as is possible, the circuit should be representative of that used in the end equipment for which the transformers are designed. Of particular importance are lead dress, chassis layout, shielding between circuit elements, tube type and connections, avc bias, and tube shields – all of which should resemble the production model.
3. *Bogie tubes* – those with center values – should be procured from the manufacturer after due consultation as to their ultimate application.

4. Leads from both signal generator and VTVM should be as short as possible, and the use of a probe-type voltmeter is recommended.
5. The power supply should furnish a well regulated and adjustable voltage and should be provided with meters permitting a constant check on its performance.

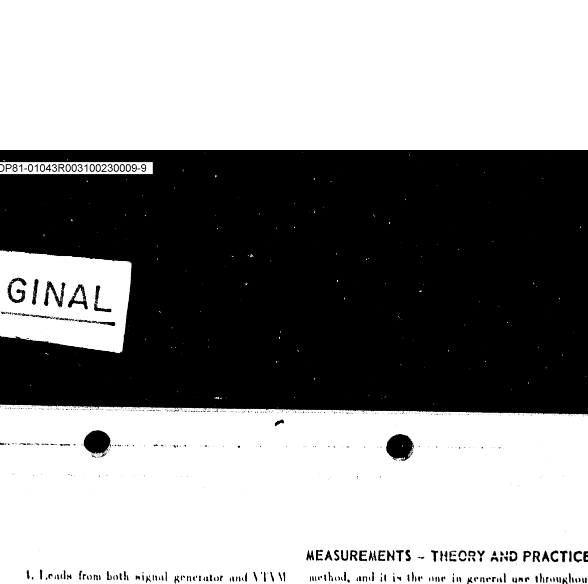

Fig. 12-24 Representative test circuit for use with r-f transformers.

Test jigs of this general type are calibrated by connecting the signal generator to the grid of the second tube at the point marked X and, of course, to ground. The attenuator is then adjusted until the vacuum tube voltmeter shows a convenient output indication which then becomes the standard output. *Stage gain*, which it should be noted includes the gain of the first tube, is then measured by moving the generator connection to the grid of the first tube and then reducing the output voltage by means of the attenuator until the VTVM indicates the standard output value, whereupon the ratio of the two attenuator settings will represent the stage gain of the transformer.

Selectivity can be determined by the identical setup, with the procedure for locating the 2X points being first to double the generator output at resonance and then to swing the generator to those points above and below resonance where standard output is obtained. In a similar manner, any desired point may be obtained and a response curve plotted from the readings.

From the description of this procedure, it is apparent that plotting a selectivity curve in this manner is a somewhat slow and laborious process. As a result, many engineers now use sweep generators in conjunction with oscilloscopes on which the complete response curves appear directly. This system is very much faster than the point-by-point method, and it is the one in general use throughout the coil industry for production testing. As a laboratory or design method, it is open to some criticism, particularly in the case of extremely narrow band transformers or high Q traps, since unless the sweep rate is extremely low, there will be insufficient time for the response to build up, and the indicated curve may not represent the true performance of the unit under test. While unquestionably a valuable and time saving method of viewing transformer performance – especially in a comparative manner – it is recommended that all high Q sharply selective components be checked during the actual design process by the slower point-by-point method.

MEASUREMENTS AS A BASIS OF TRANSFORMER SPECIFICATIONS

It was stated in the opening paragraph of this section that electrical measurements of high frequency transformers not only serve to define performance but also provide a means of specifying fundamental design parameters. Up to this point in our discussion of measurements only performance has been considered.

Since it is frequently necessary to so define a transformer that it may be readily duplicated, it is apparent that merely to record its performance will not provide the information needed to produce an identical unit.

Proper specification of transformers is one of the major problems facing the coil industry today. There is a serious lack of knowledge as to which electrical parameters are significant and which ones may be neglected in describing and specifying a transformer. Because of this situation, it is

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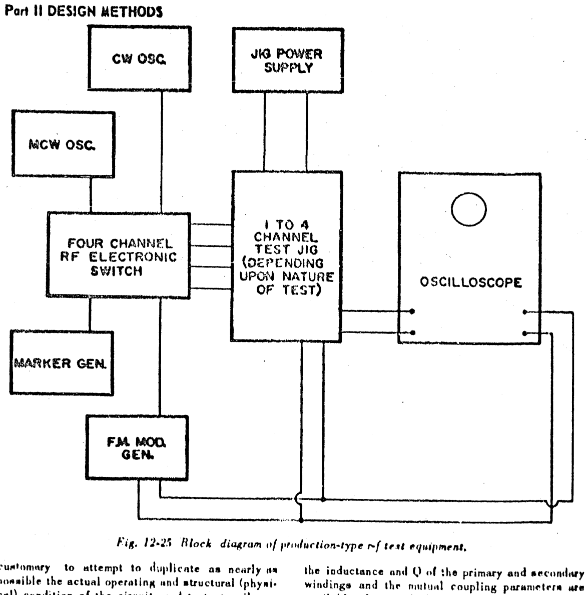

Fig. 12-25 Block diagram of production-type r-f test equipment.

customary to attempt to duplicate as nearly as possible the actual operating and structural (physical) condition of the circuit, and to test coil performance in a test jig so constructed. So basic is this matter of transformer parameters that it has been made the subject of a Signal Corps sponsored Research and Development Project which is currently in progress.[8]

Considered in the light of accepted practices within the industry, there seems to be general recognition of the impossibility – or at least of the improbability – of successfully duplicating a high frequency transformer by reference only to d-c measurements. In other words, the mere fact that the inductance and Q of the primary and secondary windings and the mutual coupling parameters are available along with the physical measurements of the unit and its component parts is not to be taken as assurance that a unit with equal values of the above parameters will give equal performance. If all measurements were made on the same instrument by the same operator using the same techniques, duplication would most likely be successful; but if, for example, the inductance and mutual were measured on two different bridges by two different operators, it is more than likely that the resulting coils would not be identical. Furthermore, experience of the coil industry accumulated over a period of years indicates that even in those cases where all measurements are apparently duplicated, there still exists a firm basis for reasonable doubt as to the accuracy of the readings.

So serious is this problem, that little effort is expended by the coil industry on absolute measurements except for basic design purposes. Duplication of coils is almost universally accomplished through the use of so-called *standards.* The idea behind the use of standards is simply that through their use a way is opened to measure coils comparatively rather than absolutely, thus making exact duplication much easier.

In practice, the system works in this general manner. A transformer is selected from among a number of units as being representative of the average performance desired. This particular coil is arbitrarily designated as the *standard,* whereupon it is sealed to prevent tampering, and all parameters are carefully measured and recorded along with the serial numbers of the equipments used and the name of the operator making the measurements.

The standard having been approved, the next step is to prepare a number of transformers which match as nearly as possible the performance and the various measured parameters of the accepted standard. These transformers too are sealed, identified, measured, and all values recorded. The designation given to this group of coils varies within the industry with some companies referring to them as *masters* and others as *working standards.* In any case, the treatment accorded the two types of coils differs sharply. The *standard* is stored in a safe place from which it is taken only in the event of serious question concerning values or performance of that particular coil, or for a periodic check of the working standards to insure that they have not changed in value through use, neglect, or willful action on the part of production personnel. The *working standards* form the basis for the actual comparative testing that controls the quality of transformers in production. To insure uniform quality, it is customary to check the working standard against the standard at regular intervals.

It is apparent that this system is one of comparison in which absolute values are of little concern. The use of standards represents a system which is not particularly satisfactory but which does possess the obvious advantage of minimizing the need for exact calibration and/or correlation of measuring equipment. Since, in practice, both customer and supplier have a standard, and since these coils were developed and checked by one operator on one set of equipment, successful duplication becomes a matter of matching the performance of the standard rather than attempting to duplicate a series of absolute values.

If this exchange of standards is accompanied by information concerning the circuit in which the coil was developed, comparative testing becomes entirely feasible and production may be held within satisfactory limits. In the event of disagreement, the standards are always available to both supplier and customer and offer a means whereby differences may be adjusted.

For those cases where certain absolute measurements are necessary – usually in the course of basic design – certain precautions are essential. It was suggested earlier in this discussion that whenever possible, every parameter should be determined by two different methods of measurement. In this manner it is possible to maintain a check on the general accuracy of the readings since any considerable error obtained by one system would be evidenced by wide variations in the value of the same parameter as determined by the optional method.

A general check on the accuracy of equipment can be maintained by periodic checking against a standard coil. Such coils may be purchased or they may be made up by the user for a specific purpose. In every case they should be of sturdy construction, exhibiting a maximum of stability and should, if possible, be shielded to avoid interference due to stray field phenomena. Once designated as standards, these coils should be used only for equipment checks and at all other times should receive the care usually accorded to standards of any kind.

Accurate measuring of transformers in a test jig by the use of a standard signal generator and a vacuum-tube voltmeter requires certain precautions. For one thing, at the beginning of the test the output of the signal generator should be connected directly to the VTVM as a mutual check of generator output and the meter calibration. It is more than likely that a majority of cases will show some disagreement between the reading of the vacuum-tube voltmeter and the indicated output of the signal generator. Fortunately, this is not too serious, and if the disagreement does not exceed 10 per cent, it will be possible to correct for the discrepancy and to record sufficiently accurate stage gain measurements.

Once the vacuum-tube voltmeter has been calibrated, the attenuator of the signal generator

[8] Contract No. DA 36-039 sc-64522 entitled "Characterization of I-F Transformers".

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should be checked at all settings. If the generator has a piston-type attenuator, it will most probably be found to be highly accurate. If, however, the signal generator is an older model using a ladder-type attenuator, the check may be expected to indicate some discrepancies between steps and also between the top and the bottom of the slide wire. Should the errors be large, the generator should be returned to its manufacturer for repairs and adjustments.

When making stage gain measurements, it is desirable to avoid the use of modulation. By so doing, full advantage is taken of the accepted fact that the output accuracy of a signal generator is best when the r-f carrier is unmodulated. Care should be taken to avoid overloading a transformer under test through the use of too high output from the generator. As far as possible, it is always well to keep the applied voltage near the level at which the transformer will work in the end equipment.

There has been little in this discussion about the problems of drift in measuring equipment because the topic is a general one in all electronic devices. In vacuum-tube voltmeters drift can be a serious matter especially if readings are being taken in the vicinity of 0.1 volt. A good rule to follow when using a VTVM is to allow the instrument to operate for some time with the input shorted on its most sensitive range — the setting in which drift will be most apparent.

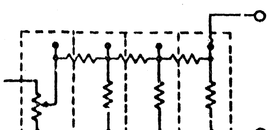

Fig. 12-26 Ladder-type attenuator.

It probably should be recognized at this point that while it was advocated earlier in this discussion that stage gain measurements be taken with a constant output and a variable input, there are engineers who prefer to work with a constant input and a variable output. Either system can be used, but it is generally accepted that the attenuator of an average signal generator is more accurate than is the average voltmeter scale. By holding output constant, in a stage gain measurement the VTVM actually serves more as a galvanometer than as a voltmeter, and a higher degree of accuracy is therefore maintained in the readings.

In all instances, setups should be planned to utilize the shortest possible leads. This is particularly true at high frequencies and is illustrated by the fact that leads with a length of only 2 inches can, when inserted between a signal generator and a conventional high frequency, probe-type, vacuum-tube voltmeter, introduce errors of up to 20 per cent in measurements made at 100 Mc. A simple test to determine whether lead length is introducing a measurement error consists merely of making slight changes in the length of the connections and noting the difference, if any, resulting from the new leads. If no change in readings is apparent, it may be assumed that the leads are not introducing an appreciable error into the measurements.

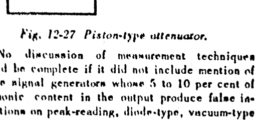

Fig. 12-27 Piston-type attenuator.

No discussion of measurement techniques would be complete if it did not include mention of those signal generators whose 5 to 10 per cent of harmonic content in the output produce false indications on peak-reading, diode-type, vacuum-type voltmeters. Thermocouples, bolometers, square-law operated crystals, and other types of true rms output meters are less affected by high harmonic output and are recommended for use with carriers high in harmonic content. In extreme cases, as for example, when measuring a high-pass filter stage, it may be found advisable to use a tuned output meter which might take the form of a stable receiver capable of detecting only the fundamental of the carrier, thus effectively minimizing the influence of the harmonics present in the carrier wave.

MEASUREMENT OF IMPEDANCE

It is often desirable for the coil designer to quickly determine the impedance (dynamic resistance, R_d) of parallel resonant circuits without dis-

turbance to the circuit assembly. This can be accomplished by the use of the Boonton Q-Meter and a suitable coil (hereinafter called the *R_d Coil*), having a minimum Q of about 200 at the measurement frequency, if used on the 250 Q-scale of a Model 260A Q-meter or 50 if used on the 60 Q-scale. The system has some limitations imposed by the accuracy of the Q-Meter and the skill of the operator in making this particular type of measurement, but many find it useful and it is presented here in deference to those engineers who subscribe to this method.

The *R_d Coil* should preferably be shielded to avoid stray effects from surrounding objects, of sufficient inductance to tune to the measurement frequency with the Q-Meter capacitor and have an R_d when tuned to resonance of the order of the impedance to be measured.* Thus, when the R_d of the impedance under measurement equals the R_d of the *R_d coil*, the Q-Meter deflection will be exactly halved, the prerequisite for maximum precision of measurement. For higher values of impedance the 250 Q-scale should be used and for lower values of impedance it is desirable to use the 60 Q-scale.

Measurement Procedure

The Q-Meter set-up for the R_d measurement of a typical 455 kc double-tuned transformer is shown in Fig. 12-28. The high and low potential terminals of the *R_d coil* are connected to the high and low potential coil terminals (A and B respectively) of Q-Meter. All low potential terminals of the transformer, including the shield, are connected to the ground terminal C of the Q-Meter as shown. The dynamic resistance of the *R_d coil* is first determined as follows:

Connect the high potential terminals of the secondary and primary (grid and plate terminals 1 and 3, respectively, to ground thereby shorting both transformer windings. Now adjust the frequency dial to the desired frequency (in this case 455 kc) and resonate the *R_d coil* with the Q-Meter capacitor and record the Q and resonating capacity; designating them as Q_1 and C_1 respectively. In a specific example of record, these values were found to be: $Q_1 = 200$ $C_1 = 155$ $\mu\mu f$.

Then,

$$R_d = \frac{Q_1}{2\pi f C_1}$$

$$= \frac{200}{6.28(.455)10^6(455)10^{-12}} = 154{,}100 \text{ ohms.}$$

The dynamic resistance (R_{d_s}) of the secondary is next determined, leaving the capacitor setting and frequency dial untouched. Terminal 1 of the secondary is connected to the high potential Q-Meter capacitor terminal D, meanwhile maintaining the primary shorted (terminal 3 to ground). Now tune the secondary of the transformer for maximum deflection of the Q-Meter, and record the reading as Q_2. In this example Q_2 was measured as 160. Q_1 (*R_d coil* alone), of course, remains 200.

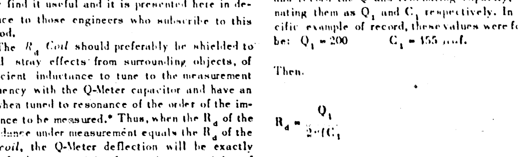

Fig. 12-28 Top view of Boonton Q-Meter Terminals showing coil connections for the determination of R_d.

* See Section 14 for a detailed analysis of the impedance of a parallel resonant circuit; this impedance at resonance is shown to be equal to the parallel loss resistance (R_d) of the coil.

Then: $R_{d_s} = R_d \left(\frac{Q_2}{Q_1 - Q_2}\right)$ where R_d is the dynamic resistance of the auxiliary coil *(R_d coil)*.

$$R_{d_s} = .1541(10^6)\left(\frac{160}{200-160}\right) = 154{,}100(4)$$

$$= 616{,}400 \text{ ohms.}$$

The dynamic resistance of the primary R_{d_p} is next measured in identically the same manner with the Q-Meter capacitor setting and frequency dial still untouched. The secondary is shorted (terminal 1 is connected to ground) and the primary (terminal 3) is connected to terminal D of the Q-Meter. The primary is adjusted for maximum deflection of the Q-Meter and the Q reading recorded as Q_3. In this example Q_3 measured 147.

Then $R_{d_p} = .154..10^6)\left(\frac{147}{200-147}\right) = 151{,}100\ (2.77)$

$$= 426{,}800 \text{ ohms.}$$

The effective primary impedance with the secondary coupled $R_{d_{(p+s)}}$ is next measured, Q-Meter settings still being the same. Terminal 1 is disconnected from the Q-Meter ground, thereby removing the secondary short circuit and the secondary circuit adjusted for *minimum* Q-Meter deflection, Q_4. In this case Q_4 was measured as 113.

Then $R_{d_{(p+s)}} = .1541(10^6)\left(\frac{113}{200-113}\right) = 154{,}100(1.3)$

$$= 200{,}300 \text{ ohms.}$$

The $R_{d_{(p+s)}}$ is used to determine the ratio of the actual coefficient of coupling to the coefficient at critical coupling according to the following equation:

$$\tau^2 = \left(\frac{R_{d_p}}{R_{d_{(p+s)}}}\right) - 1 \qquad \text{where } \tau = \frac{\kappa_{actual}}{\kappa_{critical}}$$

$$\tau = \sqrt{\left(\frac{R_{d_p}}{R_{d_{(p+s)}}}\right) - 1} = \sqrt{\left(\frac{426{,}800}{200{,}300}\right) - 1} = 1.07$$

This transformer is overcoupled by 7% due to the fact that it has been measured unloaded. Rechecking with the primary and secondary shunted by resistors simulating the output and input resistance of a converter and i-f tube, respectively, (950,000 ohms and 4 megohms) we find the following:

$$R_{d_s} = .1541(10^6)\left(\frac{155}{200-155}\right) = 154{,}100(3.44)$$

$$= 530{,}100 \text{ ohms.}$$

$$R_{d_p} = .1541(10^6)\left(\frac{131}{200-131}\right) = 154{,}100(1.9)$$

$$= 292{,}800 \text{ ohms.}$$

$$R_{d_{(p+s)}} = .1541(10^6)\left(\frac{104}{200-104}\right) = 154{,}100(1.08)$$

$$= 166{,}500 \text{ ohms.}$$

$$\tau = \sqrt{\left(\frac{R_{d_p}}{R_{d_{(p+s)}}}\right) - 1} = \sqrt{\left(\frac{292{,}800}{166{,}500}\right) - 1} = .875$$

This shows the transformer to be 12.5% undercoupled when measured with a load equivalent to the loading supplied by the tubes under operating conditions.

CONCLUSION

An understanding of the various equipments and techniques used in the measurement of high frequency transformers is essential to a well-informed coil engineer. Toward this end, it is urged that catalogs of the various test equipment manufacturers and the instruction books that accompany their instruments be studied carefully.

Electrical measurement is a large subject which has been dealt with at length in many standard text books. It has been possible here to present only the bare fundamentals of the most common instruments and techniques. Many common methods – for example, those involving beat-frequency

techniques – have been omitted in the interest of brevity.

It is hoped that this discussion – will have served to point out to the reader something of the importance of measurements in the design and specification of high frequency transformers. If it has contributed to an understanding of the problems surrounding specification and duplication of these units, it will have served its purpose.

BIBLIOGRAPHY

Beatty, R.T.
Radio Data Charts, Second Edition
Iliffe, London

Brookes-Smith, C.
"Measurement of Small Values of Inductance and Effective Resistance"
Electrical Communication, Vol. 13, July 1934

Dwight, H.B.
Electrical Coils and Conductors, Their Electrical Characteristics and Theory
McGraw-Hill Book Company, Inc., New York, 1945

Hartshorn, L.
Radio Frequency Measurements by Bridge and Resonance Methods
John Wiley & Sons, Inc., New York, 1940

Henney, Keith
Radio Engineering Handbook, Fourth Edition
McGraw-Hill Book Company, Inc., New York, 1950

Hund, A.
High Frequency Measurements
McGraw-Hill Book Company, Inc., New York, 1939

Langford-Smith, F.
Radiotron Designer's Handbook
Wireless Press, Sydney, 1952

Laws, F.A.
Electrical Measurements, Second Edition
McGraw-Hill Book Company, Inc., New York, 1938

Terman, F.E.
Radio Engineers Handbook
McGraw-Hill Book Company, Inc., New York, 1943

Terman, F.E.
Measurements in Radio Engineering
McGraw-Hill Book Company, Inc., New York, 1935

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Section 13

TECHNIQUES OF FABRICATION

GENERAL

Many precautionary measures and short-cuts are used in every industry. They may vary from novel shop practices which are not written into manufacturing specifications to precautionary measures that all well known by those who are experienced in the art.

SIMPLIFIED METHODS OF MAKING EXPERIMENTAL PARTS

Plastic materials, high frequency magnetic cores and certain other parts are inexpensively produced in large quantities by high speed production methods employing expensive tooling. The production of sample quantities of coils and transformers requiring these parts and materials, for which there are no production tools, presents a major procurement problem. Such parts and materials can, of course, be machined individually in a model shop. This is costly and time consuming. Following are suggestions for making sample lot quantities of often required parts.

A Method of Making Molds for Cast Plastic Parts:

In the development of a new transformer design, it is often necessary to have small quantities of non-standard plastic parts. As was pointed out in Section 6, plastic molds are expensive items, and when once completed, do not readily permit the introduction of changes in the molded parts.

Until very recently, this fact placed a severe restriction upon the design engineer who wished to try out new and different methods of assembling electronic parts made from plastics. The only possible solution has been to machine the desired shapes from larger pieces of the desired plastic material. Obviously an expensive means to an end. In the case of pilot runs to test out a design, it has proved a most impractical system.

Today, however, a simple means of making small quantities of even relatively complex plastic parts is available at low cost to even the smallest laboratories. This system combines the use of casting resins with plastisol molds.

Plastisols are paste-like dispersions of polyvinyl materials in a liquid plasticizer. They are easily compounded to produce any desired degree of rigidity following fusion at 350F. Available at moderate cost from a number of sources, these materials can be used to make simple molds capable of producing up to fifty or more plastic parts before reaching the end of their useful lives. Conventional steel molds for producing typical i-f and r-f transformer plastic parts frequently cost hundreds of dollars, whereas plastisol molds to produce cast-plastic parts of the same shape can be made for a few pennies.

A major advantage of the process lies in the fact that no elaborate equipment is required and that neither special skills nor any knowledge of chemistry is needed to make good plastisol molds. An oven capable of sustained heat at 350F, a reasonably accurate balance or scale for weighing out the materials, and suitable mixing apparatus comprise the necessary equipment.

The basic mixture consists of

100 parts resin (Geon Paste Resin No. 121)*
70 parts plasticizer (Plasticizer G.P. No. 261)*

To this is added 2 parts of stabilizer (tribasic lead phosphate)** for every 100 parts of resin.

The recommended mixing procedure calls first for weighing out the desired amount of plasticizer. To this is added the resin, and the mixture is

* B. F. Goodrich Company

** National Lead Co., Brooklyn, N. Y.

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POOR ORIGINAL

stirred thoroughly, after which the stabilizer is added. Thorough mixing is necessary for the success of the process. Unless stirred to the point where the mixture is absolutely uniform and free from lumps, the finished molds will have tiny holes throughout the plastisol. The use of a device such as a Waring Blendor provides a most satisfactory means of insuring thorough mixing, but where such a device is not available, satisfactory results can be obtained by the use of a mortar and pestle. Mixing by simple stirring is not impossible but will be found to consume a great deal of time since the resin "wets" slowly in the plasticizer.

The degree of flexibility present in the final product may be controlled by varying the amount of plasticizer in the mixture. The proportion given above represents a good average and is the point at which experimentation should start. If it appears that a more rubbery material would be desirable, the amount of plasticizer should be increased; if a stiffer material is sought, the proportion of resin should be increased. In practice, it will be found that the contours of the piece being molded will in large measure determine the type of plastisol to be used. For example, the presence of undercuts on a piece creates a requirement for a mold which will stretch easily, thus releasing the piece without damaging the mold.

If stored in temperatures in the order of 40F, plastisol mixtures may be kept for weeks without danger of spoilage. This means that a quart or so can be made up at a time, thus serving the needs of an average laboratory for several weeks.

To make a mold, it is only necessary to make from metal or wood one piece of the exact size and shape desired. This piece is then placed on the bottom of a metal container (such as a can in which typewriter ribbons are shipped) and covered with the plastisol mixture to a depth of from an eighth to a quarter of an inch. The container is then placed in a 350F oven for a period of time sufficient to bring the mass up to the temperature of the oven. The actual curing process is one of fusion, which means that timing is not important so long as the material becomes heated through. For the average small mold 15 to 25 minutes will be sufficient after which the mass can be removed from the oven and allowed to cool.

At this point, it will be found that some shrinkage has taken place and the mold may be readily removed from the metal holder. The piece can then be removed by flexing the mold which is of a rubbery nature. This should be done carefully, of course, to avoid tearing or damaging the mold, and the mold is then ready to receive its first "shot" of casting resin.

As simple as this process is, it is capable of a high degree of accuracy, and is one that can be used with pieces of almost any shape, including those with negative draft or with undercuts. It offers the coil engineer an inexpensive and fast means of molding small lots of plastic parts without resorting to the use of conventional metal molds and requiring no more equipment than is to be found in the average coil laboratory.

Special Shapes of Powdered Iron Cores:

The development of new coil types, especially those involving miniaturization, often requires core configurations that are not available from existing tools. It is, of course, impractical to have tools prepared to produce, by conventional production methods, the few cores that are necessary to prove out an idea.

Sample quantities can be machined without extensive equipment and within reasonably short time if proper preparations are made and followed. Chapter 3 on Magnetic Materials gave suggestions for machining powdered iron cores and suggested several types of grinding wheels known to give satisfactory results.

The following suggestions are offered:

1. Do not design or make samples of a part which is not practical for mass production by pressing or other established production techniques.*
2. Obtain from a core supplier solid blanks of the required material which are as near to the desired size, at least in diameter, as is possible. Most core manufacturers have standard tools for producing cylindrical cores in a wide range of sizes. Check their catalogues. A blank that is longer than required will allow material for chucking in a lathe or other machine tool.
3. Inside diameters or cavities should be first machined in a lathe with a tool post grinder, operated at high speed, using small mounted grinding wheels.
4. Outside grinding operations including threading should follow the inside machining. The cutoff operation should be performed last, using a small thin grinding or cutting wheel.
5. If several identical cores are to be made it may be advisable to perform the same operation on each of the samples rather than attempt to complete all operations on a core before starting on the next one.
6. If facilities are available, cooling the parts during grinding with an oil such as D.A. Stuart Oil Co., Superkool 201M, is recommended. This not only prolongs the life of the grinding wheels and helps to produce a smoother finish but also will prevent the fine iron particles removed from catching fire.

* See paragraph on "Practical Shapes" on page 8 of Section 3.

Special Shapes of Ferrite Cores:

Ferrite cores, like powdered iron cores, are often required for developmental purposes in shapes that are not commercially tooled. Unfortunately, they are not easily produced by machining operations (see Section 3, *Machining Ferrites*).

Due to their extreme hardness they are difficult to grind and even diamond grinding wheels do not prove entirely satisfactory. Cavities or holes are even more difficult to machine than outside dimensions.

It is possible to perform machining operations on unfired or partially fired ferrite material but complete shrinkage data and proper facilities for the high temperature firing must be available. This generally requires that the machined pieces be returned to the ferrite manufacturer for final firing, which is very unsatisfactory from the standpoint of time and cost.

It is recommended that in cases where special ferrites are required the designer work with a reliable ferrite manufacturer who is equipped to completely fabricate whatever is necessary. The design of the desired parts should be carefully discussed so that development is not centered around a part which is impossible to produce by production methods.

WINDING SUGGESTIONS

There are a number of well established shop practices that are very helpful to the engineer who must produce his own developmental samples. It is suggested that Section 10, Windings, be reviewed as additional material to the following.

Measurement of Coil Spacing:

The physical act of accurately measuring the spacing between two coils of either solenoid or universal type is not a simple problem. Several methods of measurement are commonly employed and specifications should reflect the instrument used and the points between which the measurements are to be taken.

Among the instruments for the measurement of coil spacing, listed in ascending order of accuracy, are the scale, vernier calipers, and the optical micrometer. While good steel scales can be used with a relatively high degree of accuracy by a skilled operator, readings taken in this manner are subject to error in duplication and are not sufficiently accurate for close tolerance transformers. Vernier calipers represent a compromise between the low cost and low accuracy of the scale and the high cost and very much higher accuracy of the optical micrometer. With the "inside-outside" type of calipers readings may be taken either between the inside surfaces of the two coils or between the outside faces. Reasonable care on the part of the operator should remove the possibility of damage to the windings, and experience will soon dictate the amount of pressure which can be safely applied when the reading is taken.

For accuracy, no method of measurement of coil spacing equals that afforded by the optical micrometer. These instruments consist merely of a low power microscope, equipped with cross hairs, which moves laterally on a bed driven by a micrometer screw. Not only does this instrument provide a highly accurate means of measurement, but it offers the further advantage of operating without the necessity of physical contact with the object being measured, thus removing all possible danger of damage to the windings.

Another use for which an optical micrometer will be found of value in a coil laboratory is for measuring turns in a solenoid coil made from small wire. By starting with the cross hairs just at the right of the first turn and then moving from left to right across the winding, an accurate count can be made — again without danger of damage to delicate wires.

The exact point at which measurements of coil spacings should be taken for specification purposes is a subject on which there is no universal agreement within the coil industry. From a theoretical standpoint it would appear that spacings of universal windings should be specified from center to center. In practice, this is difficult and impractical because of the problem attached to location of the exact center of a pi winding. Far more workable are the systems calling for measurements between the inside faces of the two coils, or from the inside face of one coil to the outside face of the other coil. It will be noted that the latter method is effectively the same as measuring center to center, particularly if performed in both direc- tions and the results averaged. Scale measure- ments made in this manner will be found to corre- spond closely to vernier caliper measurements made either between inside or outside faces.

When in search of accurate measurements from which to specify design procedures, it is well to measure by at least two methods and to average the results, since only in this manner can the en- gineer eliminate to the maximum extent the pos- sibility of errors in his measurements.

Solenoid windings present a somewhat different problem in measurement. In practice, few solenoid windings will be found in which the leads leave the coil forms at exactly the same angle. Since varia- tions in this angle will be reflected in variations in in spacing if measured at different points, the problem becomes one of choosing an identifiable point at which to take all readings.

One system which has given good results calls for the measurements to be taken between the out- side of the first turn of the primary and the start of the second turn of the secondary. With either an optical micrometer or scale, this restriction on the location at which the readings must be taken poses no problem, and it does have the very positive ad- vantage of defining the method and location, thus increasing the possibility of duplication of readings by different operators.

It should be stressed that in the matter of coil measurements it is of the utmost importance not only to make all measurements in the same manner with the same sort of measuring devices but also to specify clearly the exact manner in which the measurements are to be taken. Only when such procedures are followed, can there be reasonable hope that duplication of transformers will be suc- cessful.

Control of Inductance:

Precise control of inductance can be a problem of considerable magnitude in the production of close tolerance windings. Because such variables as wire diameter, coil form diameter, wire tension and number of turns may, and do, have an effect upon the inductance of both solenoid and universal windings, some simple method for adjusting the final inductance value is necessary if excessive rejects are to be avoided.

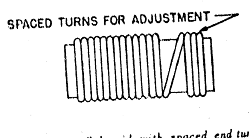

Fig. 13-1 Solenoid with spaced end turns for inductance adjustment.

In the case of solenoid windings, this adjust- ment can be accomplished by the relatively simple expedient of winding a large majority of the turns in normal fashion and then when near the end, space the final turns of the coil in a way permitting phys- ical adjustment (Fig. 13-1) (positioning) of these turns to add or subtract inductance. This system is commonly used in the coil industry as a means of matching inductances and also for tracking oscilla- tors in radio receivers.

The actual adjustment of the end turns is usu- ally done with the coil in an oscillating circuit. Obviously, varnish-treated coils cannot be ad- justed; if such treatment is specified, it must be applied following test and adjustment. Some quick- setting adhesive must be applied at the time the leads are positioned to insure that the coils retain their setting.

Universal coils which fall outside of tolerance or which must be adjusted precisely present a somewhat more difficult problem. The very nature of universal coils eliminates the method of adjust- ment recommended for solenoid windings. High inductance can readily be corrected by the removal of turns — providing the windings have not been varnish-treated or otherwise encased in a manner preventing unwinding. It should be kept in mind that universal coils, because of their large number of turns, are less critical with respect to variations in total turns than are solenoids. In other words, in a small winding, of say seven turns, a variation of one-half turn may result in an inductance value far outside of an acceptable tolerance, whereas in high inductance universal windings, a variation of 10 to 15 turns may still produce an acceptable unit.

While obviously, the exact degree of variation from the specified number of turns that can be tol- erated in a winding is a matter to be decided in each individual case, a good rule of thumb to guide coil design is to hold small solenoid windings to plus or minus one quarter turn and universals to

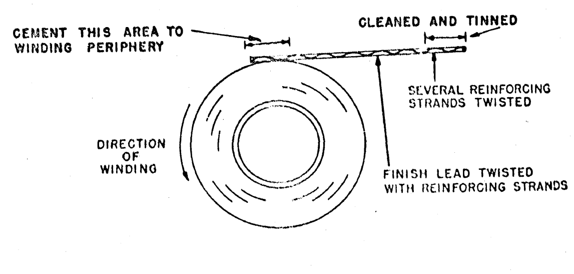

Fig. 13-2 Method of Lead Reinforcement.

plus or minus 2 turns in small inductance values and up to perhaps as much as plus or minus 1% of the total number of turns in the case of large coils. Since modern winding machines are equipped with automatic stops, accurate to within one turn, wide tolerances, even if permissible, need not be ac- cepted in most cases.

Because the average universal coil is somewhat springy in nature, it is possible to introduce a certain amount of inductance variation by squeezing the winding. The result of this squeezing or de- forming of the winding is a displacement of turns with a consequent change in the flux linkage within the coil. The extent to which a winding may thus be adjusted depends upon its shape and size. Since this procedure may also affect the Q, it follows that inductance control by deformation of the winding is an individual process, the exact details of which must be determined by each case. Squeezing the winding to a smaller diameter de- creases the inductance while increasing the dia- meter by squeezing parallel to the coil axis, in- creases the inductance. Many coil companies have adopted the use of wooden pinchers by which coils may be squeezed either parallel to the axis or at right angles to the axis, thus introducing a sub- stantial amount of the desired type of correction into the individual windings. Such corrective action must be used with caution to prevent induct- ance drift which will result if the turns of the *squeezed windings* shift position.

Reinforcing of Coil Leads:

When it is known that the leads of a coil wound from magnet wire of size No. 36 or smaller are to be subjected to considerable amounts of handling, it will often be found advantageous to reinforce both the start and finish leads to lessen the danger of breakage.

Nearly every coil manufacturer has his par- ticular method of accomplishing this end. A gen- erally satisfactory system will be found to consist of the addition to the coil leads of a loop of several turns of wire having a size somewhere in the order of No. 36, and the same type insulation as is used in the winding. This wire may be wound in the form of a hank using pins for spacers. Once re- moved from the pins, the wires are twisted a few times, placed along side the lead to be strength- ened, and then fastened in place.

In the case of a start lead, this latter operation may consist simply of cementing the reinforced lead to its proper place on the coil form and then proceeding with the winding operation. For the finish lead, the plan shown in Fig. 13-2 may be used. Here, it will be seen, the reinforcing wires are cemented to the outer turns of the coil and aligned in the same direction as the coil is wound Since the coil lead becomes a part of the reinforced lead during the tinning process, the result is the formation of a small cable which is firmly attached to the coil and therefore serves as a support for the

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very delicate coil leads.

The use of reinforced leads is particularly recommended for those cases where large inductances are wound with extremely small wires. It must be remembered that a start lead, broken off close to the coil, represents an item which can be neither be used nor repaired. Finished leads can often be repaired by carefully removing a turn or two from the coil and then recementing the point of termination. Obviously, this method is possible only in windings having a large number of turns where the loss of two or even more turns will have a negligible effect upon the overall characteristics of the coil.

As a matter of practical interest, in the case of very small wires (No. 39 and smaller), no attempt should be made to loosen the end turns without first applying a solvent to soften the cement. Needless to say, the solvent used for this purpose should be compatible both with the wire insulation and the treatment material.

Handling Litz Wire:

The manufacture of high Q coils often necessitates the use of stranded conductors commonly known as Litz wire. As shown by the curves of Fig. 1-4 in Section 1, the effectiveness (Q) of Litz wire coils drops off rapidly as strands are broken.

It is necessary to contact all strands when making connection to the winding. When solderable* wire is used in the make-up of the Litz it is not difficult to contact all strands since the entire group can be wrapped around the point of connection and soldered together without first removing the insulation from each individual strand.

When enamel or Formex wire is used, removal of the insulation is required prior to soldering. It is advisable to exercise extreme care during removal and then to twist the strands together and tin the twisted group so that individual strands will not become separated or broken when connection is made to lugs or terminals.

Windings should be checked for broken strands after soldered connections are made. This can be performed as a simple resistance test allowing suitable tolerances to cover the number of broken strands which are permitted. It is common commercial practice to allow one broken strand in wires of 3, 5 or 6 strands and as many as two in Litz wires having 7 to 10 strands and possibly 3 or more in wires having 12 or more strands. The number of strands that can be permitted should be governed by whatever decrease in Q can be tolerated below that obtained from an average coil having all strands.

* Wire whose insulation coating is readily removed by the heat of a soldering operation. See Section 1, page 3.

Stripping Formex (Formvar) Wire:

There are two common methods for the removal of Formex (Formvar) insulation, mechanical and chemical. Each have their undesirable and desirable features and brief descriptions of each are included.

Mechanical:

For laboratory purposes where relatively few wires are involved, the simplest method is to carefully remove the insulation with a small piece of sandpaper held between the tips of the thumb and index finger. If care is exercised, even Litz wire may be stripped by this method without breaking strands and no elaborate or expensive equipment is required.

For mass production operations, mechanical stripping is best accomplished by rotating wire or fibre-glass brushes. Machines available commercially* for performing this operation consist of a suitable chassis for bench mounting, having two small circular wire or fibre-glass brushes revolving in opposite directions. The end of the wire to be stripped is inserted between the revolving brushes and the insulation is quickly removed. This equipment can also be used to remove fabric insulation, as well as Formex or enamel.

Care must be exercised to see that the brushes are kept in proper adjustment, to avoid severe abrasion or breakage of strands during the stripping operation. One advantage of mechanical stripping is the absence of chemical elements (see following discussion of chemical stripping) which may later cause corrosion of small wires resulting in reduced performance or even failure of the winding.

Chemical:

A solution of 85% (C.P. Grade) Formic Acid is maintained in a suitable container such as a beaker or bottle and a thin layer of Mineral Oil (C.P. Grade, chemically neutral, sulfur free) is floated on the surface.

The wire to be stripped is immersed in the stripping solution to the depth of the required strip and allowed to remain until the insulation has softened. This may require between 30 and 90 seconds, depending upon the size of the wire and the thickness of the insulation. Immediately after removal from the solution, the wire is cleaned of its insulation by wiping with a clean, dry cloth, such as clinical gauze or equivalent. It is important that the stripping solution be kept from unwanted portions of windings, either by splashing, operator's hands, or wiping cloths. The stripper is irritating to the skin and contact should be avoided. If it comes in contact with the skin, it should be immediately removed with soap and water. Food should not be brought into the work area.

* Rush Wire Stripper Division, The Eraser Co., Inc., Syracuse, N.Y., Model D-1.

The chemical method does not cause breakage of small strands, due to reduction of diameter, as is often the case with mechanical stripping. There are some who feel that it is impossible to entirely remove the chemicals subsequent to stripping and the use of this method, therefore, represents a potential source of trouble thereafter.

There are a number of methods for chemically removing Formex (Formvar) in addition to the heretofore described.*

EXPERIMENTAL DETERMINATION OF COUPLING

One of the very necessary operations in transformer design is that by which the proper degree of coupling between primary and secondary windings is determined. Many attempts have been made to reduce this determination to one which can be solved mathematically in terms of a physical arrangement of two coils. Unfortunately, no simple procedure has yet been evolved, and the combination of magnetic and capacitive coupling that is the natural result of coil separation, coil size, lead placement, and other factors of equal import has successfully resisted all but the empirical approach.

It therefore has become standard practice in the coil industry to finalize any design by empirical determination of coil spacing. This is, of course, an operation which must be performed with care — particularly with respect to uniformity of lead placement in successive tests and also in regard to the handling of the coils so as not to damage them or change their characteristics.

Probably the easiest and most common method for producing variable spacing transformers involves locating one of the windings on a very thin coil form whose ID is such as to just fit over the regular form. Commonly called a "slider", this collar allows relatively free movement of the coil along the regular form, thus making it possible to locate the point of desired coupling with a high degree of accuracy. To accomplish this a single layer of No. 40 wire is close wound directly over the coil form approximately where the desired winding will eventually be located. This single-layer winding should be slightly longer than the length of the desired winding and one end should be accessable. A layer of thin paper is placed over the winding and the desired coil wound over the paper, after which the single-layer winding is pulled out from under the paper, leaving the desired winding slightly loose on the coil form. This enables free movement for purposes of adjusting coupling, without danger of damage to the winding.

In using a variable coupling set-up of this general type, it is advisable to begin the series of measurements with the coils widely spaced. This normally should represent an under-coupled condition, the presence of which would be substantiated if the second reading taken with the coils moved closer together were to indicate higher gain and wider bandwidth. Successive adjustments of spacing accompanied by gain and bandwidth measurements will establish the proper spacing for whatever degree of coupling is desired. After the correct spacing is empirically established, it is recommended that several transformers be wound in accordance with this established spacing, so that the correctness of the dimension can be verified.

GENERAL PRECAUTIONS FOR HANDLING VARIOUS MATERIALS

Coils and transformers are complex components made up of many materials and the failure of any one may result in the failure of the inductive component. Many materials are entirely satisfactory when operated or used in a suitable environment, or when properly handled, but may be unsatisfactory when improperly used or handled. The following suggestions are given for guidance.

Iron Cores:

Some types of iron cores are attacked by solvents or hot wax and may disintegrate under such conditions. If windings which are to be impregnated in hot wax or coil lacquer use iron cores, it is advisable to make sure that the core material is not affected by the impregnating material.

* For other methods, reference is made to, "Methods of Removing the Insulating Film from Formex Wire" by E. J. Flynn and G. W. Young, General Electric Review, June 1946.

Handling of Mica:

Mica is often used as the dielectric for capacitors either as silvered mica or between metallic plates. In order that the Q of such capacitors remain as high as possible under conditions of high humidity, it is important that the mica be kept clean and free from oil and especially from finger prints and perspiration. If the mica does accidentally pick up dirt or other foreign material, it may be removed by a suitable wash with alcohol.

Section 14
THEORY AND DESIGN

INTRODUCTION

It is essential in the rational practice of design of r-f coils and transformers, that the coil engineer possess a full familiarity with the basic tools of such design — namely, a thorough knowledge of the inter-relationships of such coil and circuit parameters as Q, bandwidth, inductance, capacitance, source and load impedances, etc. It is the purpose of this Section to develop a simple and applicable set of such basic relationships which should become the everyday tools of the coil designer. Obviously, any degree of academic sophistication can be introduced in such a development, ranging from an elemental presentation of only the simplified formulas and step-by-step examples of their use, to the precise mathematical development of the exact general equations which have questionable utility in the practical area of design with which we are concerned here. The approach to be presented will attempt to straddle the advantages of both treatments, with references to selected bibliography for more complete analysis of the points under discussion.

It must be recognized too, that it is impossible to disassociate practical coil design from any consideration of the circuit in which the coil is to operate. Stray capacitances in amplifier grid and plate circuits, for example, must be taken into account in establishing the value of the resonating capacitors in the tuned circuits used to couple tubes; similarly, coil Q can be appreciably lowered by the loading effects of plate and grid circuits in some applications, and deviations from optimum design of coil impedances due to neglected circuit parameters can have serious effects on the amplifier gain and bandwidth of the r-f and i-f systems in which the coils are to be used. The coil designer, therefore, must also have a working knowledge of the elements of r-f amplifier design and the relationships of these elements to coil characteristics. In like fashion, the r-f circuit designer must have some knowledge of the behavior, limitations, capabilities, and variations possible in his use of coils as coupling media in his circuits. Accordingly, this Section will also present a review of the r-f class A amplifier with emphasis on those parameters of amplifier design which come within the coil designer's area of interest. A number of examples will be used to illustrate how these circuit parameters are treated in the evolution of practical coil and transformer designs.

Effort has also been made to reduce the more frequently used design operations to simple nomographs and tables, and examples of use of such aids are included.

Finally, to stimulate further interest in the art and science of r-f coil design on the part of the engineer and to point out the more powerful tools that are available to him as he matures in the field, an introduction to network theory is presented, and simplified approaches to circuit analysis and coil design are discussed. Examples of transformer designs using equivalent lattice networks are used to illustrate the advantages of these advanced techniques.

It will be assumed in the following pages that the electrical engineer has a basic knowledge of a-c theory, involving the use and manipulation of such quantities as capacitance, inductance, resistance, reactance, frequency, angular velocity, and the operator "j" in the mathematical expressions. These terms are well established but a brief redefinition and discussion, where applicable, are included in this Section to clarify the more specific nature of the term in the application then under discussion.

The finalized formulas developed in the text are included with the related illustrations for the purpose of rapid reference in subsequent design use of the manual.

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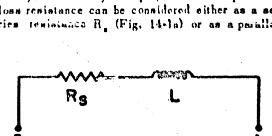
THE SIMPLE COIL

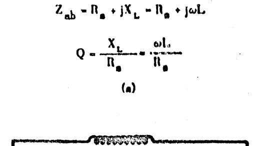
In its most elemental form as illustrated in Fig. 14-1, a physical coil designed for r-f applications may be considered as possessing inductance (designated by L) with a corresponding reactance (designated as X_L) as its prime attribute, and an unavoidable amount of equivalent resistance (designated by R). This equivalent resistance represents the total electrical losses inherent in the coil due to several factors including resistance of the wire conductor itself, losses in the magnetic core that may be used with the coiled conductor, and the aggregate of losses in the adjacent mounting hardware, supporting chassis, and the shield, if any. For analytical purposes this equivalent loss resistance can be considered either as a series resistance R_s (Fig. 14-1a) or as a parallel



$$Z_{ab} = R_s + jX_L = R_s + j\omega L$$

$$Q = \frac{X_L}{R_s} = \frac{\omega L}{R_s}$$

(a)



$$Z_{ab} = \frac{R_p}{Q^2} + j\omega L$$

$$Q = \frac{R_p}{X_L} = \frac{R_p}{\omega L}$$

(b)

Fig. 14-1 Series and parallel representations of a simple coil.

resistance R_p (Fig. 14-1b); in the series representation, the value of the equivalent loss resistance R_s will be low, while in the parallel representation this loss resistance R_p will be high — but in both cases these resistors, though differing in value represent the identical quantity of electrical loss.

In the series circuit, the total impedance Z_{ab} of the coil is

$$Z_{ab} = R_s + j(\omega L) \qquad (1)^1$$

where: R_s = equivalent series loss resistance
ω = angular velocity = $2\pi f$
ωL = reactance of inductor L = X_L

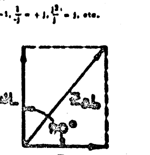
The ratio of reactance to resistance is a measure of the efficiency or quality of the coil and is designated as Q. A perfect coil would have $R_s = 0$ and therefore Q = infinity; in contrast, a perfect resistor R_s would have no inductance, so that $\omega L = 0$ and Q would be zero. The term Q accordingly helps to define the degree of perfection of the coil and is a very useful parameter in coil design. *At radio frequencies, the Q's of coils are almost always greater than 10, a fact which permits considerable simplification of the various working formulas, as will be shown.*

For the series circuit under discussion:

$$Q = \frac{\omega L}{R_s} \qquad (2)$$

[1] The quantity "ωL" is known as inductive reactance and represents the impediment offered to the flow of a-c current by the inductance L. The reactance is preceded by the mathematical operator "j" to indicate that it has a +90° (quadrature) phase relationship with respect to the resistance vector R_s, as indicated below. The impedance Z_{ab} is the vector sum of the two vectors (R_s and ωL), as shown. By giving the operator j a value = $\sqrt{-1}$, it is possible to treat it algebraically, that is:

$$j^2 = -1, \frac{1}{-j} = +j, \frac{j^2}{j} = j, \text{ etc.}$$



Consider now the equivalent parallel circuit representation of Fig. 14-1b. Here, with the two components (resistance and inductance) in parallel, the impedance equals the product of these components divided by their sum, or

$$Z_{ab} = \frac{R_p(j\omega L)}{R_p + j\omega L}$$

Multiplying numerator and denominator by $R_p - j\omega L$ and re-arranging:

$$Z_{ab} = R_p\left(\frac{\omega^2L^2}{R_p^2 + (\omega L)^2}\right) + j\omega L\left(\frac{R_p^2}{R_p^2 + (\omega L)^2}\right) \qquad (3)$$

$$\text{and } Q = \frac{\text{reactance}}{\text{resistance}} = \frac{\omega L\left(\frac{R_p^2}{R_p^2 + \omega^2L^2}\right)}{R_p\left(\frac{\omega^2L^2}{R_p^2 + \omega^2L^2}\right)} = \frac{R_p}{\omega L} \qquad (4)^1$$

Since Q is an inherent property of the coil, it is the same in either the series or parallel representation; then

$$Q = \frac{\omega L}{R_s} = \frac{R_p}{\omega L}$$

From these relations, we have:

$$R_p = Q^2R_s \qquad (5)$$

$$\text{and } R_s = \frac{R_p}{Q^2} \qquad (6)$$

It will be found convenient to remember these two simple relations since conversion of a series representation of losses to a parallel representation (or vice versa) occasionally simplifies an analysis of a coil circuit.

Having established that $Q = R_p/\omega L$ for the parallel circuit equivalent, equation (3) can be re-written

$$Z_{ab} = R_p\left(\frac{1}{Q^2 + 1}\right) + j\omega L\left(\frac{Q^2}{Q^2 + 1}\right)$$

If Q>10, as is usually the case, less than 1% error will be introduced by neglecting the "1" in the denominators, so that for all practical purposes the impedance can be expressed as

$$Z_{ab} \simeq \frac{R_p}{Q^2} + j\omega L. \qquad (7)$$

Contrasting this final parallel circuit relation with the series circuit equation

$$Z_{ab} = R_s + j\omega L. \qquad (1)$$

the equivalence of R_s and R_p/Q^2 can be seen for the resistive portion of the impedance. Note that the reactive term still is ωL.

The relations expressed in equations (2), (4), (5), and (6) are frequently used in coil calculations and should be absorbed with understanding so that they can be used almost axiomatically in subsequent operations. These equations are repeated below for easy reference:

Inductor with series resistor; $Q = \frac{\omega L}{R_s}$ (2)

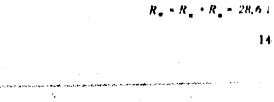
Inductor with parallel resistor; $Q = \frac{R_p}{\omega L}$ (4)

from which $R_p = Q^2R_s$ (5)

$R_s = \frac{R_p}{Q^2}$ (6)

Design Examples:

(a) *A 1 millihenry experimental coil has a Q = 143 at 455 kc. A Q' of only 100 is desired. What value of resistor should be added in series with the coil to depress the Q to the value desired?*



$R_e = R_x + R_s = 28.6\,\Omega$

[1] Note that in the series circuit $Q = \frac{\omega L}{R_s}$, but here in the parallel equivalent $Q = \frac{R_p}{\omega L}$, a difference that should be fully understood.

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Original Coil: $R_s = \frac{\omega L}{Q}$ (2)

$$\text{or } R_s = \frac{(2\pi\, 455\times10^{3})\,(1\times10^{-3})}{143} = \frac{2860}{143} = 20 \text{ ohms}$$

$$\text{Altered Coil: } R_s = \frac{\omega L}{Q'} = \frac{2860}{100} = 28.6 \text{ ohms}$$

$R_x = R_s - R_s = 28.6 - 20 = 8.6$ *ohms should be added in series.*

(b) *What value of resistor should be added in parallel with the coil of (a) above to depress the Q to the value desired?*

Original Coil: $R_p = \omega LQ$ (4)

$$= (2860)\,(143) = 409{,}000 \text{ ohms}$$

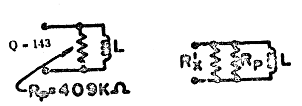
Altered Coil: $R_s' = \omega LQ'$

$$= (2860)\,(100) = 286{,}000 \text{ ohms}$$

Since the resistor R_x' is being added in parallel with the coil resistance R_p, use the formula for two resistors in parallel to find R_x'.

$$R_x' = \frac{R_p\,(R_s')}{R_p - R_s'} = \frac{(409{,}000)\,(286{,}000)}{409{,}000 - 286{,}000}$$

$= 950{,}000$ *ohms should be added in parallel*



$Q' = 100$

$$R_s' = \frac{R_x'(R_p)}{R_x' + R_p} = 286\ \text{K}\Omega$$

The coil selected for the illustration would not be uncommon for 455 kc, and the example points up the following important considerations in the use of such a coil in practical circuits:

(1) It is sometimes difficult and uneconomical to design a coil to specific Q value in production. Accordingly, the coil designer will, in those cases, prepare the coil specifications to give him a value of Q somewhat higher than actually needed in the circuit, and then specify a value of resistor to be placed across the coil when wiring the circuit to give the exact value of Q needed. This practice is particularly common in wide-band circuits where fairly low Q's are needed for the coils (see Fig. 14-1). It is seldom that a series resistor is used for this purpose since, as should be obvious from the example, the value of resistors needed would be very small, uncommon, and too frequently of such low value as to require the use of an expensive wire-wound resistor instead of the much less expensive carbon composition type (the latter are available from 10 ohms to 20 Megohms).

(2) The example also illustrates the sensitivity of coil Q to resistances placed in parallel with the coil. Such resistances could be the plate resistances of tubes, grid resistors, or even ohmic leakages across, say, tube socket terminals which might be effectively in parallel with the coil (such a leakage could be caused by use of acid solder flux, for example). The subsequent discussions on amplifier circuits will bring out the manner of taking into account the tube and circuit parameters so that the "effective Q" of a coil can be calculated when it is in its operating circuit.

The reader should rework the examples, this time calculating the value of R_p using the relation $R_p = Q^2 R_s$ as developed in equation (5). The calculation should be made for the unloaded coil (Q=143) and then for the loaded coil (Q=100).

Rather than using the derived equations for calculating R_s, Q, or R_p precisely, it is sometimes adequate and more convenient to determine approximate values using the reactance chart given in Fig. 14-2. The chart permits the direct evaluation of the Q of a circuit if the reactance and either the parallel or series resistance are known, or the value of the equivalent series or parallel resistance if either the Q or the reactance and the other form of resistance are known. In most cases the values given in the chart can be used directly, but in certain instances it will be necessary to apply proper decade multipliers to the lines.

Example:

Referring to Example "a" (page 3) above

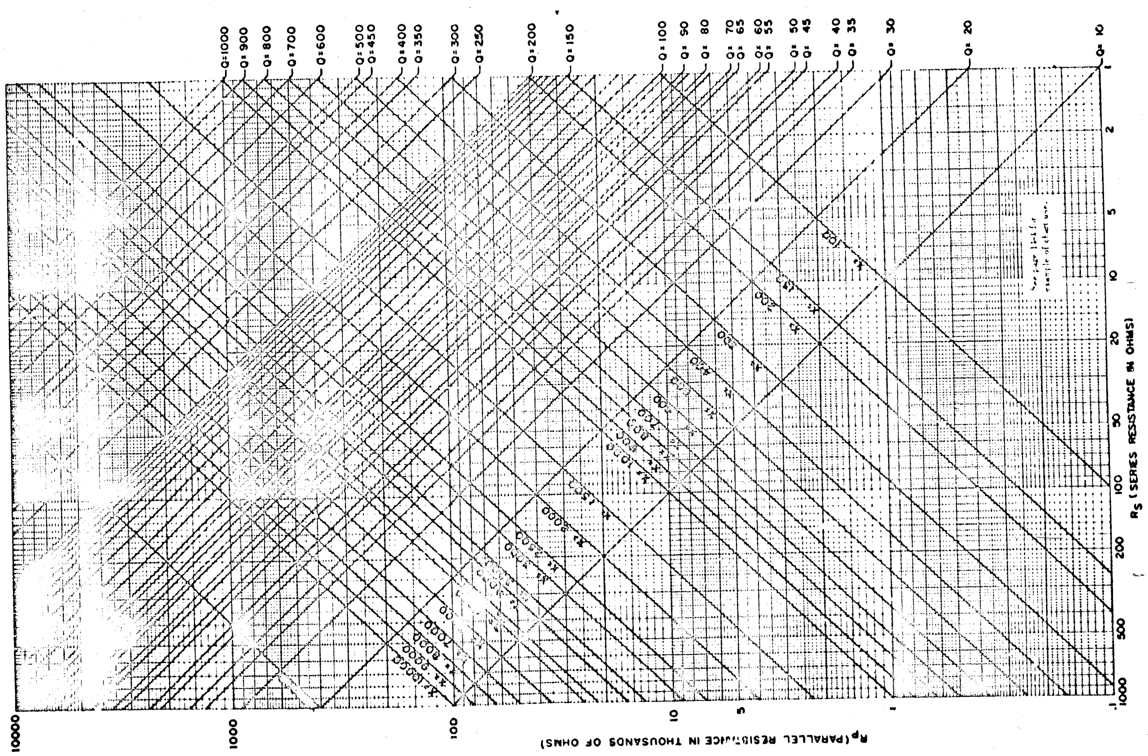


Fig. 14-2 Reactance Chart.

where the reactance is 2860 Ω (≈3000 Ω approx.) and Q=113 (≈150 approx.), the value of R_s can be estimated by looking at the cross-point of the X=3000 and Q=150 lines.

The value of R_s=20 is read on the abscissa. The value of R_p is determined from the ordinate corresponding to the same cross-point, i.e., R_p = 450,000 ohms.

THE SERIES RESONANT CIRCUIT

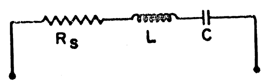
A series circuit consisting of R_s, L, and a capacitor C is depicted in Fig. 14-3. Here the impedance at any frequency is

$$^{1}\ Z_{ab} = R_s + jX_L - jX_C$$

$$\text{or } Z_{ab} = R_s + j(\omega L - \frac{1}{\omega C}) \qquad (7)$$

where:

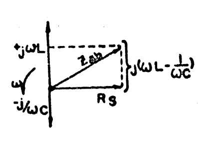
R_s = effective series resistance of circuit
ωL = inductive (positive) reactance = X_L
$\frac{1}{\omega C}$ = capacitive (negative) reactance = X_C



$Z_{ab} = R_s + j(\omega L - 1/\omega C)$

$Z_o = R_s$

Fig. 14-3 Series circuit of R, L, and C.



1 Just as the inductive vector was considered as leading the resistance vector by 90°, so the capacitive vector is considered lagging the resistance vector by 90° (see figure above). The net reactance at any frequency is the difference in magnitudes between ωL and 1/ωC; the impedance is the vector sum of this difference and R_s

In the analysis of practical circuits involving capacitors, no loss resistance is generally ascribed to the capacitor as had been done for the inductor. This stems from the fact that the capacitor loss resistance is generally so low compared to the coil resistance that the former can be ignored in the circuit calculations without introducing much error. Stated in another way, the Q's of the common air variable and mica capacitor used in R-F circuits are so high (Q's up to 5000 and higher are not unusual) that the capacitor losses in comparison to those of the average coil can be neglected. When exceptionally high coil Q's are encountered, say, Q=300, then the losses of the capacitor with, say, a Q=3000 cannot be ignored unless an error of almost 10% in the assumed Q=300 can be tolerated; the value of R_s in such a case should be augmented by the calculated value of the loss resistance of the capacitor – that is, the value of R_s in Fig. 14-3 should be the total losses of the capacitor as well as the inductor.

At one specific frequency, the net reactance ($\omega L - 1/\omega C$) of equation (7) reduces to zero. This frequency is called the *resonant frequency* and will be designated hereafter as f_o. The corresponding angular velocity ($\omega = 2\pi f_o$) will be designated ω_o and the impedance at resonance will be Z_o.

Since the net circuit reactance is zero at resonance, then equation (7) at $f = f_o$ reduces to

$$Z_o = R_s \qquad (8)$$

indicating that the only component remaining to limit current flow is the series resistance. It follows then that with constant voltage E applied to such a series circuit of R, L, and C, current will vary with frequency in the manner shown in in Fig. 14-4, reaching a maximum at resonance when the circuit impedance is a minimum.

Since at resonance the magnitudes of the reactances are equal, namely $\omega_o L = 1/\omega_o C$, therefore

$$\omega_o^2 = \frac{1}{LC} \qquad (9)$$

Since $\omega_o = 2\pi f_o$, then

$$f_o = \frac{1}{2\pi\sqrt{LC}} \qquad (10)$$

Equations 9 and 10 express a very fundamental relation between the series resonant fre-

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quency and the L and C constants of the circuit; it will be shown later that similar relations exist for parallel resonance. Since this resonance formula is used repeatedly in coil and circuit design, it should be known well enough to be used without reference.

If the relation $C = \frac{1}{\omega_0^2 L}$ from (9) is now substituted in the general impedance equation (7), then

$$Z_{ab} = R_s + j\omega L\left(\frac{\omega^2 - \omega_0^2}{\omega^2}\right)$$

If the last term of the above equation is multiplied by $\left(\frac{\omega\,\omega_0}{\omega\,\omega_0}\right)$, which is equivalent to 1, we have

$$Z_{ab} = R_s + j\frac{L\omega_0}{\omega\,\omega_0}\left(\omega^2 - \omega_0^2\right)$$

Since $\omega = 2\pi f$ and $\omega_0 = 2\pi f_0$

$$Z_{ab} = R_s + j\frac{L\omega_0\, 4\pi^2}{4\pi^2 f f_0}\left(f^2 - f_0^2\right)$$

$$= R_s + jL\omega_0\left(\frac{f^2 - f_0^2}{f f_0}\right) \qquad (11)$$

which can be written

$$Z_{ab} = R_s + j\omega_0 L\left(\frac{f+f_0}{f}\right)\left(\frac{f-f_0}{f_0}\right)$$

For values of f near f_0,

$$\frac{f+f_0}{f} \approx 2$$

And since $f - f_0 = \Delta f$, equation 11 reduces to the following simple form which assumes arithmetic symmetry in the resonance curve: [1]

then

$$Z_{ab} = R_s + j\omega_0 L\left(\frac{2\Delta f}{f_0}\right) \qquad (12)$$

[1] Actually, the resonance curve as defined by equation (11) has geometric symmetry about f_0—that is the absolute values for points on the resonance curve are equal at frequencies f and f_0^2/f having f_0 as their geometric mean. For Q's above about 20 the assumption of arithmetic symmetry (the absolute values for points on the resonance curve are equal at frequencies f and $2f_0 - f$ having f_0 as their arithmetic mean) introduces negligible error and at the same time tremendously simplifies the resonance formula (equation 12). For wide band amplifiers (see subsequent paragraphs on staggered single tuned circuits page 27) it is necessary to use the basic equation (equation 11).

The current for a small deviation Δf can now be written as

$$I = \frac{E}{Z_{ab}} = \frac{E}{R_s + j\omega L(2\Delta f/f_0)}\,.$$

Since the current at resonance was $I_0 = E/R_s$, then the manner in which the current falls off from the peak resonance value can be best expressed by the ratio of I/I_0, or

$$\frac{I}{I_0} = \frac{E/Z_{ab}}{E/R_s} = \frac{1}{1 + j(\omega L/R_s)(2\Delta f/f_0)}$$

$$= \frac{1}{1 + jQ(2\Delta f/f_0)} \qquad (13)$$

Expressed in absolute [2] values, this relation becomes

$$\left|\frac{I}{I_0}\right| = \frac{1}{\sqrt{1 + \left(Q\frac{2\Delta f}{f_0}\right)^2}} \qquad (14)$$

or in terms of attenuation from the response at resonance, db Attenuation [3] $= t = 20\log\left(\frac{I}{I_0}\right)$

$$= -20\log\left[1 + \left(Q\frac{2\Delta f}{f_0}\right)^2\right]^{1/2} \qquad (15)$$

and in terms of *times down* [3] = T

$$= \left[1 + \left(Q\frac{2\Delta f}{f_0}\right)^2\right]^{1/2} \qquad (16)$$

A curve, generally referred to as the *selectivity or response curve* can be drawn based on equation 14, and can be used to quickly establish the value of current for any deviation of frequency around resonance knowing just the Q and the resonant frequency f_0 (see Fig. 14-4). It will be shown later in the discussions of the parallel resonant circuit, that the identical relation is applicable to parallel resonance with response being measured in terms of the impedance ratio rather than the current ratio.

[2] The symbol | | means that the included quantity, in this case $\frac{I}{I_0}$, is being expressed in absolute terms and not in the complex form using the j operator. The absolute value is equal in magnitude to the square root of the sum of the squares of the resistive and reactive terms.

[3] See pages 14 and 15.

Referring to Fig. 14-4, it is common and convenient to express the falling characteristic on each side of resonance in terms of attenuation, either as a numerical factor (such as "2 times down" when the response falls to 0.5 of the maximum value, or "10x down" as when the response falls to 0.1), or in terms of decibels. [1] In the latter case, the characteristic is said to be "6db down" when the response falls to 0.5, for example -- or "20db down" if it falls to 0.1. Note that for every value of "db down" there is a specific value of Δf associated with a given coil -- the given coil in the illustration (Fig. 14-4) having a Q of 50. The quantity Δf is sometimes referred to as the "half-bandwidth"; the quantity $2\Delta f$ is better known as the "bandwidth" and actually is the difference in frequency between the same "db down" points on each side of the resonance curve.

[1] The decibel notation is explained in the discussion of amplifiers (see pages 14 and 15).

When the term "bandwidth" is used without specifying the point at which this bandwidth is to be measured, it is generally assumed to refer to the half power or 3db point.

This bandwidth at the 3db point will be designated as "BW" in the following text. It should be noted that if the quantity $\left(Q\frac{2\Delta f}{f_0}\right)$ of equation 14 is set to equal one, then the ratio of $\frac{I}{I_0}$ becomes 0.707, the 3db point referred to above.

It follows then that

$$\left(Q\frac{2\Delta f}{f_0}\right) = 1 = k$$

or

$$2\Delta f = BW = \frac{f_0}{Q} \qquad (17)$$

This simple relation is a basis for simple determinations of Q or BW.

If the quantity

$$a = Q\frac{2\Delta f}{f_0} = Q\left(\frac{\text{cycles off resonance}}{\text{resonant frequency}}\right)$$

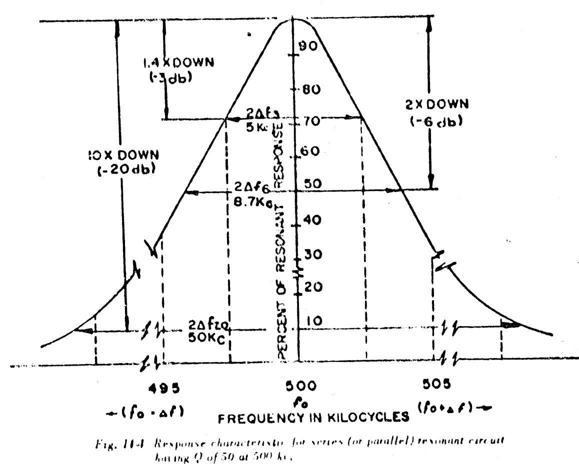


Fig. 14-4 Response characteristic for series (or parallel) resonant circuit having Q of 50 at 500 kc.

of equation 14 is plotted against the current ratio I/I_0, a curve called the *Universal Resonance Curve* (Fig. 14-5) is obtained which is applicable to all coils, irrespective of the values of Q, L, or f_0. With the aid of this curve, the attenuation of the circuit at any frequency may be obtained.

Assume the circuit is tuned to 455 Kc and it is desired to know the impedance at a frequency 9,510 Kc above or below 455 Kc. In this case

$$a = Q\frac{\Delta f}{f_0} = (143)\frac{9510}{455,000} = 3.0$$

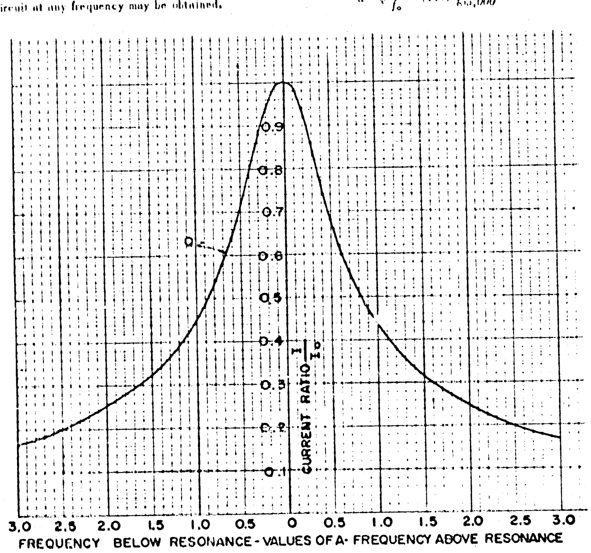


Fig. 14-5. Universal resonance curve for series or parallel resonant circuit.

Example:

Consider a series LC circuit having a 1 millihenry coil which has a Q of 143 at 455 Kc. Since the losses of the capacitor are negligible, the equivalent series resistance and, therefore, the impedance at resonance is 20 ohms (see page 14-3).

The ordinate (Fig. 14-5) corresponding to 3.0 is 0.16, which indicates that the current for this deviation is 0.16 times the current at resonance; this can only occur if the impedance has increased (1/0.16) *over the impedance at resonance, i.e.,* 6(20)=120 *ohms. In a similar fashion the cycles off resonance corresponding to a given current ratio can be conveniently determined.*

It is to be noted that at series resonance the current is equal to E/R_s and that the voltage E_L across the coil is ωL times the current. This means that at resonance

$$E_L = \left(\frac{\omega L}{R_s}\right)E$$

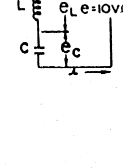
and since

$$\frac{\omega L}{R_s} = Q, \text{ it follows that } E_L = EQ \qquad (18)$$

Since at resonance $\omega L = \frac{1}{\omega C}$, it follows that the voltage across the capacitor is also equal to EQ. It is apparent that with Q's in the order of 100, this voltage can become considerable in some cases, and therefore must be considered in the design of series-resonant assemblies.

Example:

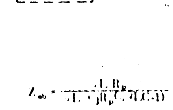
As an illustration of this increase in voltage, consider a series L-C circuit to which 10 volts is applied and in which the coil has an inductance of 10 mh and a resistance of 100 ohms. At 200 kc the reactance of this coil is $2\pi fL$ = 12560 ohms. If this circuit is resonant at 200 kc, then the current will be $\frac{10 \text{ volts}}{100}$ = .1 amperes. A current of .1 amperes flowing through a reactance of 12560 ohms produces a voltage across either of the reactances of 1256 volts--a far cry indeed from the original 10 volts impressed upon the circuit. The designer must recognize this possibility and select suitably rated components where necessary.



The development of the series resonance relations was taken somewhat slowly--first, to permit growing familiarity with the notations and expressions common to coil and circuit design and analysis, and second, to familiarize the engineer with some of the algebraic manipulations inherent in the treatment of coil circuits and in the use of the j operator. The following reviews of parallel resonance and mutually coupled circuits will proceed much faster; the similarity in the series and in the parallel resonant circuit relations will be developed and the applicability of the *Universal Resonance Curve* to parallel circuits will be discussed and illustrated.

THE PARALLEL-RESONANT CIRCUIT

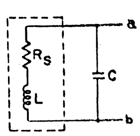
A *parallel-resonant circuit* contains the same elements (R, L, and C) as a series circuit; it is only in the manner of connection that the two types differ. As shown in Fig. 14-6, the parallel circuit can be considered as an inductance L, its parallel loss resistance R_p, and a capacitance all in parallel. Since this loss resistance can also be expressed as a series resistance R_s, another form of the parallel circuit is shown in Fig. 14-7. Both forms of the parallel circuit represent the same performance characteristic.



$$Z_{ab} = \frac{\omega L R_p}{\omega L + jR_p(\omega^2 LC - 1)}$$

$$Z_0 = R_p$$

Fig. 14-6 Parallel-resonant circuit (loss resistance in parallel).



$$Z_{ab} = \frac{\omega^2 L^2}{R_s + j\omega L(\omega^2 LC - 1)}$$

$$Z_o = \frac{L}{R_s C} = R_p$$

Fig. 14-7 Parallel-resonant circuit (loss resistance in series).

In the following analysis of the circuits of Fig. 14-6 and Fig. 14-7 the presentation is somewhat simplified if *admittances* (1/Z) rather than impedances are considered first. Since admittances of parallel branches can be added directly then

$$\frac{1}{Z_{ab}} = j\omega C + \frac{1}{j\omega L} + \frac{1}{R_p}$$

$$= \frac{j\omega C(j\omega LR_p) + R_p + j\omega L}{j\omega LR_p}$$

$$= \frac{R_p(1 - \omega^2 LC) + j\omega L}{j\omega LR_p}$$

Now, inverting, to return the equation to terms of impedance:

$$Z_{ab} = \frac{j\omega LR_p}{R_p(1 - \omega^2 LC) + j\omega L} = \frac{\omega LR_p}{\omega L + jR_p(\omega^2 LC - 1)} \tag{19}$$

A condition of resonance (at which Z_{ab} becomes maximum) exists when $(\omega^2 LC - 1) = 0$, in which case equation 19 reduces to

$$Z_{ab} = Z_o = R_p \tag{20}$$

This condition, as in the case of the series resonant circuit, yields

$$\omega_o^2 = \frac{1}{LC}$$

$$\text{or } \omega_o L = \frac{1}{\omega_o C} \tag{9}$$

and

$$f_o = \frac{1}{2\pi\sqrt{LC}} \tag{10}$$

Since the parallel loss resistance

$$R_p = Q^2 R_s \tag{5}$$

it follows that the parallel representation of Fig. 14-6 can be converted to the series representation of Fig. 14-7. Thus, equation 20 can be expressed in terms of R_s as follows:

$$Z_o = R_p = Q^2 R_s \tag{21}$$

Also, since $Q = \omega L/R_s$ and $\omega L = 1/\omega C$ at resonance, the resonant impedance Z_o can be expressed in any of the following ways:

$$Z_o = R_p = Q^2 R_s = \omega LQ = \frac{L}{R_s C} \tag{22}$$

It will be convenient to remember these impedance relations for parallel resonance since they are used frequently in the course of design operations.

If $LC = 1/\omega_o^2$ is substituted in the general impedance equation (19) for the parallel circuit, the impedance relation becomes:

$$Z_{ab} = \frac{\omega LR_p}{\omega L + jR_p\left(\frac{\omega^2}{\omega_o^2} - 1\right)} \tag{23}$$

If we multiply the denominator by $\frac{\omega\omega_o}{\omega\omega_o}$, which is equal to 1, we have

$$Z_{ab} = \frac{\omega_o LR_p}{\omega_o L + jR_p\left(\frac{1}{\omega\omega_o}\right)(\omega^2 - \omega_o^2)}$$

And since $\omega = 2\pi f$ and $\omega_o = 2\pi f_o$ then

$$Z_{ab} = \frac{\omega_o LR_p}{\omega_o L + jR_p\left(\frac{f^2 - f_o^2}{ff_o}\right)} \tag{24}$$

The term $\frac{f^2 - f_o^2}{ff_o}$ reduces to $\frac{2\Delta f}{f_o}$ for small deviations of frequency from resonance (see development of equation 12). Therefore, equation 24 can be expressed as

$$Z_{ab} = \frac{\omega_o LR_p}{\omega_o L + jR_p\left(\frac{2\Delta f}{f_o}\right)} \tag{25}$$

The ratio of this impedance to that at resonance can now be expressed by dividing equation 22 by $Z_o = R_p$ to give (in absolute value):

$$\left|\frac{Z_{ab}}{Z_o}\right| = \frac{1}{\sqrt{1 + \left(Q\frac{2\Delta f}{f_o}\right)^2}} \tag{26}$$

This relation will be recognized as the *selectivity curve* derived for the series resonant circuit except that here the impedance rather than current is the dependent variable; accordingly, the *Universal Selectivity Curve* of Fig. 14-5 is applicable to the solution of parallel resonant circuits remembering, however, that the impedance is the quantity normalized at resonance in this case.

Example:

Assume a parallel L-C circuit consisting of a 1 mh coil with a Q of 100 at 455 Kc, and a capacitor of 122 μμf. The coil's parallel loss resistance ($R_p = \omega LQ$) is 286000 Ω and the circuit is resonant at 455 Kc ($f_o = \frac{1}{2\pi\sqrt{LC}}$). Therefore, at resonance, the impedance of the circuit (R_p) is also 286000 Ω. Referring to the Universal Resonance Curve (Fig. 14-5), at a frequency 9100 cycles off resonance

$$a = Q\frac{\Delta f}{f_o} = 100\left(\frac{9100}{455000}\right) = 2$$

The ordinate corresponding to 2 is 0.24, indicating that the impedance for this deviation is .24 times the impedance at resonance, or

.24 (286000) = 68640 ohms

AMPLIFIERS

The basic analytical treatment of simple coils and parallel and series circuits given up to this point provides a sufficient background to take up the very important matter of the practical electrical environment in which coils are operated in electronic circuits. The most frequent application of r-f coils is found in the radio frequency Class A amplifier so commonly used in the r-f (front end) stages and in the i-f sections of various types of receivers. The tubes used in these amplifiers have finite input and output impedances of varying degrees depending on the specific tube construction, the frequency of operation, operating point set by the applied voltages and several other factors. Also, the physical circuitry (sockets, wiring, shields, etc.) connected to the tube introduces stray capacitances and loss resistances across the coil so that the performance of the coil usually undergoes a substantial change from that calculated on the basis of the coil elements alone. Thus, the coil and coil-capacitor combinations cannot be regarded as simple devices any longer and the designer must take into account these unavoidable coil loadings imposed by the amplifier circuitry.

Gain

A brief review of some of the basic elements of Class A amplifier performance will be helpful in developing the coil-amplifier relationships to be used by the designer.

Accordingly, consider the amplifier circuit in Fig. 14-8a in which a very small (differential) change (ΔE_{in}) in the grid voltage produces a change ΔI_p in the plate current which in turn produces a voltage ΔE_{out} across the load impedance Z_L[1]. The equivalent circuit is shown in Fig. 14-8b in which the tube is considered as a signal generator with an output voltage $E_p = \mu\Delta E_{in}$ and an internal resistance equal to the plate resistance r_p.

[1]The performances of electron tubes (triodes, tetrodes, pentodes) are characterized by the following terms:
μ = amplification factor = $(\partial E_p/\partial E_g)$ with I_p constant
r_p = dynamic plate resistance = $(\partial E_p/\partial I_p)$ with E_g constant
g_m = transconductance = $(\partial I_p/\partial E_g)$ with E_p constant
The basic relationship among these characteristics is given by $\mu = g_m r_p$.

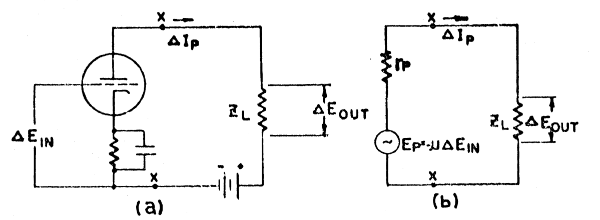


Fig. 14-8 Basic amplifier

Thus:

$$\Delta I_p = \frac{\text{Generator Voltage}}{\text{Tot. Circuit Impedance}} = \frac{(-\mu \Delta E_{in})}{r_p + Z_L}$$

Since $(\Delta I_p)(Z_L) = \Delta E_{out}$, this relation can be rearranged to show

$$\frac{\Delta E_{out}}{\Delta E_{in}} = \frac{E_{out}}{E_{in}} = \frac{-\mu Z_L}{r_p + Z_L} = \text{Voltage Gain of Amplifier}$$

$$= A \qquad (27)$$

Equation 27 is applicable to triodes, tetrodes, and pentodes but a more convenient relation can be expressed for the case of a pentode for which the plate resistance is very high compared to the load resistance Z_L, so that the term Z_L can be neglected in the denominator for all practical purposes. Equation 27 can now be written (noting that $\mu = g_m r_p$):

$$\text{Voltage Gain} = g_m Z_L = A \qquad (28)$$

When several such amplifiers are cascaded in series, as in an i-f amplifier system, the over-all gain is a product of the individual stage gains:

$$A_{over\text{-}all} = (A_1)(A_2) \text{------} (A_n)$$

If the individual stage gains are equal, then

$A_{over\text{-}all} = A^n$ where A = voltage gain of each stage

The arithmetic manipulations involved in calculating system gains by this process are obviously awkward and have led to the adoption of the very convenient "Decibel System" for designating the degree of gain in an amplifier. This system of rating amplifier performance has its definition in the unit called the "Bel" which is the common logarithm of the ratio of output to input power, or:

$$\text{Bel} = \text{Log}_{10} \frac{P_{out}}{P_{in}}$$

Since the normal human ear can detect changes in sound power levels of about one-tenth of a Bel, a more useful unit called the Decibel has been adopted for practical usage with electronic equipment. Thus,

$$\text{Decibel} = \text{db} = 10 \ \text{Log}_{10} \frac{P_{out}}{P_{in}}$$

In the special case where the input and output loads are equal, then:

$$\text{db} = 10 \ \text{Log}_{10} \frac{P_{out}}{P_{in}} = 10 \ \text{Log}_{10} \frac{E^2_{out}/R_{out}}{E^2_{in}/R_{in}}$$

$$= 10 \ \text{Log}_{10} \frac{E^2_{out}}{E^2_{in}} = 20 \ \text{Log}_{10} \frac{E_{out}}{E_{in}}$$

Since E_{out}/E_{in} = Voltage Gain = A, then

$$\text{db} = 20 \ \text{Log}_{10} A \qquad (29)$$

The db voltage gain of n cascaded stages of an amplifier system[1] can now be written as a simple sum, thus:

$$\text{Over-all Voltage Gain (db)} = \text{db}_{stage\,1} + \text{db}_{stage\,2} + \text{------} \text{db}_{stage\,n}$$

If the stage gains are equal, then

$$\text{Over-all Voltage Gain (db)} = n(\text{db per stage})$$

The decibel notation is also a convenience in referring to the losses in power (attenuation or insertion loss) in a network in which case the ratio of P_{in}/P_{out} rather than P_{out}/P_{in}, is used as a basis for reference to avoid fractions. The falling characteristic of voltage or current of a resonant circuit on either side of resonance can also be expressed in terms of "db down"; thus, when the voltage (or current) falls to 1/2, the point is sometimes identified as a simple ratio "2 times down" (abbreviated to "2X down") or as "6db down" (Since db=20Log$_{10}$2=6 db).

MILLER EFFECT

The input impedance of a tube can be considered as consisting of a loss resistance (r_g) and a parallel capacitance from grid to cathode. At low frequencies (up to about 5Mc) the value of r_g is extremely high so that for all practical purposes it can be considered to have a negligible loading effect on any coil circuits feeding the tube. The tube capacitances, however, present a serious problem for they can effectively detune any associated resonant coil circuit or, particularly at high frequencies, they can represent most if not all the capacitance needed to resonate an associated coil. Moreover, this input capacitance is not just the physical grid-cathode capacitance alone, but due to a phenomenon known as *Miller Effect* this grid-cathode capacitance is effectively augmented by the grid-plate capacity multiplied by one plus the stage gain. In Fig. 14-9 are shown the two most important capacitances which contribute to the Miller Effect – the grid to plate and the grid to cathode capacitances.

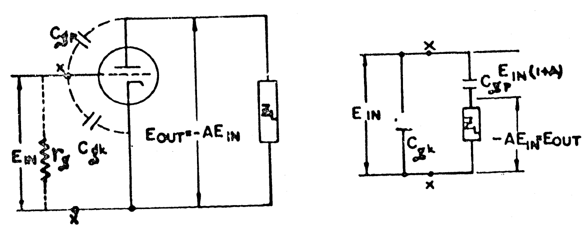


Fig. 14-9 Miller Effect capacitances.

[1]It is somewhat unfortunate that the term *db* has been corrupted in use so that it is frequently applied to rating amplifier voltage gains with complete disregard of the load impedances. As popularly used in this way, *db* has the connotation of voltage gain only and bears no relation to *db* power gain. Thus, a "100 db" amplifier generally refers to one in which the voltage gain is 10^5; unless the input and output impedances are equal, the power gain is *not* 100 db. This misuse, though technically improper, does not cause difficulties when it is clearly evident that voltage gains only are meant, as in the case of i-f, audio pre-amplifier, etc., systems where power, as such, is of no interest.

To fully understand the serious effects of the Miller phenomenon, the output voltage, (E_{out}) is shown equal to the input voltage (E_{in}) times the stage gain (A), or $E_{out} = -AE_{in}$ (the minus sign indicates the phase reversal existing between grid and plate voltages). The voltage across capacitor C_{gp} is the difference between the signal voltage and the output voltage, that is E_{gp}=

E_{in}-(-AE_{in}) = E_{in} (1+A). The figure also shows that both capacitors are effectively in parallel. Since the total charge on the effective input capacitance C_t is q_t, and noting that q=CE, then

$$q_t = q_{gp} + q_{gk} = E_{in}(1+A)C_{gp} + E_{in}C_{gk}$$

From this it follows that

$$C_t = \frac{q_t}{E_{in}} = C_{gk} + (1+A)C_{gp}$$

which means that the grid to cathode capacitance appearing across the input has been effectively increased by the grid to plate capacitance multiplied by one plus the stage gain. It is this "(1+A)C_{gp}" which represents the capacitance due to the *Miller Effect.*

Example:

To give an idea of the magnitude which this total capacitance may assume, consider a 6J5 (triode) tube which has C_{gp} = 3.4 mmf, C_{gk} = 3.4 mmf, and assume a stage amplification of 25. In this case C_t= 3.4+(1+25) 3.4 = 91.8 mmf — a very sizeable input capacitance.

Consider now a 6J16 (pentode) tube which, because of an r-f grounded (screen) grid between the signal grid and plate, has a substantially lower grid to plate capacitance. In this tube C_{gk}= 5.4 mmf, C_{gp}=0.0035 mmf and assume a stage amplification of 130. In this case C_t =5.4+(1+130) 0.0035=5.9 mmf.

These specific examples have shown that a triode may have, due to its higher grid to plate capacitance, nearly sixteen times the effective input capacitance of a pentode. Since the stage amplification is such an important parameter in determining the effective input capacitance of a triode, and since the stage gain can change with fluctuating voltages and tube aging it becomes apparent that considerable detuning and consequent high degree of instability can be expected in most amplifiers using triode tubes. Naturally, this already bad situation would be aggravated in those cases where all, or even a large part of the total tuning capacity across a coil is provided by the tube — a condition that is quite common in high frequency amplifiers.

It may be stated at this point that the Miller Effect is not always an undesired phenomenon. Because by varying the bias on a vacuum tube, the g_m and thereby the stage amplification (A) may be changed, there becomes available through the Miller Effect an instantaneously varying capacitance which may be used for automatic frequency control, variable rate discharge devices, and similar electronic circuit applications.

The Miller Effect superiority of the pentode over the triode for r-f class A amplifier service, plus the fact that much higher stage gains and much lower loading effects on associated coil circuits are realized with pentodes, has made the triode almost obsolete for r-f voltage amplifier applications. Accordingly, the following discussions and design applications of coils will be confined almost exclusively to pentode circuits.

Gain-Bandwidth Product

Consider a pentode amplifier with a parallel resonant circuit as the plate load (Fig. 14-10). At the resonant frequency f_o:

$$A = g_m Z_L = g_m \omega_o LQ = g_m R_p$$

(Eq. 28 modified where $Z_L = \omega_o LQ$)

$$\text{and} \qquad B\overline{W} = 2\Delta f = \frac{f_o}{Q} \qquad (17)$$

The product of A and B$\overline{W}$, called the *gain-bandwidth product*, is a useful design parameter since it forms a convenient basis for selection of the coil constants, particularily in broad band amplifiers. Multiplying Equations 28 and 17:

$$(A)(B\overline{W}) = (g_m 2\pi f_o LQ)\left(\frac{f_o}{Q}\right) = g_m 2\pi f_o^2 L.$$

$$\text{Since } f_o^2 = \frac{1}{4\pi^2 LC}, \text{ then}$$

$$\text{Gain-Bandwidth Product} = AB\overline{W} = \frac{g_m}{2\pi C} \qquad (30)$$

Equation 30 indicates that for a desired single-stage gain and bandwidth, the controlling circuit factor is the ratio of tube transconductance to the total capacitance resonating the coil. Furthermore, if a specific tube is selected (that is, if g_m is fixed) then the value of the resonating capacitor needed to satisfy the gain-bandwidth product can

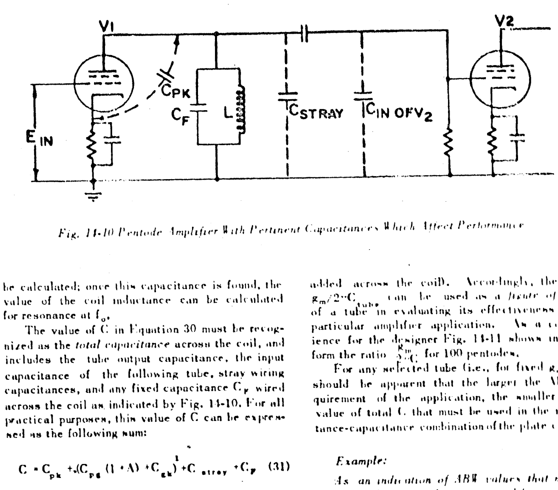


Fig. 14-10 Pentode Amplifier With Pertinent Capacitances Which Affect Performance

be calculated; once this capacitance is found, the value of the coil inductance can be calculated for resonance at f_o.

The value of C in Equation 30 must be recognized as the *total capacitance* across the coil, and includes the tube output capacitance, the input capacitance of the following tube, stray wiring capacitances, and any fixed capacitance C_F wired across the coil as indicated by Fig. 14-10. For all practical purposes, this value of C can be expressed as the following sum:

$$C = C_{pk} + (C_{pg}(1+A) + C_{gk})^1 + C_{stray} + C_F \qquad (31)$$

In a carefully wired circuit, the value of the stray capacitance can be assumed to be about 5 mmf, and the values of the tube interelectrode capacitances can be found in a tube manual, so that the value of C_F, the only unknown, can be calculated. The manner of calculating these capacitances and the use of the gain-bandwidth concept will be illustrated in the subsequent design examples of single-tuned amplifiers.

It should be noted that the ratio of g_m to the tube capacitances depends on the tube selected and is a significant factor in determining the value of the coil inductance to be used, particularly at the higher frequencies where *only* the tube capacitances and the wiring strays are used to resonate the coil (that is, no external fixed capacitor is added across the coil). Accordingly, the ratio $g_m/2\pi C_{tube}$ can be used as a *figure of merit* of a tube in evaluating its effectiveness for a particular amplifier application. As a convenience for the designer Fig. 14-11 shows in chart form the ratio $\frac{g_m}{2\pi C}$ for 100 pentodes.

For any selected tube (i.e., for fixed g_m), it should be apparent that the larger the AB$\overline{W}$ requirement of the application, the smaller is the value of total C that must be used in the inductance-capacitance combination of the plate circuit.

Example:

As an indication of ABW values that may be encountered in single stage amplifiers in commercial design practice, the following examples may be considered representative:

Medium Frequency Amplifiers (455 kc), AM Service

Based on stage gain of 200 and BW of 10 kc

$$AB\overline{W} = 200\,(10)^4 = 2\,(10)^6$$

High Frequency Amplifiers (10.7 Mc), FM Service

Based on stage gain of 60 and BW of 280 kc

$$AB\overline{W} = 60\,(2.8)\,10^5 = 16.8\,(10)^6$$

Very High Frequency Amplifiers (44Mc), General Broad Band Service

Based on stage gain of 10 and BW of 6 Mc

$$AB\overline{W} = 10\,(6)\,10^6 = 60\,(10)^6$$

[1] It is assumed here that the following amplifier stage will use the identical tube, so that the input capacitance of the following tube is $C_{pg}(1+A) + C_{gk}$.

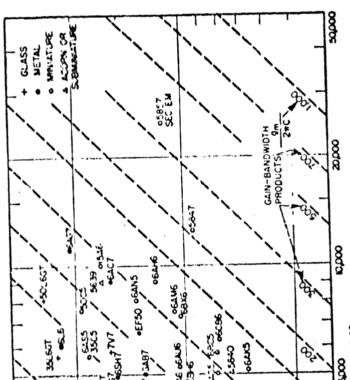


Fig. 11-11 $g_m/2\pi C$ for 100 Pentodes

From these examples it is apparent that for any selected tube the designer may use 30 times as much C in the 455 kc amplifier as is possible in the 44 Mc application and 8 times as much as in the 10.7 Mc application.

It will be shown in the discussions of the various types of cascaded amplifiers that the bandwidth undergoes significant changes as stages are added to an amplifier. In an amplifier of n stages, the over-all bandwidth BW_{oa} bears a specific quantitative relationship to the single stage bandwidth BW (i.e., $KBW = BW_{oa}$ where K is a *bandwidth shrinkage factor*[1]. If the over-all gain of such an amplifier with n stages is designated as A_{oa}, then the mean gain of each stage is $A_{oa}^{1/n}$. It follows that the gain-bandwidth product of the single stage of such an amplifier can be expressed in terms of the over-all parameters as follows:

$$\text{Single Stage Gain Bandwidth Product} = A(BW) = A_{oa}^{1/n}\left(\frac{BW_{oa}}{K}\right)$$

This relation is sometimes expressed as

$$KABW = A_{oa}^{1/n}(BW_{oa})$$

where the quantity ABW *(K)* is called the gain-bandwidth product of the *amplifier* and it is equal to the mean stage gain multiplied by the over-all bandwidth. In this text this *amplifier* product will be distinguished from the single stage gain-bandwidth product by italicizing [i.e., (K)ABW = *ABW*]

The convenience of the gain-bandwidth product as a design parameter will be illustrated in the several examples to be given in connection with the discussion of multi-stage amplifiers. A simple academic example to show the mechanics of using this product is as follows:

Example:

In designing a receiver with a 44 Mc i-f system, calculations show that one of the i-f stages requires a bandwidth of 2 Mc. A tube with g_m = 5000 μmhos and C_{tube} = 7 μμfds is to be used.

(a) Determine the coil constants assuming a stray wiring capacity of 7 μμfds.

(b) What is the maximum stage gain obtainable?

(c) If for some reason this stage gain is considered too high, how can it be reduced by 50%?

(d) What could be done to increase the gain by 50% while still retaining same BW?

(a & b) For maximum gain, the tube and stray capacities only will be used to resonate the coil. Thus

$$ABW = \frac{g_m}{2\pi C} = \frac{(5000)(10^{-6})}{2\pi(7+7)(10^{-12})} = 56.8(10^6)$$

$$\text{Stage Gain } A = \frac{(ABW)}{BW} = \frac{56.8(10^6)}{2(10^6)}$$

$$= 28.4 \text{ max. stage gain}$$

$$\text{Now } Q = \frac{f_o}{BW} = \frac{44(10^6)}{2(10^6)}$$

$$= 22 \text{ (This is effective circuit Q)}$$

$$L = \frac{1}{\omega_o^2 C} = \frac{10^6}{(2\pi)^2(44)(10^6)^2(7+7)10^{-12}} = 0.93 \ \mu h.$$

(c) *For a given ABW product, the gain can be decreased only at the expense of increased bandwidth. If the bandwidth specified is to be retained, the simplest recourse is to add sufficient fixed capacity to reduce the stage gain-bandwidth product, $\frac{g_m}{2\pi C}$, by the desired amount. In this instance 50% of 28.4 = 14.2 = reduced gain. The reduced gain-bandwidth product therefore is*

$$(14.2)(2 \times 10^6) = \frac{g_m}{2\pi C} = \frac{5000(10^6)}{6.28\,C} \text{ or } C = 28 \ \mu\mu fds.$$

Since 14 μμfds are already available as tube and stray capacities, a fixed capacitance of 14 μμfds should be wired across the coil.

(d) *It is impossible in this instance to increase the gain of the stage beyond that permitted by the use of the lowest net capacitance (14 μμfds). The only recourse is to use a tube with a higher figure of merit $\left(\frac{g_m}{2\pi C_{tube}}\right)$ so that the gain bandwidth product of the stage can be increased by 50%.*

Bandwidth Ratio

In the subsequent discussions of tuned ampli-

[1] The bandwidth shrinkage factor K is discussed in detail on pages 23 and 24.

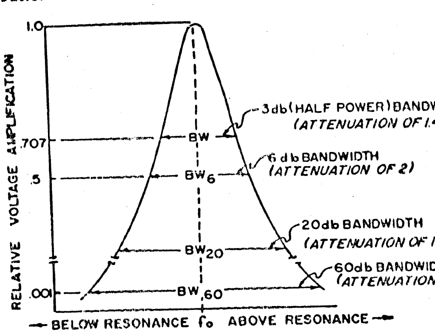


Fig. 14-12 Bandwidth measurement points.

fiers, reference will also be made to one additional characteristic by which the performance of amplifiers can be described — namely, the "bandwidth ratio", sometimes called the "skirt selectivity" or the "skirt ratio". Referring to Fig. 14-12, the ratio of the bandwidth at the 60 db point (BW_{60}) to that at the 6 db point (BW_6) is a measure of the steepness of the sides of the selectivity characteristic and is therefore a measure of the effectiveness of the amplifier tuned circuits in discriminating against unwanted off-channel signals. This quantity is thus defined:

$$\text{BW Ratio} = \frac{BW_{60}}{BW_6}^{1} \qquad (32)$$

The bandwidths at the 60 and 6 db points can always be calculated using the selectivity curve relations applicable to the tuned circuits being used in the amplifier. Since the 3db bandwidth (BW) is usually stated as a design requirement, it is convenient to express other bandwidth measurement points in terms of the 3 db bandwidth. For *n* single-tuned stages in cascade, the bandwidths at the important measurements points are as follows:

$$BW_6 = BW\sqrt{2^{2/n}-1} = K_6 BW$$

$$BW_{20} = BW\sqrt{10^{2/n}-1} = K_{20} BW$$

$$BW_{40} = BW\sqrt{100^{2/n}-1} = K_{40} BW \qquad (33)$$

$$BW_{60} = BW\sqrt{1000^{2/n}-1} = K_{60} BW$$

1 When only one stage performance is being described, it is common to define BW Ratio in terms of the 20 db and 6 db bandwidths. The reason for this follows from the fact that the 60 db bandwidth for a single-tuned stage (or even in a double tuned stage) is exceedingly large. A more practical value for the ratio results using the moderate 20 db point for reference.

INDUCTIVE COUPLING DEVICES

Inductive coupling is a term covering a variety of circuit configurations having inductive elements which are used, primarily, for the purpose of impedance matching and coupling the output of one circuit to the input of a subsequent circuit. There are two basic ways by which plate to grid coupling may be accomplished:

(a) Impedance Coupling: By the use of a high-impedance inductive network isolated from the following stage by a capacitor, or

(b) Transformer Coupling: By the use of a transformer, either single or double-tuned, in which the primary and secondary circuits are completely insulated, one from the other, or by the use of tapped inductors (auto transformers).

Each of the above can be further sub-divided as shown in Fig. 14-13. Fig. 14-14 shows schematically these coupling methods. The discussions that follow will take up each method and show the coil design considerations that apply in designing an amplifier system. Each will be illustrated by a practical example.

The simplified analytical treatments of coils and series and parallel resonant circuits, previously given will suffice for analysis and design of impedance-coupling elements. A general analysis of mutual-inductive coupling will be presented as part of the discussion of transformer-coupled systems.

IMPEDANCE COUPLING

The high-impedance reactive network commonly known as impedance coupling is the simplest

type of inductive-coupling device. It consists of a single impedance, generally in the plate circuit of an amplifier stage and coupled to the grid circuit of the following stage by a coupling capacitor, which isolates the plate voltage of the first tube from the grid of the following tube. Impedance coupling may be further classified into untuned and single-tuned circuits. Fig. 14-14a and 14-14b illustrate schematically these classifications.

Disregarding stray-wiring capacitance and the input and output capacitance of the associated tubes, the impedance of the untuned coil will be ωL. If this coil is tuned to resonance, the impedance will be increased Q times ($Z = \omega LQ$) thereby resulting in a Q-times increase in stage gain with the same coil since stage gain = $A = g_m \omega LQ$. The tuned circuit not only provides a higher load impedance without change in the inductor, but also provides frequency discrimination (band-pass) not inherent in the simple untuned coil. However, simple impedance coupling does not provide impedance transformation to match unequal impedances unless special divider networks are used.

UNTUNED COILS

While in the true sense, there is no such thing as an untuned coil (distributed capacitance and external circuit capacitance are always present), it is common to refer to a circuit as *untuned* when

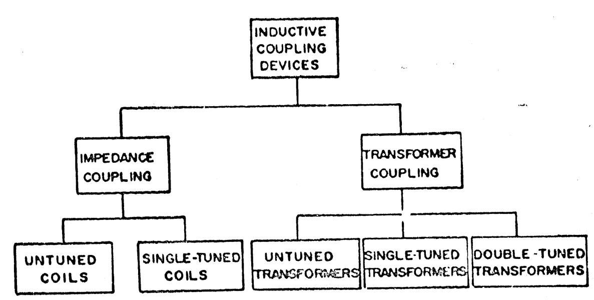


Fig. 14-13 The family of common inductive-coupling devices.

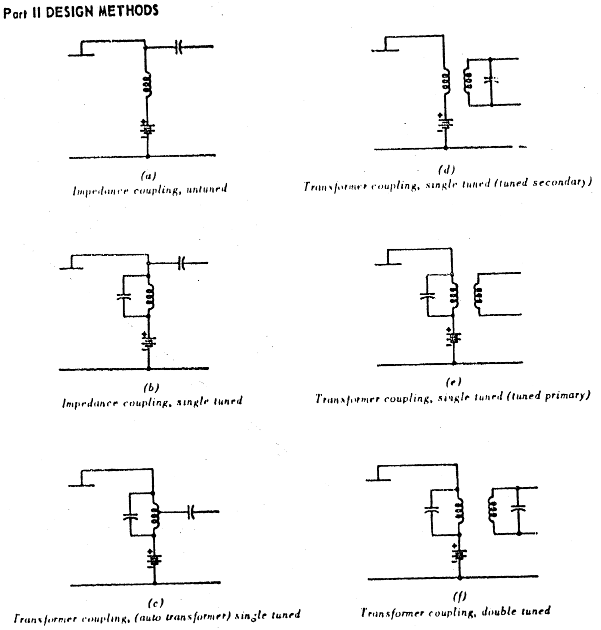


Fig. 14-14 Methods of using inductances as coupling devices.

there is no attempt made to provide means for tuning, and the distributed and external circuit capacitances in combination with the inductor resonate at a frequency either above or below the desired operating band.

A practical example of an untuned coil is the common plate-choke often used in vacuum tube circuits (Fig. 14-14a). The value of inductance is chosen so that it will resonate outside of the desired operating range with the external circuit capacitance and its own distributed capacitance.

SINGLE TUNED COILS[1]

Single-tuned circuits are common in both wide and narrow-band amplifiers and are used at all frequencies. When used between tubes, they represent the simplest form of frequency-selective coupling.

Figure 14-10 shows a single-tuned amplifier stage with the various associated capacitances which are identified in the text accompanying the figure. The single-stage characteristics have been given in part in the preceding general discussions of amplifier performance. A full characterization in terms of the various design parameters follows.

Single-tuned circuit characteristics (one stage):

$$\text{Gain: } A = g_m Z_L = g_m \omega L Q = g_m R_p \qquad (34)\ (28\ \text{Modified})$$

$$\text{Attenuation (in db)} = -20 \text{ Log } [1 + (Q \frac{2\Delta f}{f_o})^2]^{\frac{1}{2}} \qquad (15)$$

or

$$\text{Attenuation (in times down)} = [1 + (Q \frac{2\Delta f}{f_o})^2]^{\frac{1}{2}} \qquad (16)$$

$$\text{3db bandwidth: } BW = 2\Delta f = \frac{f_o}{Q} \qquad (17)$$

$$\text{Bandwidth ratio: BW Ratio} = \frac{BW_{60}}{BW_6} = 577 \qquad (32)$$

$$\text{BW Ratio} = \frac{BW_{20}}{BW_6} = 5.76 \qquad (35)$$

$$\text{Gain bandwidth product: } ABW = \frac{g_m}{2\pi C} \qquad (30)$$

Note particularly the high BW_{60}/BW_6 ratio, indicating the rapid flaring of the skirt of the selectivity characteristic and denoting relatively poor rejection of unwanted signals outside of the pass band (the bandwidth at the 3db point). This skirt selectivity can be improved by additional stages cascaded and tuned to the same center frequency (synchronously tuned).

[1] This discussion is based in part on Chapter 4 of the first edition of Vacuum Tube Amplifiers by George E. Valley and Henry Wallman published by McGraw Hill Book Co. Inc., 1948. The original material has been substantially abbreviated, and for purposes of uniformity, there are changes in notations none of which affects in any way the contents of the material or its accuracy.

Cascaded Synchronous Single-tuned Circuits (n stages):

Let overall gain of n stages = A_{on} and overall bandwidth at 3db = BW_{on};

then:

$$\text{Overall gain: } A_{on} = A^n \qquad (36)$$

$$\text{Gain of one stage: } A = A_{on}^{1/n} \qquad (37)$$

$$\text{Attenuation (in db)} = -(n)\,20 \log [1 + (Q \frac{2\Delta f}{f_o})^2]^{\frac{1}{2}} \qquad (38)$$

or

$$\text{Attenuation (in times down)} = [1 + (Q \frac{2\Delta f}{f_o})^2]^{n/2} \qquad (39)$$

$$\text{3 db bandwidth: } BW_{on} = BW \sqrt{2^{1/n}-1} = K_1 BW \qquad (40)$$

Overall Bandwidth ratio: (41)

$$BW_{on} \text{ Ratio} = \frac{BW_{60}}{BW_6} = \frac{\sqrt{1000^{2/n}-1}}{\sqrt{2^{2/n}-1}}$$

Amplifier Gain Bandwidth Product: (42)

$$ABW = (A_{on}^{1/n})\, BW_{on} = (A_{on}^{1/n})\, BW \sqrt{2^{1/n}-1}$$

$$= ABW \sqrt{2^{1/n}-1} = K\, ABW.\ [2]$$

A comparison of the n stage and single stage attenuation, (selectivity) equations 38 and 15 indicates that for the same excursions of Δf on each side of the resonant frequency, the attenuation (in db) is n times greater for the cascaded circuit. Thus in Fig. 14-15 the 3 db attenuation for one stage is indicated by point a. For the same Δf, as additional stages are added, this attenuation falls 3 db per added stage, or n (3db) for n stages; i.e., the attenuation of two stages is 2 (3db) or 6 db as shown by point b; and points c and d indicate attenuation for four and eight stages respectively.

Equation 40 shows the overall bandwidth of a synchronous cascaded amplifier to be functions of the single-stage bandwidth and the number of stages. Since the expression $\sqrt{2^{1/n}-1}$ is always numerically smaller than 1, the over-all bandwidth

[2] On page 19 in the discussion of ABW it was shown that $ABW = A\Delta WK$. The quantity K is here identified as $\sqrt{2^{1/n}-1}$ for this type of amplifier.

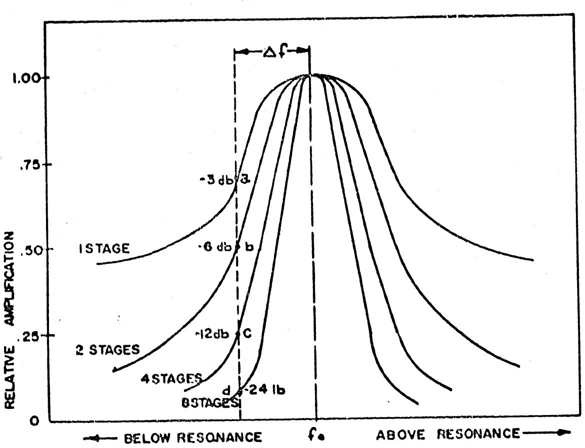


Fig. 14-15 Attenuation for Δf due to additional stages (synchronous tuned amplifier).

is always narrower than the single-stage bandwidth as shown by Fig. 14-15. This quantity $(\sqrt{2^{1/n}-1})$ is known as the *bandwidth shrinkage factor*, previously identified as *K* (pg. 19). Fig. 14-16 presents in tabular form the values of the bandwidth shrinkage factor for n-stage synchronous amplifiers showing how rapidly the bandwidth shrinks as stages are added. This rapid narrowing of bandwidth can be considered the principal weakness of this type of amplifier for wide-band applications.

The last column of Fig. 14-16 shows the 60 to 6 db bandwidth ratio, as the number of synchronous stages is increased. This ratio indicates the steepness of the skirts of the over-all selectivity curve and consequently, the adjacent channel attenuation characteristics of the amplifier. Comparing the bandwidth ratios, it should be noted that up to five stages there is a rapid increase in skirt slope, but beyond five stages the skirt slope changes slowly.

The table of Fig. 14-16 is based upon the resonance curve defined in Fig. 14-38 and this table, in conjunction with Eq. 33, will be found convenient for the computation of system selectivity characteristics as will be shown by the following example.

Example:

The effects of bandwidth shrinkage can be appreciated if we consider a nine-stage synchronous single-tuned amplifier having a required bandwidth of 4 Mc.

By Eq. 40 and the table of Fig. 14-16, the single-stage bandwidth is

$$BW = \frac{BW_{sys}}{K} = \frac{4}{.28} = 14.3 \text{ Mc}$$

Bandwidth factor: $K = \sqrt{T^{2/n}-1}$

Total number of circuits n	T=1.12 t=1db	T=1.26 t=2db	T=1.41 t=3db	T=2.0 t=6db	T=4.0 t=12db	T=7.0 t=17db	T=10 t=20db	T=20 t=26db	T=40 t=32db	T=70 t=37db	T=100 t=40db	T=1000 t=60db	T=10⁴ t=80db	$\frac{BW_{60}}{BW_6}$
1	.51	.77	1.00	1.73	3.9	6.9	10.0	20.0	40.0	-	-	-	-	577
2	.35	.51	.64	1.00	1.7	2.2	3.0	4.4	6.3	8.3	10.0	-	-	33
3	.28	.41	.51	.77	1.2	1.7	1.9	2.5	3.3	4.0	4.5	10.0	21.7	13
4	.24	.35	.44	.64	1.0	1.3	1.5	1.9	2.3	2.7	3.0	5.5	10.0	8.6
5	.22	.31	.39	.57	.86	1.1	1.2	1.5	1.8	2.1	2.3	3.9	6.3	6.8
6	.20	.28	.35	.51	.77	.96	1.1	1.3	1.6	1.8	1.9	3.0	4.6	5.9
7	.18	.26	.32	.47	.70	.86	.96	1.16	1.4	1.54	1.65	2.5	3.6	5.3
8	.17	.24	.30	.44	.64	.79	.88	1.09	1.23	1.38	1.47	2.2	3.0	5.0
9	.16	.23	.28	.41	.60	.74	.82	.98	1.13	1.26	1.33	1.9	2.6	4.7

Fig. 14-16 Bandwidth factor K for synchronous single-tuned amplifiers and bandwidth ratio for 60 to 6 db.

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The complete selectivity characteristic can be determined by the use of the appropriate value of $\underline{k}$ from the table of Fig. 14-16 and Eq. 33.

By Eq. 33:

$$BW_6 = \underline{k}_6 BW = .41\,(14.3) = 5.85 \text{ kc}$$
$$BW_{20} = \underline{k}_{20} BW = .82\,(14.3) = 11.7 \text{ kc}$$
$$BW_{40} = \underline{k}_{40} BW = 1.33\,(14.3) = 19.0 \text{ kc}$$
$$BW_{60} = \underline{k}_{60} BW = 1.9\,(14.3) = 27.2 \text{ kc}$$

The cascaded-synchronous single-tuned amplifier is the simplest type for wideband application but, as shown above, it suffers from rapid bandwidth shrinkage. The usefulness of this amplifier for a desired application, must be adjudged in each individual case by analysis of the over-all requirements in terms of the number of stages needed and the reasonableness of the calculated individual stage bandwidth.

Cascaded-synchronous single-tuned amplifiers are used primarily in applications such as panoramic receivers, radar search equipment, and in radio frequency-actuated relays or other applications where the narrow bandwidth can be tolerated or is desired. It should also be noted that as the number of stages increases, the *bandwidth shrinkage factor* approaches zero and, therefore, the amplifier gain-bandwidth product (42) also approaches zero.

If the designer is confronted with the problem of a specified over-all gain, for which the condition of maximum bandwidth is desired, it can be shown that this maximum condition can be realized when the gain per stage is equal to $\sqrt{e}$ (= 4.34 db). Any design which would add more stages will only serve to narrow the bandwidth of the amplifier below this maximum.

Example No. 1:

Required, a six-stage 110-db synchronous single-tuned amplifier having an over-all bandwidth of 2.16 Mc and centered at 30 Mc.

$$\text{Stage gain} = A = \frac{110}{6} = 18.3 \text{ db or } 8.25$$

The bandwidth shrinkage factor from Fig. 14-16 is .35 then:

$$BW = \frac{BW_{oa}}{.35} = \frac{2.16}{.35} = 6.17 \text{ Mc}$$

$$\text{and } ABW = 8.25(6.17)10^6 = 51(10^6)$$

If we choose to use a 6 AC7 tube having a g_m of 8000 μmhos, we find

$$C = \frac{g_m}{2\pi ABW} = \frac{8000(10^{-6})}{6.28(51)10^6} = 25\,(10^{-12})$$

or 25 μμf

The value of L for resonance at 30 Mc with 25 μμf is

$$L = \frac{1}{4\pi^2 f^2 C} = 1.13 \ \mu h.$$

$$\text{Since } BW = \frac{f_0}{Q} = \frac{30(10^6)}{Q} = 6.17(10^6)$$

$$Q = \frac{30(10^6)}{6.17(10^6)} = 4.9$$

Since $A = g_m R$ then

$$R = \frac{\text{stage gain}}{g_m} = \frac{8.25}{8000(10^{-6})} = 1030 \text{ ohms}\ ^1$$

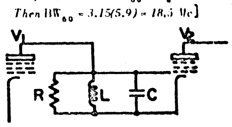
$$\text{And by (Eq. 33) } BW_6 = BW\sqrt{2^{2/n}-1}$$

$$= 6.17\sqrt{2^{1/3}-1} = 3.15 \text{ Mc}$$

$$\text{Also } BW_{60} = BW\sqrt{1000^{2/n}-1} =$$

$$6.17\sqrt{1000^{1/3}-1} = 18.5 \text{ Mc}$$

[or by Fig. 14-16 $BW_{60}/BW_6 = 5.9$
Then $BW_{60} = 3.15(5.9) = 18.5$ Mc]



AC diagram of single-tuned amplifier stage

[1] This value of R represents the total damping resistance across the tuned circuit. It will be found in this instance that the practical values of coil losses, plate resistance and input resistance of the next stage (all of which go to make up R) are by far inadequate to depress the Q to the low value (4.9) desired. Accordingly, a damping resistor of about 1000Ω has to be used in parallel with the coil to get the specified performance.

Where:

C is total circuit capacity: output capacity of V_1, plus input capacity of V_2, plus wiring capacity.

R is total parallel circuit resistance: parallel resistance of the load resistor, the plate resistance of V_1, the input resistance of V_2, plus the equivalent shunt loss resistance of L and C.

Example No. 2:

Required, a narrow band, four stage 80-db synchronous single-tuned amplifier having an over-all bandwidth of .095 Mc centered at 4.3 Mc and using battery tubes.

$$\text{Stage gain} = A = \frac{80}{4} = 20 \text{ db or } 10$$

Bandwidth shrinkage factor from Fig. 14-16 is .44, then:

$$BW = \frac{BW_{oa}}{.44} = \frac{.095}{.44} = .216 \text{ Mc}$$

$$\text{and } ABW = 10(.216)10^6 = 2.16(10^6)$$

If we choose to use a CK-569 AX tube having a g_m of 1100 μmhos, we find

$$C = \frac{g_m}{2\pi ABW} = \frac{1100(10^{-6})}{6.28(2.16)10^6} = 81.2 \ \mu\mu f$$

The value of L for resonance at 4.3 Mc with 81.2 μμf is

$$L = \frac{1}{4\pi^2 f^2 C} = 16.85 \ \mu h$$

$$BW = \frac{f_0}{Q} = \frac{4.3(10^6)}{Q} = .216(10^6) \text{ or } .216 \text{ Mc}$$

$$Q = \frac{4.3(10^6)}{.216(10^6)} = 20$$

Total parallel loss resistance

$$R = \frac{\text{stage gain}}{g_m} = \frac{10}{1100(10^{-6})} = 9100 \text{ ohms}$$

$$\text{By Eq. 33: } BW_6 = BW\sqrt{2^{2/n}-1} = .216\sqrt{.414} =$$

$$.216(.643) = .139 \text{ Mc}$$

$$\frac{BW_{60}}{BW_6} = 8.6 \text{ (from Fig. 14-16)}$$

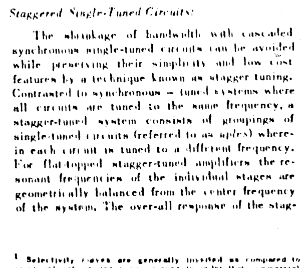
$$\therefore BW_{60} = 8.6(BW_6) = 8.6(.139) = 1.195 \text{ Mc}$$

Also

$$BW_{20} = BW\sqrt{10^{2/4}-1} = .216\sqrt{2.16} = .315 \text{ Mc}$$

$$BW_{40} = BW\sqrt{100^{2/4}-1} = .216\sqrt{9} = .648 \text{ Mc}$$

The complete selectivity characteristic of the amplifier is shown below:[1]



Staggered Single-Tuned Circuits:

The shrinkage of bandwidth with cascaded synchronous single-tuned circuits can be avoided while preserving their simplicity and low cost features by a technique known as stagger tuning. Contrasted to synchronous – tuned systems where all circuits are tuned to the same frequency, a stagger-tuned system consists of groupings of single-tuned circuits (referred to as *n*ples) wherein each circuit is tuned to a different frequency. For flat-topped stagger-tuned amplifiers the resonant frequencies of the individual stages are geometrically balanced from the center frequency of the system. The over-all response of the stage-

[1] Selectivity curves are generally inverted as compared to previously shown resonance curves in order that commercial semilog paper may be used.

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gered system is the product of the responses of individual single-tuned circuits, and can be designed to be relatively flat and have any desired bandwidth. Improvement of *ABW* product and skirt ratio, if desired, can be achieved by cascading two or more staggered groups.

Stagger-tuned amplifiers offer greater efficiency (i.e., gain-bandwidth product) than synchronous-tuned amplifiers and, therefore, are more readily adapted to wide-band applications. The stagger-tuned alignment procedure is more complex but must often be tolerated if certain bandpass characteristics are required.

Before discussing stagger tuning further, it is essential that the limitations in accuracy of some of the formula developed in the earlier part of this Section be re-stressed. It was assumed in previous analyses that Δf was an equal excursion on each side of the center frequency (arithmetic symmetry) and the selectivity characteristic so derived was indicated to be sufficiently accurate for relatively narrowband designs. In circuits with Q's of about 20 or higher, the formulas are fully adequate. However, in very broad-band designs where total bandwidths of more than 30% of the center frequency are involved, use of these simplified expressions is not valid since in the rigorous analysis the symmetry around f_o is geometric, not arithmetic. Accordingly, the following design procedure is presented for the exact case of geometric symmetry and is applicable without qualification in all broad-band designs.

Figure 14-17 illustrates a staggered pair designed to produce a nearly flat-topped over-all response. The center frequency of the over-all response is f_o, the resonant frequency of one stage is f_1 and the resonant frequency of the other stage is f_2. The frequencies are related according to the geometric relationship:

$$\frac{f_o}{f_1} = \frac{f_2}{f_o} = \alpha \qquad (43)$$

$$\text{then } f_1 = \frac{f_o}{\alpha} \qquad (44)$$

$$\text{and } f_2 = \alpha f_o \qquad (45)$$

$$\text{Note that } f_o = \sqrt{f_1 f_2} \qquad (46)$$

Figure 14-18 illustrates a staggered triple designed to produce a nearly flat-topped over-all response as indicated by the center frequency curve. The center frequency of the over-all response is f_o. One of the stages is tuned to this center frequency; of the remaining two stages, one is tuned to f_1 and the other to f_2. The frequencies f_1 and f_2 are designated by equations 44 and 45. As will be shown, staggered triples have narrower skirts than staggered pairs.

Stagger-tuned amplifiers may have almost any number of stages. It is common practice to speak of *staggered n-uples*, where n is the number of stages in the staggered group or uple. As the number of staggered stages (n) is increased, the skirt of the over-all selectivity response becomes narrower (see Fig. 14-19). A system consisting of too many staggered stages can be difficult to align in practice because of the many different frequencies involved. In addition, overstaggering (a response with multiple peaks and dips between) may result. Therefore, several staggered groups are often cascaded to meet the specified selectivity characteristics.

The mathematical analysis of staggered circuits is complex and has been treated in several texts.[1] For these reasons and for simplification of this manual, the complete mathematical analysis has been omitted and only the working relationships are presented in the following paragraphs.

For the better understanding of stagger-tuned systems, the table of Fig. 14-19 compares them with single-tuned single-stage and cascaded-synchronous amplifiers. However, before studying the table, several fundamental definitions must be well understood.

- n = number of single-tuned circuits in a staggered group, or *n-uple*.
- m = number of staggered groups *(n-uples)* in cascade.
- u = total number of single-tuned stages in system. (u = nm for staggered systems).
- A = gain of single-stage single-tuned circuit.
- A_n = gain of staggered group of *n-uple*.
- A_{oa} = gain of the over-all system of m cascaded *n-uples*. ($A_{oa} = A_n^m$)
- BW = bandwidth of single-stage single-tuned circuit at 3 db.
- BW_n = bandwidth of staggered group or *n-uple* at 3 db.
- BW_{oa} = bandwidth of the over-all system. (m cascaded *n-uples*).

1 "Vacuum Tube Amplifiers" by Valley and Wallman, Radiation Laboratory Series, Volume 18, McGraw Hill 1948; "Theory and Design of T.V. Receivers," by S. Deutsch, McGraw Hill.

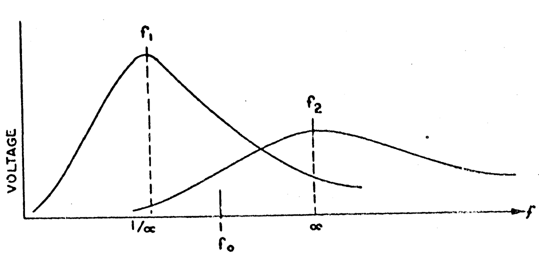

Fig. 14-17 Response of a staggered pair; geometric symmetry.

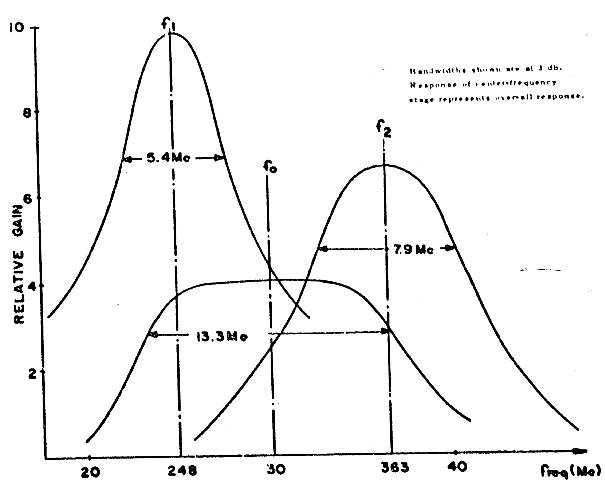

Fig. 14-18 Response of a staggered triple; geometric symmetry.

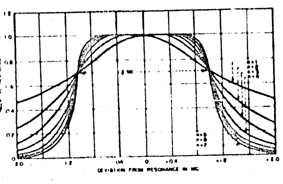

Fig. 14-19 Frequency responses of flat-topped bandpass networks.

The bandwidth shrinkage factor shown in Fig. 14-19 indicates the shrinking in bandwidth of any u-stage single-tuned amplifier in comparison with a single-stage single-tuned circuit at equal magnitude of gain. It also serves as a relative figure comparing *ABW* product of different n-stage systems.

Figs. 14-21a and 14-21b show the shrinkage factors for m-stage cascaded flat-staggered pairs and triples respectively. A revealing comparison of bandwidth shrinkage can be made between these staggered systems and a cascaded-synchronous system by comparing the tables shown in Fig. 14-21 with the table of Fig. 14-16. This comparison shows that cascaded-staggered stages do not give the degree of objectionable bandwidth shrinkage that is characteristic of synchronous single-tuned systems.

The use of Fig. 14-20 is illustrated as follows: Dividing the shrinkage factor for m-staggered pairs by the shrinkage factor for the u-stage cascaded-synchronous amplifier, we get the following ratio,

$$\frac{\sqrt[4]{2^{2/u}-1}}{\sqrt{2^{1/u}-1}}$$

Assuming the total number of single-tuned stages is six (u = 6) for both the m-staggered pairs and the cascaded-synchronous amplifier, the above ratio becomes

$$\frac{\sqrt[4]{2^{1/3}-1}}{\sqrt{2^{1/6}-1}} = \frac{\sqrt[4]{1.26-1}}{\sqrt{1.12-1}} = \frac{\sqrt[4]{.26}}{\sqrt{.12}} = \frac{.715}{.346} = 2.06$$

Or, from Table (a) of Fig. 14-21 for m = 3 we get .71 and from Fig. 14-16 for u = 6 we get .35. This ratio is

$$\frac{.71}{.35} = 2.03$$

or for all practical purposes the same as obtained by calculation.

This means that a six-stage amplifier in the form of three flat-staggered pairs has about twice the *ABW* product of a six-stage cascaded-synchronous amplifier. Hence, for the same over-all gain, the six-stage flat-staggered-pair amplifier has twice the over-all bandwidth.

For the development of the design equations of flat-staggered *n-uples* refer to Valley and Wallman. Only the final equations for staggered pairs and triples are stated below:

Let

$$\delta = \frac{BW_n}{f_o} \qquad \text{and} \qquad d = \frac{1}{Q} = \text{dissipation factor}$$

Staggered Pair:

$$d^2 = \frac{4 + \delta^2 - \sqrt{16 + \delta^4}}{2}$$

$$\delta^2 = d^2 + (\alpha - \frac{1}{\alpha})^2$$

for the frequency above f_o: $\quad BW = d\,(f_o \alpha) \qquad (47)$

for the frequency below f_o: $\quad BW = d\,(\frac{f_o}{\alpha}) \qquad (48)$

A typical design uses two stages, one staggered at $f_o\alpha$ and the other at f_o/α, each of dissipation factor d.

Staggered triple:

$$d^2 = \frac{4 + \delta^2 - \sqrt{16 + 4\delta^2 + \delta^4}}{2}$$

$$\delta^2 = d^2 + (\alpha - \frac{1}{\alpha})^2$$

AMPLIFIER CIRCUIT	OVERALL GAIN	OVERALL BANDWIDTH	GAIN BANDWIDTH PRODUCT	BANDWIDTH SHRINKAGE FACTOR
SINGLE-TUNED SINGLE-STAGE	A	BW	ABW	1
CASCADED SYNCH (u stages)	A_{oa} or A^u	BW_{oa}	$A_{oa}^{1/u} BW \sqrt{2^{1/u}-1}$ (Fig. 39)	$\sqrt{2^{1/u}-1}$ (See Fig. 14-15)
ONE STAGGERED PAIR	A_n or A_2	BW_n or BW_2	$A_2^{1/2} BW_2$	1
m STAGGERED PAIRS	A_{oa} or A_2^m	BW_{oa}	$A_{oa}^{1/2m} BW_2 \sqrt[4]{2^{1/m}-1}$	(See Fig. 14-20) $\sqrt[4]{2^{1/m}-1}$ or $\sqrt[4]{2^{2/u}-1}$
ONE STAGGERED TRIPLE	A_n or A_3	BW_n or BW_3	$A_3^{1/3} BW_3$	1
m STAGGERED TRIPLES	A_{oa} or A_3^m	BW_{oa}	$A_{oa}^{1/3m} BW_3 \sqrt[6]{2^{1/m}-1}$	$\sqrt[6]{2^{1/m}-1}$ or $\sqrt[6]{2^{3/u}-1}$ (See Fig. 14-20)

1. BW_2 and BW_3 here denote the 3 db bandwidth of a staggered pair and staggered triple, respectively.
2. For any flat-staggered *n-uple*, the *ABW* product is the same as that of a single-stage single-tuned circuit in that *n-uple*, i.e., *ABW* = ABW.
3. The response of the *n-uple* is given by

$$\text{Relative gain} = \frac{1}{\sqrt{\delta^2 + (f - \frac{1}{f})^{2n}}} \qquad \text{Where } \delta = \frac{BW_n}{f_o}$$

Fig. 14-20 Comparison of single-tuned single-stage, cascaded synchronous, and flat-topped staggered systems.

for the frequency above f_0: $BW = d\,(f_0 \alpha)$ (49)

for the center frequency f_0: $BW = d\, f_0$ (50)

for the frequency below f_0: $BW = d\,(\frac{f_0}{\alpha})$ (51)

A typical design uses two stages, one staggered at $f_0\alpha$ and the other at f_0/α, each of dissipation factor d, and a third stage centered at f_0 of bandwidth BW_3, (the over-all bandwidth).

For ready reference the design equations for d and α for flat-staggered pairs are presented graphically in Fig. 14-22 and for flat-staggered triples are presented in Fig. 14-23. These two graphs will be found convenient for designing staggered pairs or staggered triples.

2. Assume a flat-staggered triple of 20.6 Mc bandwidth is to be designed with band center at 14.3 Mc. Then f_0 = 14.3 Mc, BW = 20.6Mc, and $\delta = BW_n/f_0$ = 1.44, so that from Fig. 14-23, α = 1.84 and d = 0.60.
Therefore, the triple is to be constructed from:
One stage staggered at $f_0\alpha$ = 14.3(1.84) = 26.3 Mc of dissipation factor 0.60 and hence of bandwidth $d(f_0\alpha)$ = 0.60 (26.3) = 15.8 Mc (by Eq. 49).
Another stage staggered at f_0/α = 14.3/1.84 = 7.8Mc, of dissipation factor 0.60 and hence of bandwidth $d(f_0/\alpha)$ = 0.60(7.8) = 4.7Mc (by Eq. 51). A third stage centered at 14.3 Mc of bandwidth 20.6 Mc (by Eq. 50). From these data the Q's of the coils (Q = 1/d) can be readily calculated.

m	$\sqrt[4]{2^{1/m}-1}$
1	1.00
2	0.80
3	0.71
4	0.66
5	0.62

(a) m-flat-staggered pairs

m	$\sqrt[6]{2^{1/m}-1}$
1	1.00
2	0.86
3	0.80
4	0.76
5	0.73

(b) m-flat-staggered triples

Fig. 14-21 Shrinkage factors of flat-staggered pairs and triples.

Examples:

1. Assume a flat-staggered pair of 8 Mc bandwidth is to be designed with band center at 10 Mc. Then f_0 = 10 Mc, BW_n = 8 Mc, and $\delta = BW_n/f_0$ = 8/10 = .8, so that from Fig. 14-22 one finds that α = 1.33 and d = 0.535. Therefore, the pair is to be constructed with one stage staggered at $f_0\alpha$ = 10 (1.33) = 13.3 Mc of dissipation factor 0.535 and hence of bandwidth $d(f_0\alpha)$ = 0.535 (13.3) = 7.1 Mc (by Eq. 47).
The other stage is staggered at f_0/α = 10/1.33 = 7.5 Mc of dissipation factor 0.535 and hence of bandwidth $d(f_0/\alpha)$ = 0.535 (7.5) = 4.0 Mc (by Eq. 48).
From these data the Q's of the coils (Q = 1/d) can be readily calculated.

Design Procedures for Stagger-tuned Circuits.

The design of an amplifier made up of staggered single-tuned circuits involves the following steps:

1. Determination of the best form (type of *n-uple*) that will provide the required ABW.
2. Selection of a reasonable value of C.
3. Determination of the number of stages needed to provide the specified gain.
4. Selection of proper tube based on consideration of ABW product and selected value of C.
5. Design of the selected *n-uple* (based on Figs. 14-22, or 14-23).
6. Calculation of the required circuit Q's of the individual stages and load resistors, if required.

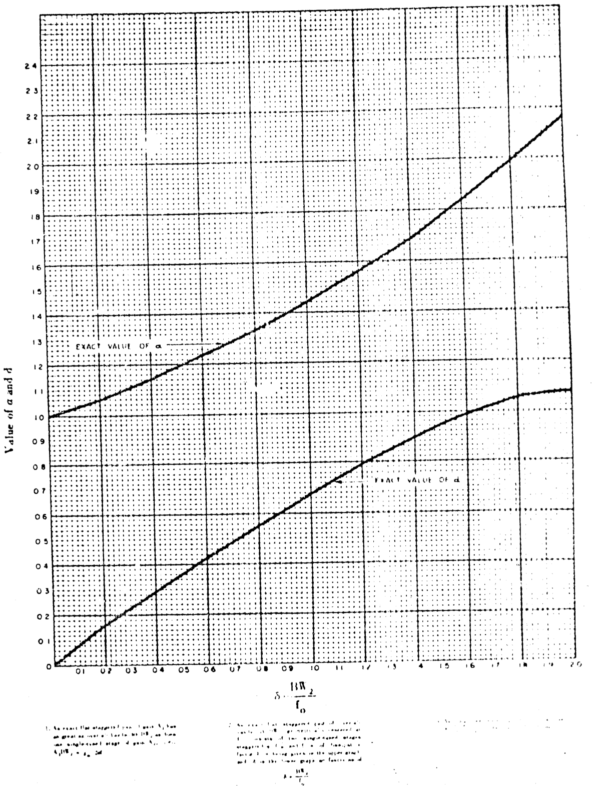

Fig. 14-22 Design curves for an exact flat staggered pair.

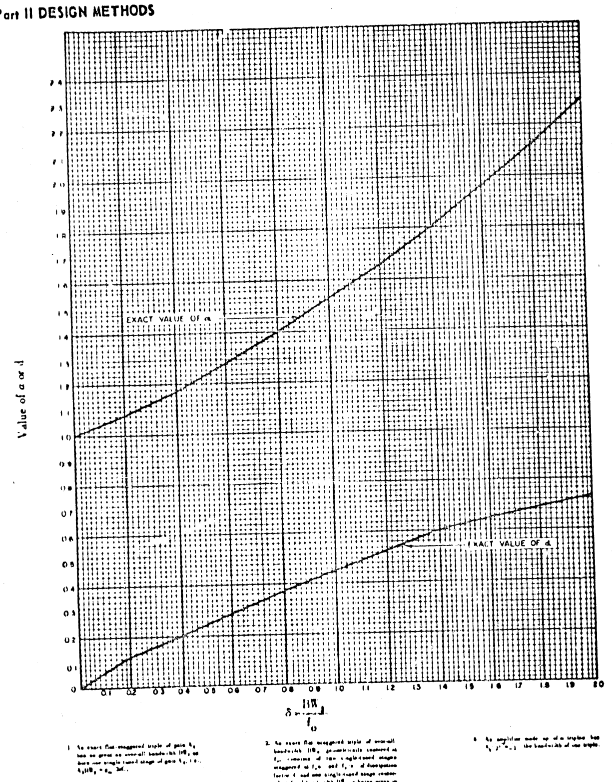

Fig. 14-23 Design curves for an exact flat staggered triple.

Example:

Required: An amplifier covering the band of 8.25 to 21.75 Mc and having a gain of 80 db.

Solution: From the requirements given above we know that the over-all bandwidth of the system is equal to 21.75 - 8.25 or 16.5 Mc we also know that the band center of a stagger-tuned amplifier is equal to the geometric mean of the band extremities. Therefore, in this case we find f_0 to be equal to $\sqrt{8.25(21.75)}$ or 14.3 Mc. (See Eq. 46).

The next step is to determine the number of stages that will be needed to provide the desired gain and bandwidth. Remembering the advantage in ABW of a flat-staggered triple over a flat-staggered pair, Fig. 14-21b is checked to see what bandwidth is required if flat-staggered triples are used. Since a gain of 80 db is specified, it seems safe to assume that at least three flat-staggered triples – a total of nine stages – will be required. From Fig. 14-21b we see that three flat-staggered triples will shrink to 80 per cent of the single stage bandwidth. The bandwidth of each triple then becomes 20.6 Mc in order to insure an over-all bandwidth of 16.5 Mc.

The requirement of a gain of 80 db means that in a nine stage amplifier each stage must have a gain of 8.9 db or 2.78 times. From this information, it becomes apparent that the ABW for each triple will be 2.78 times 20.6 or 57.3 Mc.

At this point it becomes necessary to select a value for C which, it will be recalled, is made up of the input and output capacitances of the tubes plus the circuit and distributed capacitances and whatever fixed capacitance is built into the unit. Taking into consideration the average values of interelectrode capacitances in the tubes likely to be used, and adding a desire for as much stability as is possible, the value of 25 μμf seems a satisfactory compromise.

Now since

$$ABW = \frac{g_m}{2\pi C}$$

we have only to substitute in this equation the values which have already been determined and solve for the value of g_m in order to select the tube to be used. We then have

$$g_m = 57.3(10^6)2\pi(25)10^{-12} = 9000 \ \mu\text{mhos}$$

A check of tube manuals indicates that this can be met by a 6AC7 thus insuring the required gain and bandwidth. A further check of the characteristics of a 6AC7 show the input capacitance to be 11 μμf and the output capacitance 5 μμf making a total of 16 μμf in the tube thus leaving 25 - 16 = 9 μμf for circuit and distributed capacitances – a very reasonable value for such an application.

We have now decided upon the make-up of the amplifier, selected the value of C, and have picked a tube which satisfies the requirements. The one remaining task is to design the flat-staggered triples.

From the information at hand we know that the bandwidth of each triple is 20.6 Mc, f_0 = 14.3 Mc, and δ must, therefore, be 20.6/14.3 = 1.44. Turning now to Fig. 14-23, we see that, under these circumstances, α = 1.84 and d = 0.60. As a result, each flat-staggered triple must consist of:

One stage staggered at 14.3 (1.84) = 26.3 Mc, of dissipation factor 0.60 and hence of bandwidth 26.3 (0.60) = 15.8 Mc (BW = df_0 for single stage).

One stage staggered at 14.3/1.84 = 7.8 Mc of dissipation factor 0.60 and hence of bandwidth 7.8 (0.60) = 4.7 Mc.

One stage centered at 14.3 Mc with BW = 20.6 Mc (d = 1.44).

It should be apparent from the values of d, (d = 1/Q), that the circuit Q's will be very low. The only practical way of obtaining such low Q is to load the coils with appropriate values of resistance.

The one remaining step in the design then is to calculate these resistors which is most easily done from the relationship:

$$\text{Stage bandwidth} = \frac{f_0}{Q} = \frac{1}{2\pi R_p C} \qquad \text{from which}$$

$$R_p = \frac{1}{2\pi (BW) C}$$

For the first stage of the triplet we now have

$$R_p = \frac{10^{12}}{2\pi(15.8)10^6(25)} = 403 \ \text{ohms}$$

The other stages will be found to be 1360 and 309 ohms respectively. (In practice, these resistors would most likely have the standard values of 390, 1500, and 330 ohms.)

The order in which the tuned circuits appear in the amplifier is not particularly important except that best results are usually obtained when the first stage and the stage that feeds the detector are centered. See Fig. 14-24 for schematic of single-stage and block diagram of 9 stage system.

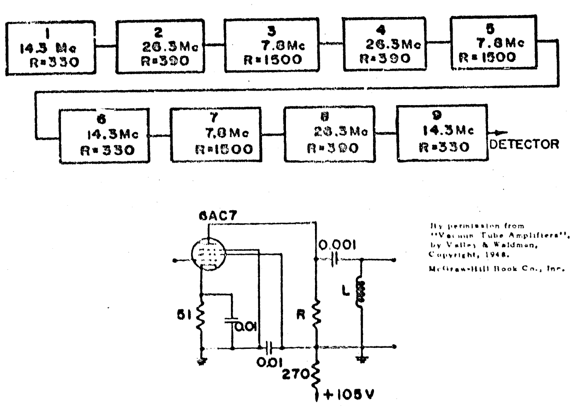

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Fig. 14-24 Block diagram of 9 stage system and single stage schematic.

TRANSFORMER COUPLING:

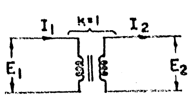
Transformers, the second major classification of Inductive Coupling Devices, includes all networks involving mutual inductive coupling between meshes. Within this group are tapped single-windings (auto-transformers) and multi-winding transformers in which there is no metallic connection between the meshes.

The basic principles governing the physical device known as the transformer were discovered in 1838 by the early American Physicist, Joseph Henry. It was not until the advent of a-c distribution systems around the beginning of the present century that this device reached the stage of efficient design. From this development came the concept of an ideal transformer as a device which multiplies the voltage by the turns-ratio and the current by the reciprocal of this ratio (in effect changing the impedance of an a-c source) and does all of this with no power loss and no magnetizing current.

For all practical purposes, the output of a transformer may be considered as the output of an a-c generator. When an a-c generator is connected to a circuit containing both resistance and reactance, maximum power transfer will take place only when the generator and the load have conjugate impedances, i.e., $Z_g = R + jX$ and $Z_L = R - jX$, or if magnitude only, but not angle, can be matched, then the generator and load impedances must be equal. This is to say that from a power point of view, conjugate (or exact in magnitude) impedance matching represents optimum design.[1]

Transformers with windings completely linked by a common flux (unity-coupled) operate on the principle of turns ratio and make ideal impedance matching devices. A common approach to unity-coupling at radio frequencies is the bifilar single-tuned transformers used as transistor coupling devices and bifilar transformers used in television video amplifiers.

Figure 14-25 shows a basic unity-coupled transformer and the following theoretical analysis is presented to illustrate the turns-ratio principle and to aid in the basic understanding of the fundamental concepts of impedance transformation.



N_1 = number of primary turns

N_2 = number of secondary turns

k = coefficient of coupling

Fig. 14-25 Basic unity-coupled transformer.

$$\text{Turns ratio} = n = \frac{N_2}{N_1}$$

$$\text{and } E_2 = nE_1 \qquad (52)$$

$$\text{Also } I_2 = \frac{I_1}{n} \qquad (53)$$

$$\text{and } \frac{E_2}{I_2} = n^2\left(\frac{E_1}{I_1}\right) \qquad (54)$$

$$\text{But since } \frac{E_2}{I_2} = Z_2 \text{ and } \frac{E_1}{I_1} = Z_1$$

$$\text{It follows that } Z_2 = n^2 Z_1 \qquad (55)$$

With these formulas it is possible to calculate with a high degree of accuracy certain performance characteristics of a unity-coupled transformer. For example, assume a unity-coupled transformer with $N_2 = 2N_1$ and a primary voltage of 100 and primary current of 2 amperes under load. Then the secondary voltage, E_2 will be 200 and the secondary current, I_2 will be 1 ampere.

Transformers may be classified into two major groups according to the number of windings. Those having only one winding tapped for purposes of impedance matching, are known as auto transformers. The more conventional type of transformer contains two or more complete and separate windings insulated one from the other. In r-f applications, transformers most frequently have two windings whose coupling is set by design to a specific value.

Of these two major types, the simplest and cheapest to manufacture is the auto transformer. According to the location of the tap and the connections to input and output circuits, auto transformers may be step-down (Fig. 14-26a) or step-up (Fig. 14-26b). In either case, if the coupling is unity (100%) the simple turns ratio principle can be applied; if coupling is not unity, as is usually the case for r-f where coupling usually ranges from 1% to 5%, more involved design considerations must be invoked to effect the desired impedance transformations.

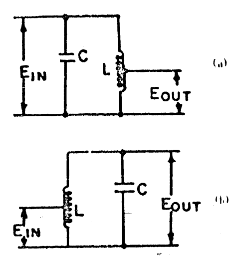


Fig. 14-26 Auto transformer, (a) step-down and (b) step-up.

Before proceeding with the detailed discussion of r-f transformer design the principal functions of an r-f transformer may be summarized as any or all of the following:

1. Provide a means of coupling between circuit elements.
2. Provide a means of isolating the plate voltage of one tube from the grid bias of the following tube.
3. Provide a means of impedance matching between various circuit elements.
4. To permit polarity inversion. In this category are transformers used for pushpull operation, full-wave detectors, ratio detec-

[1] It should be noted, however, that optimum power matching is not necessarily the best condition for the highest signal to noise ratio. In certain instances, by deviating from conjugate impedance matching, a feed-back cancellation can be effected which improves the signal-to-noise ratio. This situation is, however, of importance only in amplifiers of extremely low signal levels, and since its complete treatment would involve an elaborate treatise of the noise concept, it is considered to be outside the scope of this manual.

tors, discriminators, and similar circuit devices.

Inductively Coupled Circuits, Theory:

Circuits are said to be inductively coupled when mutual inductance exists between coils that are in different circuits. It can be shown by simple analysis that this mutual inductance is related to the primary and secondary inductances as follows:

$$M = k\sqrt{L_1 L_2}$$

where L_1 = primary inductance
L_2 = secondary inductance
k = coefficient of coupling

The mutual inductance makes possible the transfer of energy by transformer action from one circuit to the other. Examples of typical inductively-coupled circuits commonly encountered in electronic work are shown in Fig. 14-14c through f, inclusive.

The behavior of inductively coupled circuits is somewhat complex, but it can be readily calculated with the aid of the following three rules:

1. *The secondary has the same effect upon the primary circuit as though an impedance $(\omega M)^2/Z_2$ had been added in series with the primary.* This can be shown as follows (Fig. 14-27), where

M = mutual inductance; $\omega = 2\pi f$;

$Z_2 = (R_2 + j\omega L_2)$ = series impedance

of secondary circuit when considered by itself. The equivalent impedance $(\omega M)^2/Z_2$ which the presence of the secondary adds to the primary circuit is known as the *coupled impedance*. Since Z_2 is a vector quantity having both magnitude and phase, the coupled impedance is also a vector quantity having resistance and reactance components.

$$E = I_1 Z_1 + j\omega M I_2$$

Induced voltage in SEC $= -j\omega M I_1 = I_2 Z_2$

Solving to eliminate I_2, where Z_1 and Z_2 = primary and secondary impedances, respectively,

$$E = I_1\left[Z_1 + \frac{(\omega M)^2}{Z_2}\right] \qquad (56)$$

This relation shows that the effective primary impedance with the secondary present is $Z_1 + (\omega M)^2/Z_2$, of which the second term represents the coupled impedance due to the presence of the secondary.

2. *The voltage induced in the secondary circuit by a primary current of I_1 has a magnitude of $\omega M I_1$ and lags the current that produces it by 90°. In complex quantity notation the induced voltage is $-j\omega M I_1$.*

3. *The secondary current is exactly the same current that would flow if the induced voltage were applied in series with the secondary and if the primary were absent. (The secondary current therefore has a magnitude $\omega M I_1/Z_2$, and in complex quantity representation is given by $-j\omega M I_1/Z_2$.*

These relationships hold for all frequencies and all types of primary and secondary circuits, both untuned and tuned. The following procedure is recommended for computing the behavior of a coupled circuit.

(a) Determine the primary current with the aid of Rule 1.
(b) Compute the voltage induced in the secondary, knowing the primary current by using Rule 2.
(c) Calculate the secondary current from the induced voltage by means of Rule 3.

The following equations will aid the systematic execution of the aforementioned procedure.

$$\text{Impedance coupled into primary circuit by presence of the secondary} = \frac{(\omega M)^2}{Z_2} \qquad (57)$$

$$\text{Equivalent primary impedance} = Z_1 + \frac{(\omega M)^2}{Z_2} \qquad (58)$$

$$\text{Primary current} = I_1 = \frac{E}{Z_1 + \frac{(\omega M)^2}{Z_2}} \qquad (59)$$

$$\text{Voltage induced in secondary} = j\omega M I_1 \qquad (60)$$

$$\text{Secondary current} = \frac{-j\omega M I_1}{Z_2} = \frac{-j\omega M E}{Z_1 Z_2 + (\omega M)^2} \qquad (61)$$

Many important properties of coupled circuits can be determined by examining the nature of the coupled impedance $(\omega M)^2/Z_2$. When the mutual inductance M is very small, or if the secondary impedance Z_2 is large, the impedance coupled into the primary by the presence of the secondary is small. In either case the induced secondary current is small and little energy transfer takes place, and the primary current is nearly the same as though no secondary were present.

If the secondary impedance Z_2 is small and the mutual inductance is not too small, the coupled impedance $(\omega M)^2/Z_2$ is large and the voltage and current relations in the primary circuit are affected to a considerable extent by the presence of the coupled secondary.

It is important to remember that the coupled impedance has the same phase angle as the secondary impedance Z_2, with the exception that the sign of the angle is reversed. This means that an inductive secondary having an angle of 30° lagging couples into the primary circuit an impedance of 30° leading. This change from lagging to leading is equivalent to neutralizing some of the inductive reactance already possessed by the primary, and this is done electrically by postulating a capacitive reactance of suitable magnitude in series with the inductance to be neutralized (Fig. 14-27b). It should not be assumed that a resultant capacitive reactance can be obtained in the primary circuit by very large coupling, since with the maximum coupling that can exist (k=1), the coupled capacitive reactance can never exceed the value that will just neutralize the inductive reactance of the primary. When the secondary impedance Z_2 is a pure resistance, the coupled impedance will also be a resistance.

A particularly important case of coupled impedance occurs when the secondary is a resonant circuit. Then

$$\text{Coupled impedance} = \frac{(\omega M)^2}{Z_2} = \frac{(\omega M)^2}{R_2 + j\left(\omega L_2 - \frac{1}{\omega C_2}\right)} \qquad (62)$$

The coupled impedance produced by a tuned secondary circuit consequently varies with frequency according to the same general law as does the parallel impedance of the secondary circuit. At resonance the effect is greatest and is equivalent, insofar as the primary circuit is concerned, to introducing a resistance in series with the primary.

The theory just presented is the general theory of transformers, applicable to all circumstances irrespective of the degree of coupling. The commonly used method of analyzing the power transformer, involving the concept of leakage inductances and turns ratio, is a special case of the general theory which is convenient when the coefficient of coupling k approaches unity, as is the case when closed magnetic cores are used. The concept of "leakage inductance" represents those portions of the primary and secondary inductance which are not linked (coupled) by the mutual flux. Fig. 14-28 shows these leakage inductances to be $(1-k)L_1$ and $(1-k)L_2$, where k is the coefficient of coupling which varies between 0 and 1, and has the value

$$k = \frac{M}{\sqrt{L_1 L_2}} \qquad (63)$$

In most r-f applications where the coefficient of coupling is low, the voltage induced in the secondary winding will bear little relationship to the turns ratio. This arises from the fact that when the coefficient of coupling is low, as for example, 0.01, then the primary and secondary inductances (Fig. 14-28) are practically entirely leakage inductances.

Untuned Transformers

It was previously explained that there is actually no such thing as an untuned coil; this also applies to transformers. However, the arrangement shown in Fig. 14-29 is commonly considered as an untuned transformer since there is no means provided for tuning either the primary or secondary.

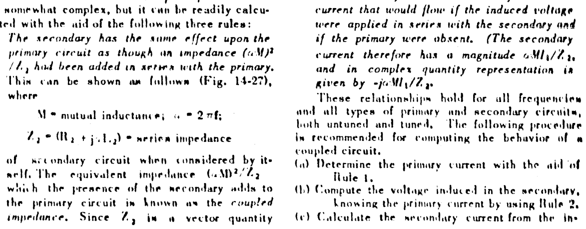


Fig. 14-27. Transformer circuit.

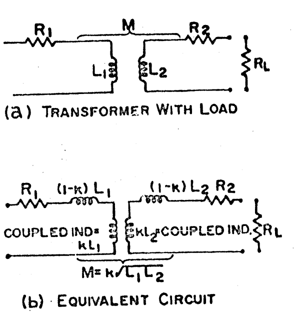


Fig. 14-28 Coupled circuits illustrating leakage inductance.

This application can be represented as a coil L_1 (primary) located near a mass of metal represented by the secondary L_2. This mass of metal can be a shield, a metal panel or any other metal object that is in the field of the coil and has an influence upon its operation.

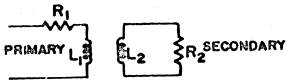


Fig. 14-29 Untuned transformer.

The secondary is essentially an inductance in series with a resistance and so has a lagging reactance. As a result, the secondary couples into the primary a resistance component and also a capacitive component. The resulting effect is to increase the effective primary resistance and to neutralize a portion of the primary inductance. With perfect coupling ($k = 1.0$) and zero secondary resistance, the primary inductance would be completely neutralized, a condition difficult to approach and impossible to attain; while with lesser degrees of coupling the effect of the secondary upon the primary inductance is correspondingly less.

When the secondary is a metal object of low resistance material, the impedance is largely reactive and the resistance component of the coupled impedance is small, primarily resulting in the reduction of primary inductance due to the coupled reactance. If a coil is to be shielded or otherwise located near a metal object, the metal used should be the best possible conductor, preferably copper or aluminum, so that the added losses will be small (see article, Electromagnetic Shielding, Section 2, page 1).

If the secondary resistance can be assumed to be zero, the coupled impedance is a capacitive reactance having no resistance component, and the equivalent primary inductance is

$$\text{Equivalent Primary Impedance} = j\omega L_1 + \frac{(\omega M)^2}{j\omega L_2}$$

$$= j\omega L_1 - \frac{j\omega M^2 L_1}{L_1 L_2}$$

$$= j\omega L_1 \left(1 - \frac{M^2}{L_1 L_2}\right)$$

$$\text{Since } k^2 = \frac{M^2}{L_1 L_2} \qquad (64)$$

then the equivalent primary inductance $= L_1(1-k^2)$

where L_1 is the primary inductance with the secondary removed and k is the coefficient of coupling between the coil and the secondary. In order to determine the reduction in effective inductance produced by a shield or other metal object, one needs to know only the coefficient of coupling.

The variation of effective inductance of a coil due to the proximity of a low-loss metal mass as expressed in equation 64 is the basic principle for *non-magnetic core tuning*, discussed on page 3-2.

Single-Tuned Transformers

The *single-tuned transformer* can be either of the secondary-tuned type as shown in Fig. 14-30a or

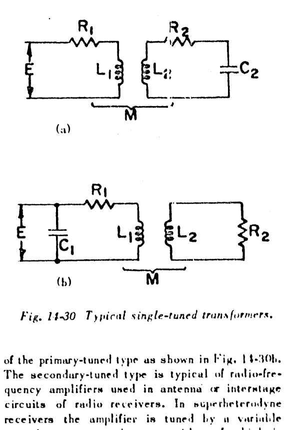


Fig. 14-30 Typical single-tuned transformers.

of the primary-tuned type as shown in Fig. 14-30b. The secondary-tuned type is typical of radio-frequency amplifiers used in antenna or interstage circuits of radio receivers. In superheterodyne receivers the amplifier is tuned by a variable capacitor or magnetic core, either of which is ganged with a similar unit in the oscillator circuit.

The primary-tuned type, Fig. 14-30b, is typical of the close-coupled single-tuned transistor i-f transformer. Either type is also used as single-tuned i-f transformers in vacuum tube circuitry, generally in the frequency range of 250 to 500 kc.

A typical tuned r-f amplifier stage is shown in Fig. 14-31a. The value of the secondary inductance L_2 is dependent upon the frequency band to be covered and the capacity range of the tuning capacitor C_2. The secondary inductance for a frequency range of 540 to 1620 kc is ordinarily between 200 and 300 μh.

The first two rules for inductive coupling stated on page 14-38 can be readily applied to the analysis of this amplifier performance. First, the value of R_1, the series loss resistance of the primary is so small in comparison to the plate resistance of a pentode that it can be ignored.

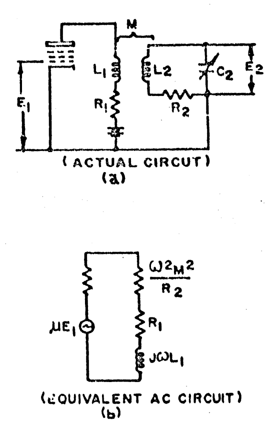


Fig. 14-31 Single-tuned transformer-coupled amplifier.

The same is true of the relatively low impedance ($j\omega L_1$) of the untuned primary (Fig. 14-31b).

By rule 1: The reflected impedance of the tuned secondary ($Z_2 = R_2$) is $\omega^2 M^2/R_2$ in series with the primary.

By rule 2: The primary current is determined almost completely by the plate resistance r_p of the pentode. Thus $I_p = \mu E_1/r_p = g_m E_1$

The induced voltage in the tuned secondary is $I_p \omega M = g_m E_1 \omega M$.

By rule 3:[1] This induced voltage is applied in series with the secondary closed circuit; since the tuned secondary impedance is only R_2, then $I_2 = g_m E_1 \omega M/R_2$. The voltage across the coil (or capacitor, since $X_L = X_C$) then is $E_2 = g_m E_1 \omega M (\omega L/R_2)$

Since $\omega L/R_2 = Q_2$ then: $E_2 = (g_m E_1 \omega M) Q_2$

$$\text{or} \quad A = \frac{E_2}{E_1} = g_m \omega M Q_2 \qquad (65)$$

[1] Stated another way, Rule 3 shows that the induced voltage is multiplied Q times across the coil and capacitor.

DESIGN PROCEDURES

From a practical standpoint, the starting point in the design of a single-tuned transformer is the determination of the inductance of the tuned circuit required to cover the desired band of frequencies. Regardless of whether the transformer is designed to operate with capacitive or inductive tuning, the controlling factor in the design will be the minimum capacitance value.

If we let f_h represent the highest frequency which the transformer is to cover and f_1 the lowest frequency, we can then say that

$$f_h = \frac{1}{2\pi\sqrt{L_{min}C_{min}}} \quad \text{and}$$

$$f_1 = \frac{1}{2\pi\sqrt{L_{max}C_{max}}}$$

Now it is equally true that

$$\frac{f_h}{f_1} = \sqrt{\frac{L_{max}C_{max}}{L_{min}C_{min}}}$$

so if the inductance is fixed as in the case of capacitive tuning, we have

$$\frac{f_h}{f_1} = \sqrt{\frac{C_{max}}{C_{min}}} \qquad (66)$$

and if the capacitance is fixed as in permeability-tuned units,

$$\frac{f_h}{f_1} = \sqrt{\frac{L_{max}}{L_{min}}} \qquad (67)$$

If we consider for a moment those cases where tuning is capacitive, it will be seen that if the minimum capacitance is computed from the sum of the tube and circuit capacitances plus the minimum capacitance of the tuning capacitor, that for a given range the maximum required capacitance is

$$C_{max} = C_{min}\left(\frac{f_h}{f_1}\right)^2 \qquad (68)$$

The complete design procedure is as follows:

(a) Determine the maximum C by Eq. 68.
(b) Using the value of C determined in (a), calculate the secondary inductance required for the lowest specified frequency.
(c) Compute the "MQ" product by Eq. 65 (use approximate band center).
(d) Calculate mutual inductance from "MQ" product. (It is necessary to assume a value of Q in order to solve for M. This can either be based upon experience or reference to published information. In lieu of either, a secondary Q of between 50 and 150 may be considered reasonable in the frequency range of 500 to 1700 kc). If low cost is important use the lower value of Q, if narrow bandwidth is necessary use the higher value of Q.
(e) Calculate the primary inductance by Eq.10.

$$L_p = \frac{1}{4\pi^2 f^2 C}$$

For highest gain in the frequency band to be covered it is customary to resonate the primary (with circuit and distributed capacitance) at a frequency just below the lowest frequency to be covered by the secondary, but in super-heterodyne receiver applications primary resonance should be kept as far as possible from the i-f frequency.

Example:

Required, an r-f transformer having an untuned primary and a tuned secondary to operate with a 1T4 (pentode) tube. The amplifier is to cover the frequency range of 540 to 1620 kc with a suitable variable capacitor having a minimum, plus circuit, tube and distributed capacity of 35 μμf. Assume a required gain of 15 (23.5 db).
Solution:

g_m of 1T4 tube is 700 μ mhos

By equation 68:

(1) $C_{max} = 35\left(\frac{1620}{540}\right)^2 = 315\ \mu\mu f$

(2) Inductance of secondary winding required to resonate at 540kc with 315 μμf:

$$L_2 = \frac{1}{4\pi^2 f^2 C} = \frac{1}{4(9.87)(.54)^2 10^{12}(315)10^{-12}}$$

$$= 276\ \mu h$$

(3) Compute "MQ_2" product by equation 65 (at 1000 kc):

$$MQ_2 = \frac{A}{g_m\omega} = \frac{15}{700(10^{-6})6.28(10^6)} = 3.42(10^{-3})$$

Note: At this point a reasonable value must be assumed for Q_2 in order to solve for M. (It is recommended, in lieu of experience, that a secondary Q of 50 to 150 be accepted as reasonable for this frequency range). For this example assume that a Litz-wire winding of 3/42 having an inductance of 276 μh has a Q of 80 at 1000 kc, (approximately mid-band).

$$(4)\quad M = \frac{3.42(10^{-3})}{80} = 42.7\ \mu h$$

(5) For highest gain in the frequency band to be covered the primary resonance should be just below the lowest frequency, in this case choose 530 kc and assume tube, circuit and distributed capacity to be 15 μμf. Then:

$$L_p = \frac{1}{4\pi^2 f^2 C} = \frac{1}{39.5(.53)^2 10^{12}(15)10^{-12}}$$

$$= 6.0\ mh.$$

The circuit specification therefore becomes:

$$L_s = 276\ \mu h$$
$$Q_s = 80$$
$$L_p = 6.0\ mh$$
$$M = 42.7\ \mu h$$

The circuit Q (secondary) can be converted to coil Q allowing for tube loading by Eq. 79.

Double Tuned Transformers

The double-tuned system is probably the most popular circuit configuration for i-f systems operating in the frequency range of 250 kc to 50 Mc. The double-tuned transformer consists of two tuned circuits (the primary and secondary) each resonant to the required frequency and coupled inductively to a degree which will give a desired shape to the selectivity characteristic. When this coupling is adjusted to a certain optimum value called *critical coupling* (k_c), the response curve has the shape shown in Fig. 14-32a. If the coupling is increased beyond this critical value, the response takes on the double-humped appearance shown in Fig. 14-32b. If the coupling is below the critical value, the response is peaked and somewhat sharper

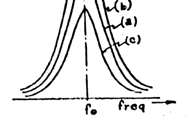


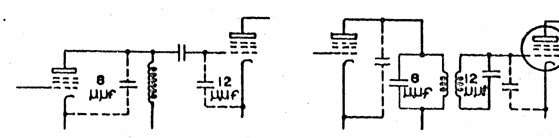
Fig. 14-32 Response curves of double-tuned circuits.

than for the critical value and the amplifier gain is reduced, Fig. 14-32c. It is the system designer's choice as to which of these response shapes would best satisfy the requirements, taking into account: complexity of the resultant circuits, the cost, the degree of tolerance that can be permitted in the non-uniformity of the peak response (degree of double-humping), and the ease of tuning during alignment. Illustrations of these factors will be made as the several examples are developed.

A significant advantage of double-tuned transformers in high-frequency application where tube and stray capacitance only are used for coil tuning is found in the increased gain-bandwidth product as compared to that of single-tuned amplifiers. This increase in ABW is due to the fact that the input and output capacitances of the tubes to be coupled are distributed between the primary and secondary of the transformer and are thus reduced in each tuned circuit. Since ABW is a function of $g_m/2\pi C$, it is apparent that in a case such as is illustrated in Fig. 14-33, the use of a double-tuned circuit can increase the gain-bandwidth product by a factor of approximately two.

In lower frequency applications where substantial fixed capacitance is added across each tuned circuit, the advantage of this capacitance reduction is not so significant; in these instances where large fixed C is used, the ABW product is actually lowered compared to the single-tuned case. This follows from the fact, as will be shown, that the gain in a double-tuned stage is only one-half that of the single-tuned stage gain while the bandwidth is increased only by $\sqrt{2}$; the ABW product in the case of large fixed capacity is therefore 0.707 of that of the single-tuned stage.

A comparison of 3 db bandwidth between multi-stage single and double-tuned amplifiers in



(a) Single-tuned-circuit coupled amplifier in which 20 µµf appears across the inductance

(b) The same capacitance values as in Fig. 14-22a showing how the effective capacitance across the inductance has been cut by about 2 to 1.

Fig. 14-33 Effect of external capacities on single and double-tuned circuits.

shown in Fig. 14-34 as compared to the 3 db bandwidth of a single stage. It is apparent that the bandwidth narrowing in a multi-stage single-tuned amplifier is much greater than in a multi-stage double-tuned amplifier.

Fig. 14-35 shows the actual circuit of a double-tuned amplifier stage; Fig. 14-36 shows the parameters of the transformer and Fig. 14-37 shows the equivalent circuit of the transformer as seen from the plate of the tube. The three basic rules for coupled circuits (page 38) can be easily applied to find the overall gain E_{out}/E_{in} of this stage:

By rule 1: Impedance of the resonant secondary is reflected into the primary as a pure resistance ω^2M^2/R_2. The current I_1 in the primary coil is determined almost entirely by the value of ωL_1 (that is, the total loss resistance in the primary coil can be ignored compared to ωL_1 since the Q is assumed to be greater than about 10). Thus

$$I_1 = \frac{E_1}{\omega L_1} \quad \text{(absolute value)}$$

By rule 2: The induced voltage in the secondary

$$I_1\omega M = \left(\frac{E_1}{\omega L_1}\right)\omega M$$

$$\text{or Induced Voltage} = \frac{E_1 M}{L_1} \quad \text{(absolute value)}$$

No. of Stages	Relative Values of 3db Bandwidth		Relative Values of BW_{A0}/BW_6	
	Single-tuned	Double-tuned[1]	Single-tuned	Double-tuned
1	1.00	1.00	577	23.9
2	.64	.80	33	5.65
3	.51.	.71	13	3.59
4	.44	.66	8.6	2.94
6	.35	.59	5.9	2.43
8	.30	.55	5.0	.
10	.27	.52	4.5	.

Fig. 14-34 Relative values of 3db bandwidth for single and double-tuned amplifiers.

[1] Based upon identical primary and secondary circuits critically coupled.

By rule 3: The secondary current equals the induced voltage divided by the net secondary impedance, which in this case is R_2.

$$I_2 = \frac{\frac{E_1 M}{L_1}}{R_2} = \frac{E_1 M}{L_1 R_2} \quad \text{(absolute value)}$$

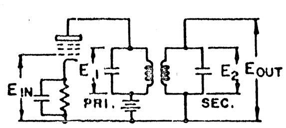


Fig. 14-35 Double-tuned transformer-coupled amplifier.

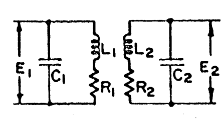


Fig. 14-36 Transformer Parameters.

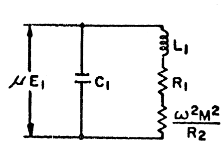


Fig. 14-37 Equivalent circuit as seen from plate of tube.

But the output voltage of the transformer

$$E_2 = I_2\omega L_2 = \frac{E_1 M\omega L_2}{L_1 R_2}$$

from which

$$\frac{E_1}{E_2} = \frac{L_1 R_2}{M\omega L_2} \qquad (67)$$

Now

$$\frac{E_1}{E_{in}} = g_m \omega L_1 Q_e$$

where Q_e is the effective Q of the equivalent circuit, that is

$$Q_e = \frac{\omega L_1}{\frac{\omega^2 M^2}{R_2} + R_1} = \frac{\text{Reactance}}{\text{Total loss Resistance}}$$

(note $M = k\sqrt{L_1 L_2}$)

Substituting and simplifying, it can be shown that

$$\frac{E_1}{E_{in}} = g_m \frac{\omega L_1}{Q_2\left(k^2 + \frac{1}{Q_1 Q_2}\right)} \qquad (68)$$

Dividing equations 68 by 67 yields the overall stage gain

$$A = \frac{E_2}{E_{in}} = g_m \frac{\omega_0 M}{k^2 + \frac{1}{Q_1 Q_2}} = g_m k \frac{\omega_0\sqrt{L_1 L_2}}{k^2 + \frac{1}{Q_1 Q_2}} \qquad (69)$$

It can be further shown by differentiating A of Eq. 69 with respect to k, that the maximum amplification is obtained when

$$k = \frac{1}{\sqrt{Q_1 Q_2}} = k_c \qquad (70)$$

This value of coupling is called *critical coupling* or sometimes referred to as *optimum coupling*, and is designated by k_c. Substituting this value of coupling into Eq. 69 yields the single-stage gain for a critically-coupled amplifier:

$$A_0 = g_m \frac{\omega_0\sqrt{L_1 L_2}\sqrt{Q_1 Q_2}}{2}$$

Since in most practical cases of tube circuit application, L_1 is made equal to L_2 and $Q_1 = Q_2$, then the above stage gain reduces to:

$$A_o = g_m \frac{\omega_o L Q}{2} = \frac{g_m R_p}{2} \qquad (71)$$

and $$k_c = \frac{1}{Q} \qquad (72)$$

Analysis of the selectivity characteristic shows that the 3 db bandwidth of the double-tuned amplifier is

$$\overline{BW} = \frac{f_o\sqrt{2}}{Q} \qquad (73)$$

Substituting Q of Eq. 73 into Eq. 71 yields

$$A_o = g_m \frac{\sqrt{2}}{4\pi(\overline{BW})C} \qquad (74)$$

Note particularily, by comparing Eq. 71 with Eq. 34 that the gain of a double-tuned stage is *half* that obtainable from a single-tuned stage with the same coil Q; and comparing Eq. 73 with Eq. 17 that the 3 db bandwidth is $\sqrt{2}$ times that of the single-tuned stage. This feature gives the double-tuned selectivity curve a "squarer" characteristic approaching the ideal rectangular shape.

While it is recognized that the above formulae are adequate for computation of the gain and response of an amplifier made up of double-tuned transformers, it is also recognized that design practices based upon these formulae are subject to, at least, two serious limitations:

1. The necessary calculations for the full selectivity characteristic are long, tedious, and somewhat involved and must often be repeated several times in the development of a new transformer.
2. For the selection of many important design parameters, or the feasibility of usage of calculated values in practical constructions, such as the tuning capacitances, the coupling, and the Q values, intelligent choice - preferably based upon experience - is still required.

Involved mathematical calculations are time consuming and for this reason emphasis is directed in this manual toward simplified methods for the design of critically-coupled and over-coupled transformers. The procedure for critically-coupled units is based upon a table and a few simple formulae already presented, while the method for over-coupled units is based upon a table and a nomograph.

A more advanced approach to circuit analysis and calculation which is more universal in nature and requires no supporting material and is known as the *Equivalent Lattice Method* will be discussed in that portion of this section which is devoted to networks.

It would be difficult, if not almost impossible, to overcome the second deficiency except by practical experience or by repetitious examples illustrating the many "mathematical designs" that are not practical because of manufacturing limitations, magnetic material inadequacies or for other reasons. It is hoped that the liberal use of illustrations in this section will provide some measure of background in this connection.

DESIGN OF DOUBLE-TUNED TRANSFORMERS - SIMPLIFIED METHODS

Optimum Coupling: The following design method[1] for optimum coupled circuits is based on the table of Fig. 14-39 and equations previously given for double-tuned circuits in general. This method greatly simplifies the calculations for the circuit parameters upon which the transformer specifications are based. It should be recognized that the specifications resulting from this method are *circuit* specifications and allowance must be made for the effects of tube loading and wiring capacitance upon the Q and resonating capacitances of the coils.

Fig. 14-38 represents a typical resonance curve where:

f_o = resonant frequency of all circuits

$\overline{BW}_T$ = bandwidth corresponding to attenuation "t" in same units as f_o

E_o = voltage at resonance

E_T = voltage at extremities of the desired bandwidth

T = attenuation, expressed as voltage ratio, E_o/E_T

t = attenuation expressed in db

[1] A method described by C. E. Dean of Hazeltine Service Corporation as an Electronics Reference Sheet published in Electronics Magazine, McGraw Hill Co.

Total number of circuits u	Bandwidth factor, $\underline{K} = \sqrt{2}\sqrt[u]{T^2-1}$												
	T=1.12 t=1db	T=1.26 t=2db	T=1.41 t=3db	T=2.0 t=6db	T=4.0 t=12db	T=7.9 t=17db	T=10. t=20db	T=20 t=26db	T=40 t=32db	T=70 t=37db	T=100 t=40db	T=1000 t=60db	T=10⁴ t=80db
2 (1 pair)	1.01	1.24	1.41	1.86	2.76	3.73	4.46	6.32	8.96	11.90	14.14	44.5	-
4 (2 pair)	.84	1.01	1.13	1.41	1.86	2.21	2.45	2.96	3.54	4.04	4.46	7.96	14.14
6 (3 pair)	.75	.90	1.01	1.24	1.57	1.41	1.95	2.26	[illegible]	[illegible]	3.02	4.46	[illegible]
8 (4 pair)	.70	.84	.93	1.13	1.41	1.60	1.72	1.93	2.15	2.32	2.45	3.32	4.46
10 (5 pair)	.66	.79	.88	1.06	1.31	1.47	1.57	1.74	1.91	2.06	2.15	2.78	3.54
12 (6 pair)	.63	.75	.84	1.01	1.24	1.38	1.47	1.57	[illegible]	[illegible]	[illegible]	2.45	3.02

Fig. 14-39 Bandwidth factor $\underline{K}$ for optimum-coupled double-tuned circuits.

n = total number of resonant circuits
$\underline{K}$ = bandwidth factor obtained from table of Fig. 14-39.
Q = circuit Q defined as

$$Q = \frac{\omega L}{R} = \frac{\underline{K} f_o}{BW_T}$$

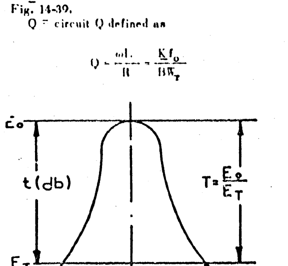


Fig. 14-38 Typical resonance curve showing voltages, frequencies and attenuations.

The equations which are the basis of the table are

$$BW_T = \frac{\underline{K} f_o}{Q} \qquad (75)$$

$$\underline{K} = \sqrt{2}\ \sqrt[4]{T^{4/n}-1} \qquad (76)$$

$$T = \sqrt[n/4]{1-\frac{\underline{K}^4}{4}} \qquad (77)$$

Since the values of Q determined by these methods are circuit, or loaded Q values (Q_L), it is necessary to convert them to unloaded Q values, (Q_u) so that they can be stated as transformer specifications.

Fig. 14-40 shows schematically the circuit conditions representing the loaded and unloaded Q values.

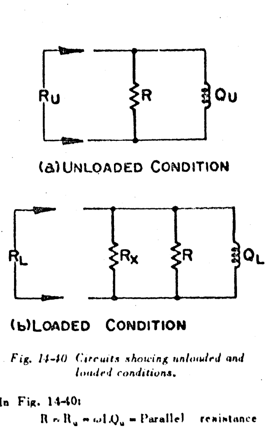


Fig. 14-40 Circuits showing unloaded and loaded conditions.

In Fig. 14-40:

$R = R_u = \omega L Q_u$ = Parallel resistance at required frequency.

R_x = Loading resistance added.

$R_L = \omega L Q_L$ = Parallel resistance of circuit with loading of R_x.

Q_u = Unloaded Q.

Q_L = Loaded Q

L = Inductance

Now:

$$R_L = \omega L Q_L = \frac{R_x R_u}{R_x + R_u} = \frac{R_x \omega L Q_u}{R_x + \omega L Q_u}$$

therefore:

$$R_x = \frac{\omega L Q_u Q_L}{Q_u - Q_L} \qquad (78)$$

and when Q_L and R_x are known:

$$Q_u = \frac{R_x Q_L}{R_x - \omega L Q_L} \qquad (79)$$

The use of these equations can best be illustrated by solving example (b) at the beginning of this Section under *The Simple Coil.*

Example: What value of resistor should be added in parallel with a coil of 1 millihenry having a Q of 143 at 455 kc in order that the Q will be depressed to 100.

We have $Q_u = 143$, $Q_L = 100$ $L = 1$ mh $f = 455$ kc

By Eq. 78:

$$R_x = \frac{6.28(.455)10^6(1)10^{-3}(143)100}{143-100} = 950{,}000 \text{ ohms}$$

Also:

If 950,000 ohms represents the vacuum tube plate load on a 1 millihenry inductor at 455 kc, compute the unloaded Q if the loaded Q is 100. By Eq. 79:

$$Q_u = \frac{R_x Q_L}{R_x - \omega L Q_L} =$$

$$\frac{(950{,}000)100}{950{,}000 - 6.28(455)10^3(1)10^{-3}(100)} = 143$$

These equations will be used in the examples to follow to convert circuit Q's to unloaded Q's for use as transformer specifications.

In general a typical i-f system problem is solved in the following manner:

(a) The bandwidth ratio for any two positions on the system selectivity curve is computed from the given data; such as BW_{60} BW_3
(b) Using this ratio, within limits as may be prescribed for the application, the number of pairs of tuned circuits is determined from the *bandwidth factors* as given in Fig. 14-39.
(c) Knowing the required overall gain and the number of stages, determine the stage gain.
(d) Using the system bandwidth at any point, by use of the *bandwidth factor* from Fig. 14-39, determine the required Q.
(e) Select a suitable tube and with Eq. 71, determine the required inductance.
(f) Knowing L, C can be determined by Eq. 10.
(g) Determine the coefficient of coupling by the use of Eq. 72.
(h) Determine the mutual inductance by Eq. 63.
(i) Finally, construct the system selectivity curve by the use of the table of Fig. 14-39 and Eq. 75.
(j) By the use of Eq. 79, compute the unloaded Q values necessary to produce the calculated circuit Q determined by step d. [1]

This design procedure can best be illustrated by considering a few typical examples which are worked out in detail.

Example 1:

Required, an i-f system having a gain of 80 db at a center frequency of 455 kc and an overall bandwidth of 8.9 kc at 3 db and no more than 35 kc at 40 db. A tube having a g_m of 4400, plate resistance of .8 megohm and input resistance of 6.8 megohms, input capacitance of 6 μμf, output capacitance of 5.5 μμf will be used. An additional allowance of 2 μμf will be made for primary and secondary wiring capacitance.

Solution:

From Fig. 14-39 $\underline{K}$ at 3 db for one pair is 1.41

$$\textit{Then}\quad Q = \frac{\underline{K} f_o}{BW} = \frac{1.41(455)10^3}{8.9(10)^3} = 72$$

also $\underline{K}$ at 40 db is 14.14

$$\textit{and}\quad BW_{40} = \frac{\underline{K} f_o}{Q} = \frac{14.14(455)10^3}{72} = 89 \text{ kc}$$

Since 89 kc is outside of the specified 40 db bandwidth, try two pair.

For two pair, $\underline{K}$ at 3 db is 1.13

$$\textit{Then}\quad Q = \frac{\underline{K} f_o}{BW} = \frac{1.13(455)10^3}{8.9(10)^3} = 58$$

also $\underline{K}$ at 40 db is 4.46

$$\textit{and}\quad BW_{40} = \frac{\underline{K} f_o}{Q} = \frac{4.46(455)10^3}{58} = 35 \text{ kc}$$

Two pair are required to satisfy the selectivity requirements. The required Q of 58 is reasonable and less than two stages could hardly be expected to fulfill the gain requirements.

For 2 stages: Stage gain = 80/2 = 40 db or 100

$$(\textit{By Eq. 71}):\quad L = \frac{\textit{Stage gain (2)}}{\omega_o g_m Q}$$

$$= \frac{100(2)}{6.28(455)10^3(4400)10^{-6}(58)} = 275\ \mu h$$

Capacity for resonance at 455 kc:

$$C = \frac{1}{4\pi^2 f^2 L} = 445\ \mu\mu f$$

[1] For further Q corrections to simplify the actual construction of a transformer, the reader is referred to page 1029 of the Fourth Edition of the Radiotron Designer's Handbook, by F. Langford Smith, distributed by the Radio Corporation of America, Harrison, N. J.

(By Eq. 70):

$$k_c = \frac{1}{\sqrt{Q_1Q_2}} = \frac{1}{58} = .0172$$

$$M = k_c\sqrt{L_1L_2} = .0172(275) = 4.7\ \mu h$$

The circuit specifications therefore are:
Pri. ind. = sec. ind. = 275 μh
Pri. cap. = sec. cap. = 445 μμf
Pri. Q = Sec. Q = 58
Mutual = 4.7 μh
(Coefficient of coupling = .0172)

Converting the circuit Q's to unloaded Q's by Eq. 79 will enable us to write the final transformer specification.

$$\text{Pri. } Q_u = \frac{R_s Q_L}{R_s - \omega L Q_L} =$$

$$\frac{.8(10^6)58}{.8(10^6) - 6.28(.455)10^6(.275)10^{-3}(58)} = 61.6$$

$$\text{Sec. } Q_u =$$

$$\frac{6.8(10^6)58}{6.8(10^6) - 6.28(.455)10^6(.275)10^{-3}(58)} = 58.5$$

The final transformer specification becomes:
Pri. ind. = Sec. ind. = 275 μh
Pri. cap. = Circuit cap. − (tube + wiring cap) = 445 − 5.5 − 2 = 437.5 μμf
Sec. cap. = Circuit cap. − (tube + wiring cap.) = 445 − 6 − 2 = 437 μμf
Pri. Q = 61.6
Sec. Q = 58.5
Mutual inductance = 4.7 μh
(Coefficient of coupling = .0172)

The complete amplifier selectivity characteristic can be calculated with the constants obtained from Fig. 14-39 and Eq. (75) and are as follows:

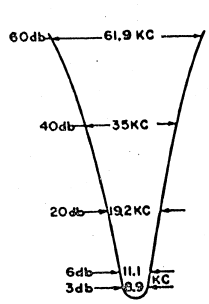
BW_3 = stated as problem requirement = 8.9 kc

$$BW_6 = \frac{\underline{K}_6 f_0}{Q} = \frac{1.41(455)10^3}{58} = 11.1\text{ kc}$$

$$BW_{20} = \frac{\underline{K}_{20} f_0}{Q} = \frac{2.45(455)10^3}{58} = 19.2\text{ kc}$$

$$BW_{40} = \frac{\underline{K}_{40} f_0}{Q} = \frac{4.46(455)10^3}{58} = 35\text{ kc}$$

$$BW_{60} = \frac{\underline{K}_{60} f_0}{Q} = \frac{7.96(455)10^3}{58} = 61.9\text{ kc}$$



Example 2:

Required, a narrow band i-f system, utilizing double-tuned optimum-coupled transformers centered at 4.3 Mc and having an overall bandwidth of 100 kc at 3 db and not greater than 1200 kc at 60 db. The overall gain of 80 db is to be obtained with battery operated tubes.

Solution:

$$\frac{BW_{60}}{BW_3} = \frac{1200}{100} \leq 12$$

Choose the number of stages for required selectivity by use of table (Fig. 14-39) as follows:

	$\underline{K}_3$	$\underline{K}_{60}$	$\underline{K}_{60}/\underline{K}_3$
1 pair	1.41	–	–
2 pair	1.13	7.96	7.0
3 pair	1.01	4.46	4.42

The $\underline{K}_{60}/\underline{K}_3$ ratio for 2 pair is the nearest obtainable to the required ratio of 12 or less, thereby indicating that the skirt selectivity at 60 db will be better than is required. This indicates the use of two stages.

Stage gain = 80/2 = 40 db or 100

As shown $\underline{K}$ = 1.13 for 2 pair at 3 db,

$$\text{then } Q = \frac{\underline{K} f_0}{BW_3} = \frac{1.13(4300)10^3}{100(10^3)} = 49$$

If we choose a CK-569-AX tube having a g_m of 1100 μmhos, we have from Eq. 71:

$$L = \frac{\text{Stage gain (2)}}{\omega_0 g_m Q} = \frac{100(2)}{6.28(4.3)10^6(1100)10^{-6}(49)}$$

$$= 137(10^{-6}) = 137\ \mu\text{ henries}$$

The value of C for resonance at 4.3 Mc with 137 μ henries is

$$C = \frac{1}{4\pi^2 f^2 L} = 9.0\ \mu\mu f$$

From Eq. (73) the coefficient of coupling is

$$k_c = \frac{1}{Q} = \frac{1}{49} = .0204$$

and mutual from Eq. 63 is

$$M = k\sqrt{L_1L_2} = .0204(137) = 2.8\ \mu h$$

The circuit specifications therefore are:
Pri. ind. = Sec. ind. = 137 μh
Pri. cap. = Sec. cap. = 9.0 μμf
Pri. Q = Sec. Q = 49
Mutual ind. = 2.8 μh
(Coefficient of coupling = .0204)

The circuit specifications can be corrected for loading as explained in example no. 1 in order to arrive at realistic transformer specifications.

The amplifier bandwidth at points other than 3 db can be calculated by use of the table of Fig. 14-39 and Eq. (75):

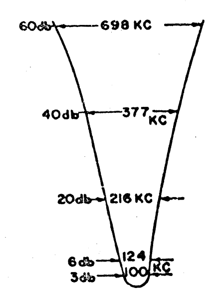
$$BW_6 = \frac{\underline{K}_6 f_0}{Q} = \frac{(1.41)4300(10^3)}{49} = 124\text{ kc}$$

$$BW_{20} = \frac{\underline{K}_{20} f_0}{Q} = \frac{2.46(4300)10^3}{49} = 216\text{ kc}$$

$$BW_{40} = \frac{\underline{K}_{40} f_0}{Q} = \frac{4.46(4300)10^3}{49} = 377\text{ kc}$$

$$BW_{60} = \frac{\underline{K}_{60} f_0}{Q} = \frac{7.96(4300)10^3}{49} = 698\text{ kc}$$

The complete selectivity characteristic is shown below:



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V_p/V_v	KQ	$Q\Delta f_p/f_0$	V_p/V_v	KQ	$Q\Delta f_p/f_0$
1.050	1.372	.94	1.26	2.02	1.77
1.055	1.39	.97	1.27	2.05	1.79
1.06	1.41	1.00	1.28	2.08	1.82
1.065	1.43	1.02	1.29	2.11	1.85
1.07	1.45	1.05	1.30	2.13	1.88
1.075	1.47	1.07	1.31	2.16	1.91
1.08	1.49	1.10	1.32	2.19	1.94
1.085	1.51	1.13	1.33	2.21	1.97
1.09	1.52	1.15	1.34	2.24	1.99
1.095	1.54	1.17	1.35	2.26	2.02
1.1	1.56	1.19	1.36	2.29	2.05
1.11	1.59	1.24	1.37	2.31	2.08
1.12	1.63	1.28	1.38	2.34	2.10
1.13	1.66	1.32	1.39	2.37	2.13
1.14	1.69	1.36	1.40	2.39	2.16
1.15	1.72	1.40	1.41	2.41	2.18
1.16	1.74	1.44	1.42	2.43	2.21
1.17	1.77	1.47	1.43	2.46	2.24
1.18	1.80	1.51	1.44	2.48	2.27
1.19	1.83	1.54	1.45	2.50	2.29
1.2	1.86	1.57	1.46	2.52	2.32
1.21	1.89	1.61	1.47	2.55	2.34
1.22	1.92	1.64	1.48	2.57	2.37
1.23	1.94	1.67	1.49	2.60	2.40
1.24	1.97	1.70	1.50	2.62	2.41
1.25	2.00	1.73			

Fig. 14-41 Table for the design of overcoupled-double-tuned transformers.

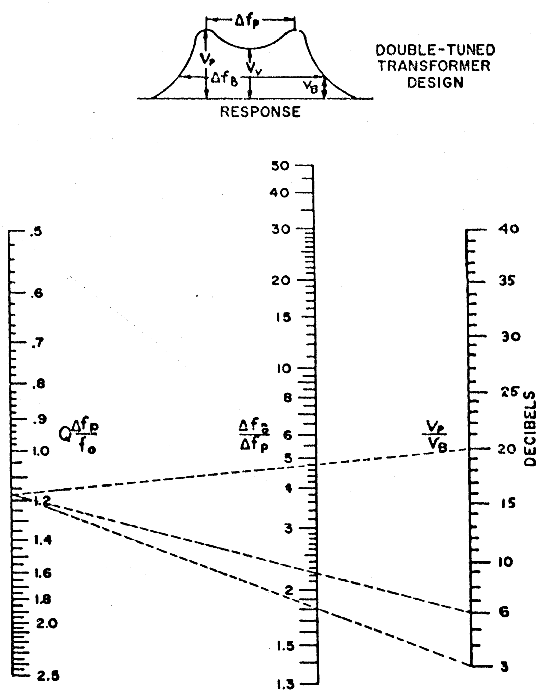


Fig. 14-42 Nomograph for the design of over-coupled double-tuned transformers. (Also see Fig. 14-41.)

Note: It should be noted that 2 double-tuned transformers actually have much better skirt selectivity than is required. Compare this with Example 2 shown on page 27 under single-tuned circuits.

Example 3:

A symmetrical double-tuned 455 kc i-f transformer has an inductance of 470 μh, Q of 60 and has the coupling adjusted for optimum. Find the mutual inductance required, and the bandwidth at 6 and 20 db.

Solution:

From 14-39 $\underline{K}_6 = 1.86$ *and* $\underline{K}_{20} = 4.46$

then: $BW_6 = \frac{K_6 f_0}{Q} = \frac{1.86(455)10^3}{60} = 14.1\ kc$

$BW_{20} = \frac{K_{20} f_0}{Q} = \frac{4.46(455)10^3}{60} = 33.8\ kc$

$k_c = 1/Q = 1/60 = .0167$

$M = kL = 7.85\ \mu h$

Over-Coupled: [1]

It is sometimes desired to increase the bandwidth of an i-f system above that which is obtained with a critically coupled system. From a single stage point of view where $BW = \sqrt{2}\ f_0/Q$, for $k_c = 1/Q$, it is possible by resorting to closer coupling to obtain a greater bandwidth factor than the $\sqrt{2}$, above.

In the following design method which is based upon the table of Fig. 14-41 and the nomograph presented as Fig. 14-42, the designer must first select the center frequency, the desired bandwidth, and the flatness of response curve. Once these choices have been made, the chart and nomograph permits one to obtain in an expedient manner the side skirt selectivity of the transformer.

In this design method it is assumed that the coil Q in most cases will actually be lowered by means of external resistors. It should be understood that this is not a positive design requirement, but rather is offered as the simplest method of obtaining the desired coil Q, since it will be found usually much simpler to lower the Q of a coil by means of resistors than to wind a coil which has the exact Q demanded by the design procedure. A typical transformer circuit complete with loading resistors is shown in Fig. 14-43.

[1] Based on a method described in "Exact Design and Analysis of Double and Triple-Tuned Band-Pass Amplifiers" by Milton Dishal, published in Proceedings of the I.R.E., Volume 35, Number 6, June, 1947.

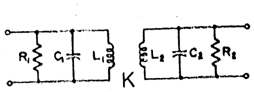


Fig. 14-43 Typical double-tuned transformer circuit.

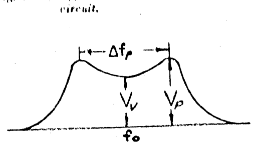


Fig. 14-44 Typical double-tuned overcoupled transformer response curve.

A typical response curve for an overcoupled double-tuned transformer is presented as Fig. 14-44. The flatness of the curve is expressed in terms of peak-to-valley ratio where V_p is the peak amplitude response and V_v is the valley amplitude response. The maximum flatness considered in the chart is 5 per cent, since values more closely approaching unity are of little practical significance.

The design procedure follows the simple steps of:

(1) Select the desired center frequency
(2) Select the desired bandwidth
(3) Calculate the ratio of the bandwidth to center frequency, $\Delta f_p/f_o$
(4) Select the desired peak to valley ratio V_p/V_v, then consult Fig. 14-41 to determine Q by using the $Q\ \Delta f_p/f_o$ column, then divide the value found in this column by the value computed for $\Delta f_p/f_o$, in step 3.
(5) Select suitable tube type.
(6) Select tuning capacitances.
(7) Determine gain per stage from the relation:

$$\text{Gain} = \frac{Q\Delta f_p}{f_0}\left(\frac{g_m}{4\pi\Delta f_p\sqrt{C_1C_2}}\right) \qquad (80)$$

(The gain thus obtained will indicate whether the choice of capacitance was appropriate; however, it should be kept in mind, that if the capacitance is lowered for the sake of increased gain, stability in frequency characteristics will be sacrified).

(8) Determine value of KQ from Fig. 14-41.
(9) Calculate coefficient of coupling K by dividing the value of KQ by the value of Q found in Step 4.
(10) Determine mutual inductance by first determining the primary and secondary inductances that resonate at the center frequency with the chosen capacitances, then obtain mutual by multiplying this value with the value obtained for K in Step 9.

The complete step-by-step procedure just outlined will be simplified if the statement of a specific design includes predetermined parameters such as center frequency and bandwidth, tube type, etc.

From the nomograph (Fig. 14-42), it is possible to determine the response at any specific point, or it is equally possible to plot the entire response curve of a transformer, should one so desire. The procedure is as follows:

(1) Using the value of $Q\ \Delta f_p/f_o$ previously determined by step 4 of the general design procedure, connect this point on column one of the nomograph with the desired db point on column three.
(2) Read value of $\Delta f_B/\Delta f_p$ on column two.
(3) Multiply value of $\Delta f_B/\Delta f_p$ just obtained by value of Δf_p to find bandwidth at desired db point.

The above procedure is illustrated in the two following examples and the dotted lines on the nomograph represent this part of the solution to example one.

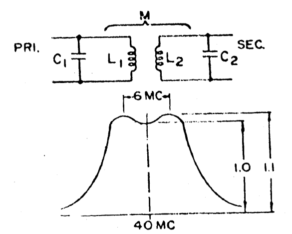
The number of stages required to produce the desired selectivity is easily obtained by dividing the desired attenuation in db's by the attenuation obtained in one stage. If the answer so obtained exceeds the number of stages required for gain, the difference in attenuation may be obtained by the use of a trap.

Examples:

To illustrate the use of this design method, the following examples have been worked out on a step-by-step basis.

Example No. 1 Typical 40 Mc Wide-band i-f stage:

Required, a wide-band i-f stage with center frequency of 40 Mc and a peak-to-peak bandwidth of 6 Mc. The peak to valley ratio shall be 1.1 and the stage will employ a 6CL6 tube having a g_m *of 10,000 micromhos.*



Solution:

Given $f_o = 40$ *Mc*
$\Delta f_p = 6$ *Mc*
$V_p/V_v = 1.1$
$g_m = 10,000$ *micromhos.*

1. Calculate value of $\Delta f_p/f_o = 6/40 = 0.15$

2. Consult Fig. 14-41 to determine Q. (Use $Q\Delta f_p/f_o$ *column).*

We find that when $V_p/V_v = 1.1$, $Q\Delta f_p/f_o = 1.19$ *and Step 1 determined* $\Delta f_p/f_o$ *to be 0.15*

Therefore $Q = 1.19/0.15 = 7.94$

3. Select value for C_1 *and* C_2.

The tube input capacity is 11 μμf and the output capacity is 5.5 μμf. To allow for socket, wiring, and distributed capacity, use 20 μμf for C_1 *and* C_2.

4. Compute gain per stage from equation 80 and the chart of Fig. 14-41:

Gain per Stage =

$$1.19\left[\frac{10^4(10^{-6})}{12.6(6)10^6(20)10^{-12}}\right] = 7.88$$

5. Consult Fig. 14-41, find value of KQ when $V_p/V_v = 1.1$, $KQ = 1.56$

6. Knowing Q from Step 2, calculate value of K,

$K = 1.56/7.88 = 0.198$

7. Determine mutual inductance,

$M = K\sqrt{L_1L_2}$

$L = \sqrt{L_1L_2} = L_1 = L_2 = \frac{1}{4\pi^2 C f_0^2}$

$\frac{1}{39.5(20)10^{-12}(1600)10^{12}} = 10^6/1.27 = 0.79\ \mu h$

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(This value of L could also have been obtained by the use of the reactance chart.

$M = .198\ (.79) = 0.156$ *microhenry*

The circuit specification, therefore, becomes:

$C_1 = C_2 = 20\ \mu\mu f$

$L_1 = L_2 = 0.79\ \mu h$

$Q_1 = Q_2 = 7.88$

$M = 0.156\ \mu h.$

and the transformer specification is

$C_1 = C_2 = 0$

$L_1 = L_2 = 0.78\ \mu h$

$Q_1 = Q_2$ = *any convenient value with circuit* $Q = 7.9$ *obtained with loading resistors.*

$M = 0.156\ \mu h$

8. *Determine the selectivity characteristic at 3, 6, and 20 db.*

(a) *For 3db connect the* $Q\Delta f_p/f_o$ *value on column one (1.175) with 3 db on column three. Read 1.81 on column two. Substituting 6 Mc for* Δf_p *in*

$$\Delta f_B/\Delta f_p = 1.81$$

we find Δf_p *or* $BW_3 = 1.81(6Mc) = 10.9\ Mc.$

(b) *Repeating the above procedure for 6 and 20 db we obtain*

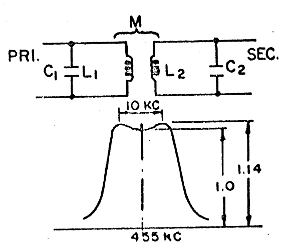
$BW_6 = 2.23(6Mc) = 13.4\ Mc$

$BW_{20} = 4.8(6Mc) = 28.8\ Mc$

The connecting lines illustrating the solution of this problem are shown in dotted outline on the nomograph of Fig. 14-42.

Example No. 2, Typical 455 kc medium wide-band overcoupled i-f:

Required, a medium wide-band i-f with center frequency of 455 kc and a peak to peak bandwidth of 10 kc. The peak to valley ratio shall be 1.14 and the stage will employ a 6SK7 tube having a g_m *of 2000 micromhos.*



Solution:

Given $f_o = 455$ *kc*

$\Delta f_p = 10$ *kc*

$V_p/V_v = 1.14$

$g_m = 2000$ *micromhos*

1. *Calculate value of* $\Delta f_p/f_o = 10/455 = 0.022.$

2. *Consult Fig. 14-41 to determine* Q_1 *(Use* $Q\Delta f_p/f_o$ *column).*

We find that when $V_p/V_v = 1.14$, $Q\Delta f_p/f_o = 1.36$

Therefore $Q = 1.36/0.022 = 61.8$

3. *Select value for* C_1 *and* C_2.
(The tube input and output capacities are 6 and 7 $\mu\mu f$ *respectively).*

For 1st trial select $C_1 = C_2 = 120\ \mu\mu f$

(Refer to discussion of Miller Effect in this section for suggestions relating to the value of tube capacity with respect to total capacity for resonance).

4. *Compute gain per stage from equation 80,*

$$\text{Gain per stage} = 60(1.36)\frac{2(10^3)(10^{-6})}{12.6(10^4)120(10^{-12})} = 178$$

5. *Consult Fig. 14-41, find value of KQ,*

When $V_p/V_v = 1.14$, $KQ = 1.69$

6. *Knowing Q from Step 2, calculate value of K,*

$K = 1.69/61.8 = 0.028$

7. *Determine mutual inductance,*

$M = K\sqrt{L_1L_2}$

$$L = \sqrt{L_1L_2} = L_1 = L_2 = \frac{1}{4\pi^2 C f_0^2}$$

$$= \frac{1}{39.5(120)10^{-12}(.455)^2 10^{12}} = 1020\ \mu h$$

$M = KL = 0.028(1020) = 28.6\ \mu h$

The transformer specification, therefore, becomes:

$C_1 = C_2 = 120\ \mu\mu f$ *(less tube capacitance)*

$L_1 = L_2 = 1020\ \mu h$

$Q_1 = Q_2 = 61.8$

$M = 28.6\ \mu h.$

8. *Determination of the selectivity characteristic as illustrated in example one gives the following:*

$BW_3 = 1.72\ (10kc) = 17.2\ kc$

$BW_6 = 2.1\ (10kc) = 21.0\ kc$

$BW_{20} = 4.45\ (10kc) = 44.5\ kc$

NETWORK THEORY

A transformer may be designed in terms of a general network. As will be shown in the following pages double tuned and triple tuned transformers or transformers with complex couplings, including bridging components can be designed by means of certain transformations and circuit equivalences.

The term network is a general term which can apply to a great variety of items as well as to widely differing circuit arrangements. In general, an electrical communication network consists of resistance, inductance, mutual inductances, and capacitances connected together in some manner. The various components which together form the network are known as *network constants* or elements.

A classification often applied to networks is based on the number of terminals to which other

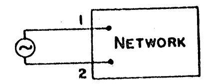


Fig. 14-45 Two-terminal network.

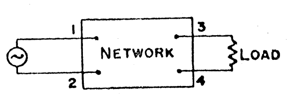


Fig. 14-46 Four-terminal network.

circuits may be connected. By far the most common are the two-terminal networks as represented by Fig. 14-45 and the four-terminal networks as represented by Fig. 14-46.

In the two-terminal network, the only applied voltage is that appearing between terminals 1 and 2 which are actually open points in the one branch of the network, and the impedance between these points is called the *input* or *driving point* impedance of the network.

The four-terminal network has one branch open at 1 and 2 where conventionally a voltage is applied while another branch is open at 3 and 4, at which point an output or load impedance is connected. This is the type of network that represents the average double- or triple-tuned transformer, filter, or similar circuit arrangement.

Driving Point Impedance

Two-terminal networks can best be defined in terms of their *driving point impedance*. An understanding of this type of network is important since the individual arms of more complex networks are in themselves basically two-terminal networks.

The driving point impedance of a reactive network may be represented graphically or mathematically. For purposes of this discussion, it seems best to start with the graphical presentation.

It is possible to represent driving point impedance by means of a simple Cartesian-system diagram where the abscissa (x axis) indicates the frequency scale and the ordinate (y axis) the magnitude of the reactance. Plotting reactance in this manner will produce a curve, the slope of which is

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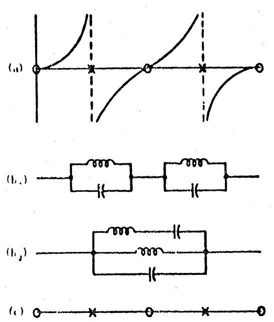
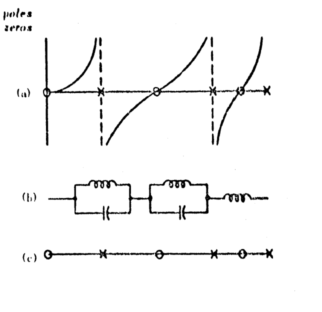
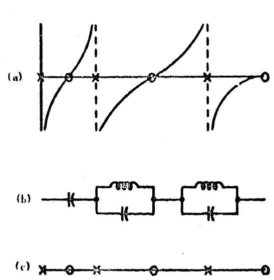


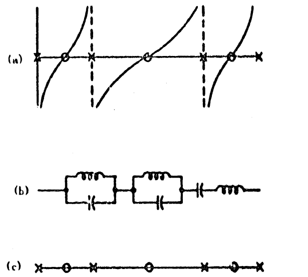
CASE I Zero at zero and infinite frequency.



CASE III Zero at zero frequency and pole at infinite frequency.



CASE II Pole at zero frequency and zero at infinite frequency.



CASE IV Poles at zero and infinite frequency.

Fig. 14-47 Four basic types of networks.

always positive. The frequencies at which the impedance is infinity are known as poles (antiresonant frequencies) and the points of zero impedance are called zeros (resonant frequencies).

It is the arrangement of the poles and zeros that determines the characteristics of the network and influences the driving point impedance which is best described as a complete specification of the ratio of the applied voltage to the resultant current at all frequencies. The usual manner of expressing this specification is in terms of poles and zeros which are called the critical frequencies.

In Fig. 14-47 are shown the four basic types of networks according to the arrangement of their poles and zeros. It should be noted that a pole must follow a zero but as long as poles and zeros alternate, they may be located at the discretion of the designer.

Because of the fact that it is the location of these critical frequencies that causes a network to pass or reject specific frequencies, a sort of shorthand system of notation has been developed as an aid to network analysis or synthesis, shown as line c in Fig. 14-47.

In Fig. 14-47, Case I, will be seen to consist of five critical frequencies and to begin and to end with a zero. Since this diagram is a representation of the driving point impedance of a network, certain facts about this network are at once apparent:

a. The presence of a zero at zero frequency (d-c) indicates the need for a continuous path through the network along which a direct current may flow.

b. The presence of a zero at infinite frequency points out the need for a continuous path through capacitative elements in order to provide a reactance path approaching zero (short-circuit) at frequencies approaching infinity.

c. Between these extremes of frequency, there are two poles and a zero - indicative of one point of zero impedance and two points of infinite impedance.

Two schematic diagrams, b_1 and b_2, are given as examples of networks which will fully satisfy the requirements of the reactance diagram of Case I. The very fact that two possible solutions are given to this case serves to point up the universal nature of a reactance diagram. It is important to recognize that either of the suggested networks will fulfill the requirements set forth in the reactance diagram, and that if both be enclosed within a "black box", they will prove indistinguishable one from the other, by any known electrical test.

A detailed analysis of the schematic diagram b_1 of Case I shows this network to be capable of satisfying fully the requirements of either a or c under Case I because of the presence of:

1. A direct current path through the two inductances to satisfy the zero at zero frequency.
2. A continuous path through the two capacitors to satisfy the zero at infinite frequency.
3. A pole (infinite impedance) resulting from parallel resonance of one of the LC combinations.
4. A pole (infinite impedance) resulting from parallel resonance of the second LC combination.
5. A zero (zero impedance) as a result of series resonance between the inductive and capacitive components of the two LC combinations.

The reader is left with the task of providing a similar analysis for b_2 under Case I and for Cases II, III, and IV.

It well to remember that the requirement for a zero at zero frequency can be met only by a continuous d-c path which means that there can be no series capacitor in the circuit. When a pole is specified at infinite frequency, this can be satisfied only by a series inductance, the reactance of which will be infinite at infinite frequency. Internal poles and zeros are merely a matter of series or parallel resonance of various LC combinations.

Foster's Reactance Theorem

Foster, a number of years ago, developed the mathematical analysis of the reactance function.[1] Any study of Foster's Reactance Theorem will show that the least number of circuit elements required in a network is one more than the sum of the the internal poles and zeros, and that the basic arrangement of circuit elements may call for either parallel or series connections.

We have shown in Case I of Figure 14-47 that a reactance function may be represented by more than one network configuration. The two basic configurations shown in Figures 14-48 and 14-49 will be found to be the most generally useful forms with which to synthesize an actual network. The fundamental equations for network synthesis as well as the particular equations specifically applying to each type of network will be given in the following paragraphs.

When the reactance function has a pole at the origin, the basic equation is:

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Driving Point Impedance = Z

$$= \pm j \frac{H}{\omega} \frac{(\omega^2-\omega_1^2)(\omega^2-\omega_3^2)\cdots(\omega^2-\omega_p^2)}{(\omega^2-\omega_2^2)(\omega^2-\omega_4^2)\cdots(\omega^2-\omega_q^2)} \quad (81)[2]$$

When the reactance function has a zero at the origin, the basic equation is:

Driving Point Impedance = Z (82)[2]

$$= \pm j\omega H \frac{(\omega^2-\omega_1^2)(\omega^2-\omega_3^2)\cdots(\omega^2-\omega_p^2)}{(\omega^2-\omega_2^2)(\omega^2-\omega_4^2)\cdots(\omega^2-\omega_q^2)}$$

In these equations, the angular velocities ω_1, $\omega_3,\ldots,\omega_p$ designated by odd subscripts correspond to internal zeros, while ω_2, $\omega_4,\ldots,\omega_q$ designated by even subscripts correspond to internal poles. It is important to remember that the subscripts indicate frequencies which are definite and fixed by design. *The only independent variable is ω. The plus sign* is used when there is a *pole at infinite frequency* and the *minus sign* when there is a *zero at infinite frequency.*

The relationships needed to determine magnitude of components in the above circuit are:

$$C_k = \left| \frac{j\omega_k}{Z_k} \right| \qquad (k = 2, 4, \ldots, q) \qquad (83)$$

There Z_k is the quantity obtained when the above formula for driving point impedance is solved with the term $(\omega^2 - \omega_k^2)$ omitted from the denominator of equations 81 and 82 and the resulting modified expressions evaluated for Z with $\omega = \omega_k$.

For every value of C_k

$$L_k = \frac{1}{\omega_k^2 C_k} \qquad (84)$$

If the network has a pole at infinity

$$L_{(q+2)} = H \qquad (85)$$

Note: Omit $L_{(q+2)}$ when network has a zero at infinity.

If the network has a pole at zero frequency

$$C_o = \left| \frac{1}{Z_o} \right| \qquad (86)$$

Note: Omit C_o when network has a zero at the origin.

where Z_o is the quantity obtained when ω is omitted from the denominator under H in Equation 81 and the resulting expression evaluated for Z with ω = O.

The simplest example of this type network is a loss-less series-tuned circuit. In this case the sign of the right hand of equation 81 will be positive since there is a pole at infinite frequency. There is only one zero at the resonant frequency $\omega_r/2\pi$. H is $L_{(q+2)}$ from equation 85 and thus equation 81 takes the form of

$$Z = j\frac{L_{(q+2)}}{\omega}(\omega^2 - \omega_r^2)$$

The value of C_o is $1/(L_{(q+2)}\omega_r^2)$ or

$$\omega_r^2 = \frac{1}{L_{(q+2)}C}$$

The reader may gain useful familiarity with the Foster reactance expression by proving the above equation equal to

$$j\omega L_{(q+2)} - \frac{j}{\omega C_o}$$

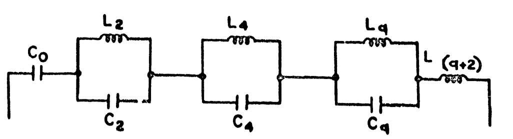

Fig. 14-48 Basic network configuration (type one).

[1] Ronald M. Foster, A Reactance Theorem, Bell System Tech. Jour. Vol. 3, pg. 259, April 1924. See also Sec. 3 of Radio Engineers Handbook by F. E. Terman, McGraw Hill, 1943

[2] The quantity H is a positive real constant that takes into account the fact that one additional piece of information is required to complete the specifications of the reactance function. See Eq. 85.

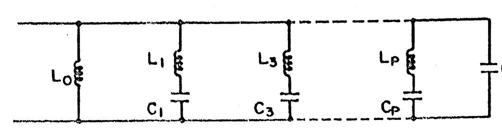

Fig. 14-49 Basic network configuration (type two).

The Foster reactance expression may seem to be too sophisticated for this simple case, but it will be apparent that it is very useful for more complex configurations and for the behavior of a combination of several such branches in a network.

When the network takes the form shown in Fig. 14-49 above, the following relationships will determine the value of the circuit components:

$$L_k = j\omega_k Z_k \qquad (k = 1, 3, \ldots, p) \qquad (87)$$

Where Z_k is obtained by omitting $(\omega^2 - \omega_k^2)$ from the numerator of equations 81 and 82 and evaluating the modified expression for $\omega = \omega_k$.

For each L_k, there will be a

$$C_k = \frac{1}{\omega^2 L_k} \qquad (88)$$

If the network has a zero at infinite frequency

$$C_{(p+2)} = \frac{1}{H} \qquad (89)$$

Note: Omit capacitor $C_{(p+2)}$ when network has a pole at infinity.

If the network has a zero at zero frequency

$$L_o = Z_o$$

Note: Omit inductor L_o when network has a pole at the origin.

where Z_o is the quantity obtained by omitting the ω that multiplies H in Equation 82 and then evaluating the modified expression for Z with ω = O.

It is helpful to remember that when ω is equal to any even subscript (ω_2, $\omega_4,\ldots,q$), the result is a pole. The reason is immediately apparent since this condition ($\omega = \omega_2$ etc.) necessitates that one member of the denominator of the basic equations is of zero value and therefore the driving point impedance becomes infinity. Conversely, when ω equals any odd sub-script (ω_1, $\omega_3,\ldots,p$) the result is a zero since this condition makes one term in the numerator equal to zero and therefore the driving point impedance becomes zero.

Another point which should be remembered is that there are a number of possible LC combinations that will satisfy a particular network requirement. However, the moment that one capacity or inductance value is decided upon, all others are at once defined in terms of the H factor.

There are occasions on which a comb filter - that is, one which passes certain frequencies while attenuating others - is desired. Such a filter can be built up in two-terminal form by properly locating the poles and zeros. In general, however, it will be found that the main usefulness of two-terminal networks is realized when such carefully designed two-terminal networks are used as branches of more complex networks.

Lattice Networks

One of the most universal methods for network design involves the use of the equivalent lattice,

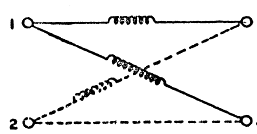

Fig. 14-50 Basic configuration of equivalent lattice.

the basic configuration of which is shown in Fig. 14-50. The equivalent lattice network of any symmetrical network may be obtained by bisecting the original network and substituting the short circuited bisected part for the series arm of the lattice and the open circuited bisected part for the lattice arms.

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Since an inductively-coupled, double-tuned transformer is a good example of a typical four-terminal network, the following schematic diagrams will show the various ways in which this type of network may be represented. Fig. 14-51 is the conventional diagram showing primary and secondary inductances and capacitances along with the mutual inductance between the windings.

Figure 14-52 shows the equivalent ladder network where the mutual inductance forms a common branch between the output and input circuits. The L_p-M and L_s-M inductances correspond to the leakage inductances in the transformer. One proof of the equivalence of these networks is the fact that short circuiting one end of either network will offer the same identical impedance at the other end.

In Figure 14-53 we have the bisected network. Here the center branch has a value double that shown in the previous drawing. This is explained by the fact that if we were to take the mirror image of this network and connect it in parallel, the center branch would then have its normal value of M.

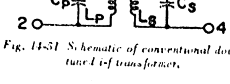


Fig. 14-51 Schematic of conventional double-tuned i-f transformer.

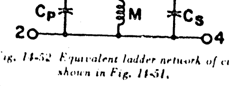


Fig. 14-52 Equivalent ladder network of circuit shown in Fig. 14-51.

Figure 14-54 shows the equivalent lattice network with one series arm (1-3) and one lattice arm (1-4). The other arms usually are shown symbolically as is done here by dotted lines and are equal to the corresponding arms shown in detail. The series arm of the equivalent lattice network corresponds to the driving point impedance of the bisected network with the 2M arm short circuited while the lattice arm is equivalent to the same bisected network open circuited.

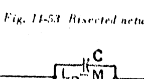


Fig. 14-53 Bisected network.

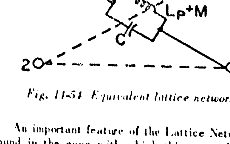


Fig. 14-54 Equivalent lattice network.

An important feature of the Lattice Network is found in the ease with which this type of circuit may be used to represent any symmetrical, balanced, four-terminal network. Fundamentally, a lattice network is a bridge circuit. This point is illustrated by Figures 14-55 and 14-56. Because of the symmetrical nature of this type of network, it is customary to work with only one half at a time and to draw the diagram as shown in Figure 14-57 with the second half in dotted lines - if shown at all. For purposes of uniformity, it is accepted that in lattice network diagrams, the series arm is that extending between terminals 1 and 3 and the lattice arm is that between terminals 1 and 4.

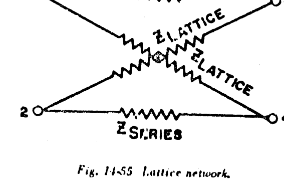


Fig. 14-55 Lattice network.

Image Impedance[1]

The image impedance of a network is of primary importance because it is only when operating under image-impedance conditions that optimum power performance is obtained from the network. The image impedance of any network is equal to the square root of the product of the short-circuited and the open-circuited impedances of the network. In the case of π and T networks, the calculations required to determine the image impedance frequently become somewhat cumbersome. In the case of lattice networks, the only necessary calculation is the extraction of the square root of the product of the impedance of the lattice and series arms. In other words,

$$Z_I = \sqrt{Z_L Z_S} \qquad (90)$$

The image-impedance of a network determines the response of that network to a transient, and in view of the increased attention being given to the transient problem in modern electronic design, it is apparent that an understanding of image impedance and its effect upon damping will prove helpful to a design engineer.

Circuit Damping

Critical damping is obtained when the image impedance is matched by the load thus providing both ideal response to a transient and a substantially flat-topped pass band. When the terminating resistance is larger than the image impedance of the network, the circuit is under-damped. Under such conditions, the introduction of a transient is likely to produce ringing[2] and the pass band will be substantially uneven on top. When the terminating resistance is smaller than the image-impedance of the network the circuit is over-damped and will show a peaked characteristic. Either form of mismatching will, in the case of sufficiently complex networks, result in echoes or ghosts.

Design of a Lattice Network

Lattice networks have two major advantages over other network forms. They are:

1. The image impedance of the network is readily obtainable since it is equal to the geometric mean of the series and lattice arm impedances.
2. By drawing reactance diagrams of the lattice and series arms, it is possible to see pass-band and stop-band regions. This statement holds, both for network analysis or synthesis.

To assist in the practical use of lattice networks, a sort of shorthand notation for the driving point impedance has been developed. An example of this is shown in Fig. 14-58 where a band-pass filter is shown in this notation.

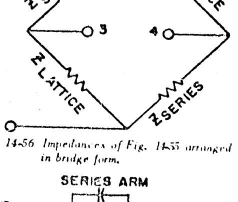


Fig. 14-56 Impedances of Fig. 14-55 arranged in bridge form.

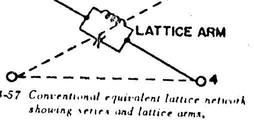


Fig. 14-57 Conventional equivalent lattice network showing series and lattice arms.

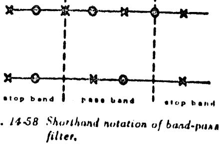


Fig. 14-58 Shorthand notation of band-pass filter.

[1] For a more detailed discussion of Image Impedance see Radio Engineers' Handbook by F. E. Terman, First Ed. 1943, McGraw-Hill or Communication Engineering by W. L. Everitt, Second Ed. 1937, McGraw-Hill.

[2] A periodic phenomena occuring after a step-type signal wave. Depending upon the extent of the damping this transient oscillation lasts for various lengths of time and is of varying amplitude.

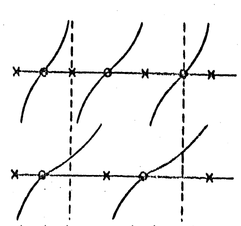


Fig. 14-59 Band-pass filter of Fig. 14-58 with reactance curves added.

The reasons behind the behavior of this filter arrangement may be somewhat clarified if, as in Fig. 14-59 we add reactance curves to the lines representing the series and lattice arms of the network. It will then be seen that the pass-band portion of the filter has its reactances of opposite sign while the stopband begins, and ends at the points where the impedances in the two arms of the lattice have the same sign.

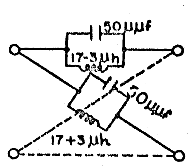
Actually, the reason for band-pass in a lattice network is a difference in the resonant frequencies $L_p - M$ in the series arm and $L_p + M$ in the lattice arm. For example, if it is desired to construct a simple lattice-type network to pass from f_1 to f_2, the following steps would produce the desired results:

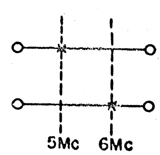
1. Select a value for C.
2. Find the L which, with the chosen C, will resonate at f_1.
3. Find the L which, with the chosen C, will resonate at f_2.

The difference between the two values of L will be equal to 2M.

Example:

Design a simple lattice-type network to pass from 5 to 6 Mc.
1. Assume a 50 μμf for C.
2. Calculate L_1 to resonate with C at 5 Mc or use reactance chart in the Appendix. L_1 = 20 μh.
3. Calculate L_2 to resonate with C at 6 Mc. L_2 = 14 μh.
The lattice then assumes the values shown below:





The pass-band will be 5 to 6 Mc. At the center frequency of 5.5 Mc the impedance of the arms according to equation 82 is:

$$Z = -\frac{\omega_c}{C(\omega_c^2 - \omega_f^2)}$$

which for the series arms is

$$Z = -\frac{5.5(10^6)}{50(10^{-12})[5.5^2(10^{-12}) - 6^2(10^{12})]}$$

$$= \frac{5.5(10^6)}{50(5.75)} = 19{,}000 \text{ ohms}$$

and for the lattice arm

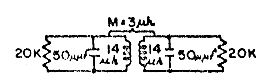
$$Z = -\frac{5.5(10^6)}{50(10^{-12})[5.5^2(10^{-12}) - 5^2(10^{12})]}$$

$$= \frac{5.5(10^6)}{50(5.25)} = 21{,}000 \text{ ohms}$$

and the image impedance according to equation 90 is:

$$Z_1 = \sqrt{19{,}000\,(21{,}000)} = 20{,}000 \text{ ohms}$$

and the equivalent transformer is shown below



It must be understood that this example is an extremely simple one and that the inclusion of more cardinal points would sharpen the sides of the pass-band while the top could be flattened appreciably by including poles and zeros exactly opposite each other.

This method can also be used for the design of more complex networks such as a capacitively-bridged transformer providing high attenuation of one particular frequency. The schematic of such a transformer and its equivalent bridged-T network and lattice network is shown in Fig. 14-60.

In this case both the inductances and the capacitances in the series and lattice arms are different, thus the two impedances can be made to be equal at one frequency. By reference to Fig. 14-55 it can be seen that under this condition the lattice (or bridge) will be balanced and the output will be zero. Actually, due to losses in the circuit, infinite attenuation will be obtained only if the resistances in both arms are also equalized. This process is called resistance cancellation. At the frequency of infinite attenuation the reactances are

$$\frac{\omega_\infty}{C_s(\omega_\infty^2 - \omega_s^2)} = \frac{\omega_\infty}{C_L(\omega_\infty^2 - \omega_L^2)} \qquad (91)$$

Therefore

$$\frac{C_s}{C_L} = \frac{\omega_\infty^2 - \omega_L^2}{\omega_\infty^2 - \omega_s^2} \qquad (92)$$

Since the left hand side of equation 92 is

$$\frac{C_1 + 2C_B}{C_1}$$

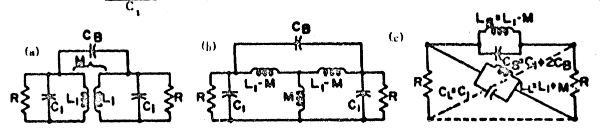


Fig. 14-60 Capacitively-bridged transformer (a); its equivalent bridged-T network (b); equivalent lattice network (c).

from Fig. 14-60, the bridging capacitance is

$$C_B = \frac{\omega_\infty^2 - \omega_L^2}{\omega_\infty^2 - \omega_s^2}\left(\frac{C_1}{2}\right) - \frac{C_1}{2} \qquad (93)$$

and from this the series and lattice inductances may be determined, since

$$L_s = \frac{1}{\omega_s^2 C_s} = \frac{1}{\omega_s^2 (C_1 + 2C_B)} \qquad (94)$$

and

$$L_L = \frac{1}{\omega_L^2 C_1} \qquad (95)$$

and the mutual inductance is

$$M = \frac{L_L - L_s}{2} = \tfrac{1}{2}\left(\frac{1}{\omega_L^2 C_1} - \frac{1}{\omega_s^2 (C_1 + 2C_B)}\right) \qquad (96)$$

and therefore the actual primary or secondary inductance is

$$L = L_s + M \qquad (97)$$

The midband image impedance and therefore R is then computed from equation 90.

Example:

For an actual example, let us assume the following requirements: a tuning capacitance (C_1) of 5 micro-microfarads, a pass-band extending from 8.5 to 12.5 megacycles and an infinite attenuation frequency (f_∞) of 14.25 megacycles. We may use in equation 93 the frequencies directly, as $4\pi^2$ can be factored out in both the numerator and denominator, and thus cancel out. Accordingly from 93

$$C_B = \frac{(14.25^2 - 8.5^2)}{(14.25^2 - 12.5^2)}(5/2) - 5/2$$

$$= \frac{201.8 - 72.25}{201.8 - 156.25}(2.5) - 2.5 = 4.5\ \mu\mu f$$

POOR ORIGINAL

Having thus obtained the bridging capacitance, the value of the mutual inductance may be computed from equation 96.

$$M = \tfrac{1}{2}\left(\frac{1}{4\pi^2(72.25)5} - \frac{1}{4\pi^2(156.25)14}\right)$$

$$= \tfrac{1}{2}(1/14{,}300 - 1/87{,}200) = 29\ \mu h$$

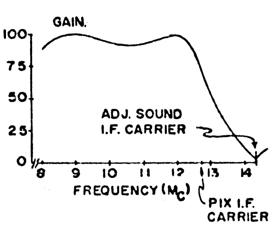
The value of L_s may be computed from equation 94 or looked up from a reactance chart yielding

$$L_s = \frac{1}{4\pi^2(72.25)5} = 11.2\ \mu h$$

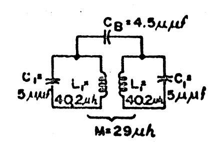
or making both the primary and secondary inductances equal.

$$L_1 = 29 + 11.2 = 40.2\ \mu h$$

The image impedance is 4,700 ohms from equation 90 computed at 10.5 Megacycles. The response of the transformer computed above is shown by



and the schematic is



FEEDBACK

In the many cases for improved stability and lower distortion as well as generally improved performance, it is customary to design amplifiers with a definite amount of feedback. In practice, this consists of providing a conductive path having electrical characteristics such as to permit a desired amount of the output voltage of the amplifier to be added to the signal voltage as it enters the amplifier.

Feedback may be either positive or negative which is to say that the voltage returned to the input of the amplifier may either support or oppose the signal voltage. Positive feedback - that is, where the feedback voltage adds to the signal voltage thereby increasing the amplifier gain - may result in oscillation and is used as the basis of most types of oscillators.

Negative feedback - where the feedback voltage opposes the signal voltage - is the type of feedback most often encountered in amplifiers for use at radio frequencies. The introduction of negative feedback adds greatly to the stability of an amplifier as well as providing reduced amplitude distortion and reduced terminal impedances.

Fig. 14-61 represents the operation of the feedback circuit in a typical feedback amplifier. A signal voltage, e_s, is impressed across the input of the amplifier which has an amplification factor of A giving an output voltage equal to E_{out}. A feedback path is provided between E_{out} and e_s which permits a definite proportion, β, of the output voltage to reach the input of the amplifier where because of a 180 degree phase difference it opposes the signal voltage and thus lowers the effective gain of the amplifier.

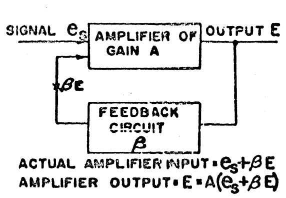


Fig. 14-61 Feedback circuit in a typical feedback amplifier.

Reference to Figure 14-61 will show the actual input of the amplifier to consist of $e_s + \beta E$ and the output to consist of $A(e_s + \beta E)$. From this we see that the gain of the amplifier, taking into account feedback, is equal to $A/(1 - \beta A)$. Since in the case of negative feedback, the quantity BA has a negative value, it follows that

Gain with negative feedback =

$$\frac{A}{1 + \beta A} \qquad (98)$$

FEEDBACK DUE TO GRID TO PLATE CAPACITY (C_{gp})

The fact that there is a definite amount of capacitance between the grid and the plate of every conventional vacuum tube introduces a certain amount of feedback into every circuit in which a tube is used as an amplifier.

The effect of this inter-electrode capacitance can probably best be made clear by reviewing certain basic relationships. For example, in Fig. 14-62, the voltage developed across the plate tank circuit (E_{out}) is equal to the voltage on the signal grid, e_s, times the amplification factor, A. Since A is equal to $g_m Z_{out}$, we can now set up the following equations:

$$E_{out} = Ae_s = g_m Z_{out} e_s$$

To further simplify the explanations of the effect of C_{gp} which are to follow, we will assume that the impedance of the input and output (grid and plate) tank circuits are approximately equal. This assumption is, of course, in accordance with the conditions most often encountered in actual practice, since standard design procedures call for the relationship $Z_{gk} = Z_{out}$.

We can represent the circuit of Fig. 14-62 in a slightly different manner as shown in Fig. 14-63. Here, C_{gp} represents the plate to grid capacitance of the tube and Z_{in} the impedance of the grid tank circuit across which the feedback voltage, e_{fb}, is applied.

Still another way of representing this same circuit appears in Fig. 14-64 where E_{out} is the voltage developed across the plate load, Z_{fb} represents the reactance of the grid-plate capacitance (C_{gp}), and Z_{in} is the impedance of grid tank circuit across which the feedback voltage (e_{fb}) develops. It is this voltage which is applied to the grid along with the signal voltage and is the βE shown in Fig. 14-61.

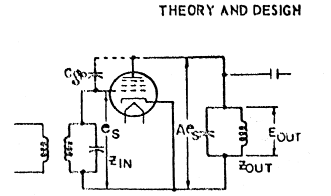


Fig. 14-62 Single stage amplifier.

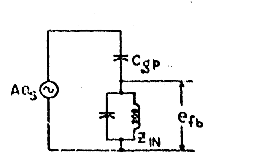


Fig. 14-63 Equivalent circuit of single stage amplifier.

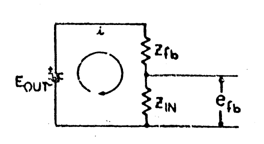


Fig. 14-64 DC represenation of single stage amplifier.

From Figures 14-63 and 14-64 we see that

$$Z_{fb} = \frac{1}{j\omega C_{gp}} \qquad (99)$$

and

$$Z_{in} = \omega LQ = Q/\omega C \qquad (100)$$

Now since i represents the total current flowing in the plate circuit, we have

$$i = \frac{E_{out}}{Z_{fb} + Z_{in}} \qquad (101)$$

From this we see that

$$e_{fb} = IZ_{in} = \frac{E_{out} Z_{in}}{Z_{fb} + Z_{in}} \qquad (102)$$

Substituting previously established values for E_{out}, Z_{in}, and Z_{fb} we can now write

$$e_{fb} = \frac{A e_g \omega L Q}{1/j\omega C_{gp} + \omega L Q} \qquad (103)$$

It now becomes possible to express the ratio between the feedback voltage and the signal voltage as

$$e_{fb}/e_g = \frac{A \omega L Q}{1/j\omega C_{gp} + \omega L Q} \qquad (104)$$

This ratio of feedback voltage to signal voltage is extremely important since it is this factor which determines the stability of any amplifier. Ratios having a magnitude of less than one are indicative of good stability.

It will be recalled that amplification with negative feedback at the center frequency may be expressed as

$$A_o = A/(1 + \beta A) \qquad (98)$$

In the foregoing expression, A and βA are vector quantities and as such have magnitudes and phase angles, both of which vary with frequency. As a result, βA may be either larger or smaller than one and may be either positive or negative in value.

It will be apparent from the feedback formula that when βA is negative, the feedback will be negative and there will be a decrease in amplification. This is the condition which is sought in the average amplifier design.

Under other conditions, βA may be positive and less than one in magnitude in which instance the feedback will be positive and amplification will be increased, although this type of system will tend to be unstable - particularly during the warm-up period.

When βA is equal to one, amplification becomes theoretically infinite and the system is completely unstable while with βA positive and greater than one the system will oscillate.

The conditions under which an amplifier will be stable have now been fairly well defined. Reduced to the simplest possible terms, amplifier stability becomes a matter of four factors, the last of which will be recognized as the Nyquist Criterion.[1] These four basic factors are:

1. For the average amplifier, βA should be negative for normal operation.
2. The gain at phase shifts of 0 to 90 degrees is degenerative (amplification is reduced).
3. The gain at phase shifts of 90 to 270 degrees is regenerative (amplification is increased).
4. At 180 degrees *phase shift*, for stable operation amplifier gain must be zero db's (unity) or less which is to say that in the Nyquist diagram the -1, 0 point must not be encircled.

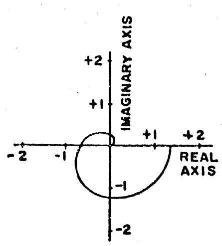


Fig. 14-65 Typical Nyquist diagram.

PHASE SHIFT

The importance of phase shift as a factor in amplifier stability is shown by the emphasis placed upon it in the foregoing discussion. It is important to remember that with linear phase shift, as the frequency increases, so does the phase angle. It is because of this fact that even the best possible amplifier designs may have their phase angles reaching 180 degrees.

The ratio of phase shift to the angular velocity constitutes a direct measure of time delay. Mathematically, this relationship may be expressed as

$$\tau = \phi/\omega \qquad (105)$$

[1] For further discussion see Regeneration Theory, H. Nyquist, Bell System Tech. Jour., Vol. II, pg. 126, January 1932. Also Sec. 5, art. 11, pg. 395, Radio Engineers' Handbook, F. E. Terman, First Ed., 1943, McGraw-Hill.

Because phase shift is essentially a time delay phenomenon, two different frequencies with the same time delay will produce different phase shifts and the higher the frequencies, the greater the magnitude of the shift. This sort of situation is illustrated by Figure 14-66.

In all amplifiers where the input tank and load circuits are purely resistive, there will be negative feedback as a result of a 180 degree phase reversal within the tube. It is significant that this phase reversal is instantaneous and has no time delay associated with it. Such phase reversal is possible only in active networks of which the vacuum tube is representative. In passive networks a change of phase can be obtained only with an accompanying time delay.

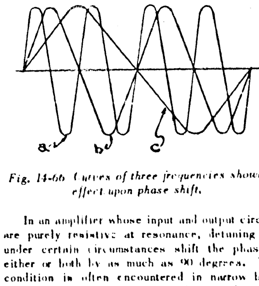


Fig. 14-66 Curves of three frequencies showing effect upon phase shift.

In an amplifier whose input and output circuits are purely resistive at resonance, detuning may under certain circumstances shift the phase of either or both by as much as 90 degrees. This condition is often encountered in narrow band, high gain amplifiers where the designer has been looking for maximum performance and has used i-f's with high inductances, high Q's, and minimum capacitances. Under such conditions, it is perfectly possible for the normal aging of a tube to introduce a change in the C_{gp} of sufficient magnitude to produce instability in the amplifier. This instability will manifest itself in the form of audible whistles resulting from beat oscillations occurring as a result of detuning.

When working with marginal designs of the type described above, it is possible because of the damping effect of the signal generator due to its low output impedance, to fully align an i-f amplifier only to have it go into oscillation the instant the generator is removed and the Q restored to the circuit. Again, it is sometimes possible to align such an amplifier and to have it remain perfectly stable until such future time as a tube ages and the Miller Effect produces detuning to an extent producing the phase shift required for oscillation. Under such conditions, the amplifier becomes unstable.

Good design obviously calls for the selection of circuit parameters which will provide a margin of safety sufficient to avoid excessive regeneration as a result of normal aging of components.

There are occasions when it is desirable to design an amplifier which will insure the maximum possible amplification-bandwidth-product (ABW). A circuit similar to that shown in Fig. 14-67 is often used under such circumstances.

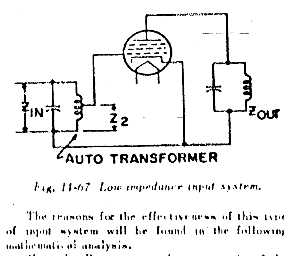


Fig. 14-67 Low impedance input system.

The reasons for the effectiveness of this type of input system will be found in the following mathematical analysis.

If we let R_t represent the turns ratio of the transformer while the subscript 1 applies to the whole winding and the subscript 2 to the tapped portion, we have seen that

$$Z_2/Z_1 = 1/R_t^2$$

Substituting, we see that this expression can be written

$$\frac{1}{j\omega C_{gk}} = \frac{1}{j\omega C_{apparent} R_t^2} \qquad (106)$$

Cancelling out $1/j\omega$ gives $C_{gk} = R_t^2 C_{apparent}$

or

$$C_{apparent} = C_{gk}/R_t^2 \qquad (107)$$

From the foregoing, it can be seen that by the use of an autotransformer in the grid circuit it is possible to reduce the effect of the grid-cathode (input) capacity of a vacuum tube by a factor equal to the square of the turns ratio in the input system. The improvement in stability resulting from a low impedance tap on the input tank will be found in many instances to be well worth the cost involved.

CONCLUSION

It is rather generally recognized throughout the coil industry that transformer design is usually conducted more as an art than as a science. A basic reason for this situation is found in the difficulties attached to the substitution of specified procedures for experience.

While every possible attempt has been made to present the problems surrounding optimum design practices in simple, straight-forward discussions, there will be constant need for selection of values on the part of the designer who uses this manual. Here again, an attempt has been made to provide sufficient background information to minimize the possibility of serious errors as a result of unwise choices on the part of an inexperienced engineer. For example, attention has been called to the importance of choosing a proper value of C in the design of parallel resonant circuits by pointing out the probable effect of using too large or too small capacitance values.

As a means of supplying that background knowledge for which no design procedure can ever completely substitute, it is recommended that those designers whose experience is limited will do well to keep complete notebooks in which are recorded the various steps through which it was necessary to progress before arriving at a satisfactory transformer design. Intelligent use of and reference to such records should soon lead to a degree of experience which in connection with the design methods set forth in this manual will lead to a minimum of design errors.

BIBLIOGRAPHY

Aiken, C. B.
Two-Mesh Tuned Coupled Circuit Filters
Proceedings of the I.R.E., Volume 25, Number 2, February 1937

Bruene, Warren B.
How to Design R-F Coupling Circuits
Electronics, May 1952
McGraw-Hill Book Co., Inc., New York, N.Y.

Dean, C. E.
Bandwidth Factors for Cascade Tuned Circuits
Electronics Reference Sheet, Electronics
McGraw-Hill Book Co., Inc., New York, N.Y.

Dishal, Milton
Exact Design and Analysis of Double and Triple-Tuned Band-Pass Amplifiers
Proceedings of the Institute of Radio Engineers, Volume 35, Number 6, June 1947

Langford-Smith, F.
Radiotron Designer's Handbook, Fourth Edition
Reproduced and Distributed by Radio Corporation of America, Harrison, N.J.

Terman, F. E.
Radio Engineering, Third Edition, 1947
Electronic and Radio Engineering, Fourth Edition, 1955
Radio Engineers' Handbook, First Edition, 1943
McGraw-Hill Book Co., Inc., New York, N.Y.

Valley and Wallman
Vacuum Tube Amplifiers, First Edition, Third Impression
McGraw-Hill Book Co., Inc., New York, N.Y.

Whyte, John R.
Choosing Pentodes for Broad-Band Amplifiers
Electronics Reference Sheet, Electronics, April 1952
McGraw-Hill Book Co., Inc., New York, N.Y.

INTRODUCTION TO APPENDIX

The following pages include data of a reference nature believed to be more readily accessible when separated from the main text. Some of this material has been referenced in various Sections of the text, while the remainder is included to aid in the general understanding of i-f and r-f transformer performance.

It is recommended that the designer study this introduction carefully prior to using the charts and data included in this appendix.

Pages A-1 and A-2 describe the test set-up for making temperature coefficient measurements. This equipment was used for taking all data pertaining to temperature coefficient of coils presented in the text.

Page A-3 graphically presents the temperature coefficients of typical universal windings on ceramic forms, each having a different type of coil impregnation. Data for low and high temperature extremes for representative impregnating materials are shown.

Page A-4 and A-5 graphically illustrate the effect of static humidity and elevated temperature, respectively, upon the Q of typical windings. Ceramic forms have been used throughout in order to minimize the effects of temperature and humidity on coil form material, so that the observed changes will truly reflect the behavior of the winding and its associated impregnating compound.

Pages A-6 through A-13 present the variation of inductance, Q, resistance and distributed capacity with turns. It should be realized that the form factor and type of wire used for these data may not be identical to that confronting the reader, but it should be noted that a similarity exists between the different families of curves. It is impossible to anticipate the many configurations that will be encountered by the user of this manual. Since coil design for similar end use generally follows an established trend, it is suggested that the reader supplement the data presented herein with similar curves derived from experimental windings representative of the type in which he is currently interested. Within a short time, the supplement will cover most of the day-to-day design problems and the necessity for making trial windings will be minimized.

Pages A-14 through A-18 deal with the various parameters of multi-pi universal windings and are intended to aid the designer of multi-pi coils in the same manner as the data on pages A-6 through A-13 aids the designer of single-pi windings.

Pages A-19 and A-20 illustrate the effects on Q and resistance of broken litz strands and cross-overs per turn, respectively. The first curve will be found useful in determining the number of broken strands that can be tolerated in production practice without seriously affecting coil performance. The second curve is useful in connection with winding practice as developed in section 10.

Pages A-21 and A-22 show the effect upon Q when the wire size is varied while maintaining a given winding-machine setup.

Page A-23 illustrates a useful form for recording coil data so that a complete log of a winding's development can be maintained for ready future reference. A series of such data sheets collected over a period of time will provide basic winding information for the immediate solution of many future designs.

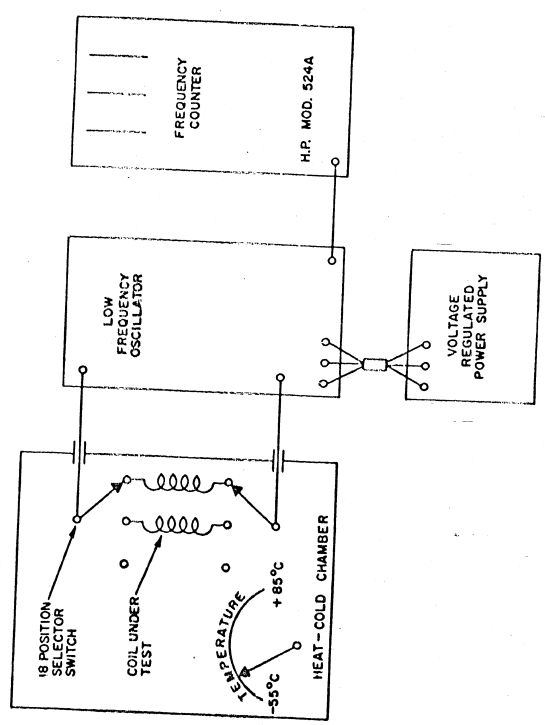
Page A-24 is the schematic diagram of a typical two stage test jig for the measurement of input or output i-f transformer characteristics.

Page A-25 defines an experimental 135 kc i-f transformer with respect to winding information, schematic diagram, and coil measurements in and out of shield, with and without iron core. The curves on pages A-26 through A-31 present the response of the i-f transformer defined on page A-25 when used both as an input and an output stage. Performance curves illustrate the effect of three different coil spacings to provide under coupling, critical coupling and over coupled conditions connected as capacity-aiding and capacity opposing.

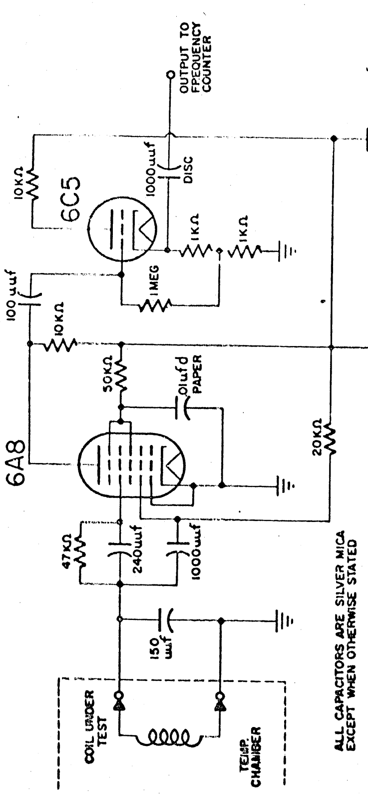
Pages A-32 through A-58 present information for transformers having operating frequencies of 262, 455, 1400 and 4300 kc, similar to that given on pages A-25 through A-31 for the 135 kc i-f transformer.

A copper wire table covering wire commonly used for r-f and i-f coils and transformers is shown on page A-59. Sizes 45 to 50 are seldom used in practice but have been included for reference.

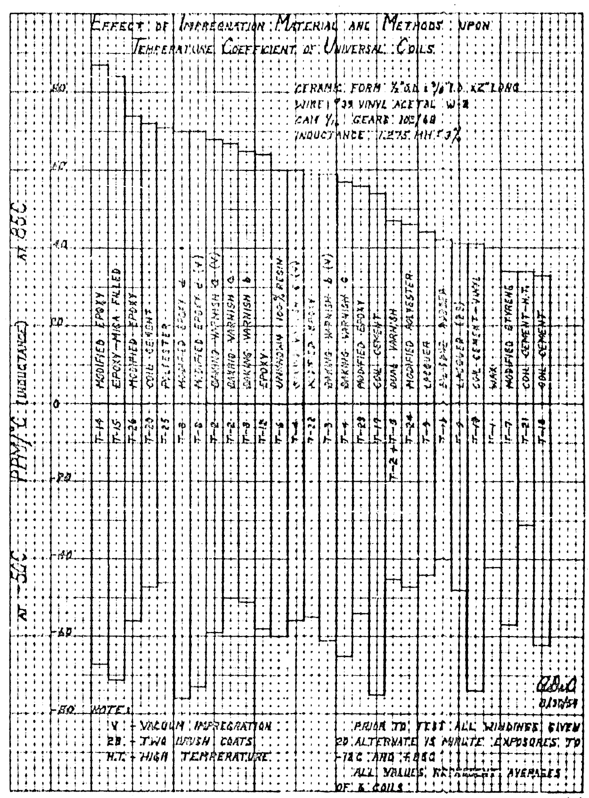
BLOCK DIAGRAM OF TEMPERATURE COEFFICIENT TEST SET UP



POOR ORIGINAL



COIL TEST OSCILLATOR
(NEGATIVE TRANSCONDUCTANCE)



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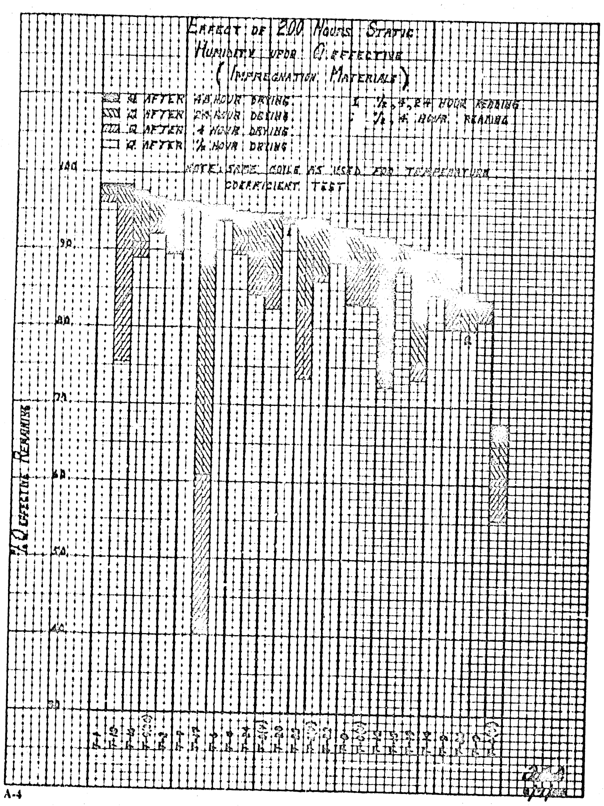
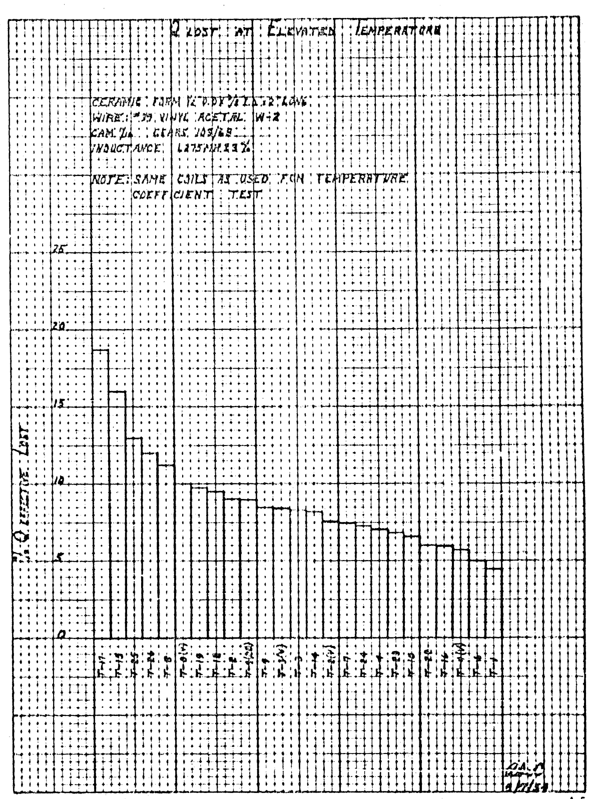
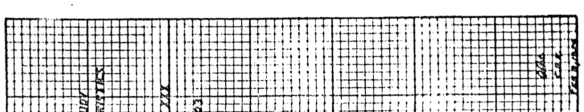


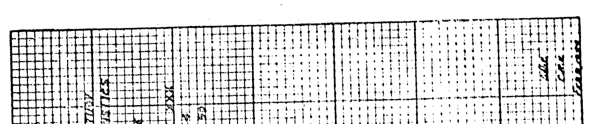
A-4



A-5

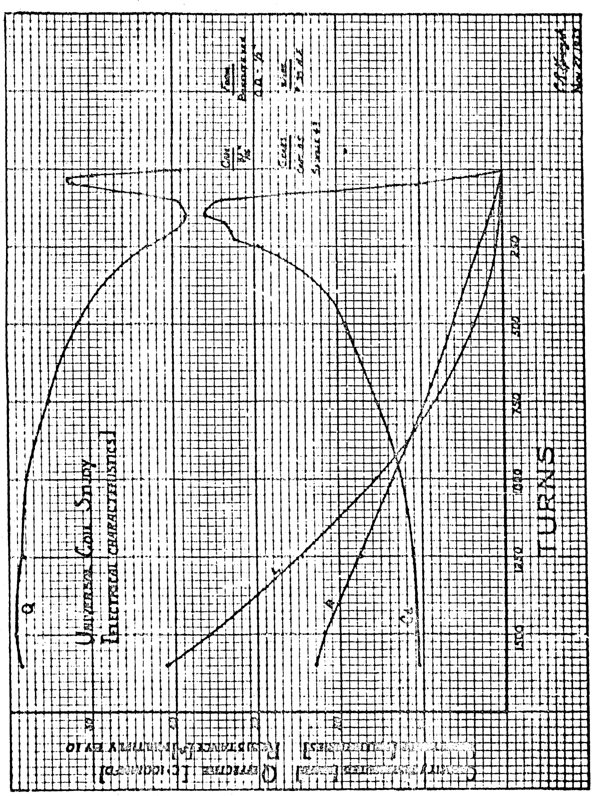


Universal Coil Study
Electrical Characteristics
TURNS

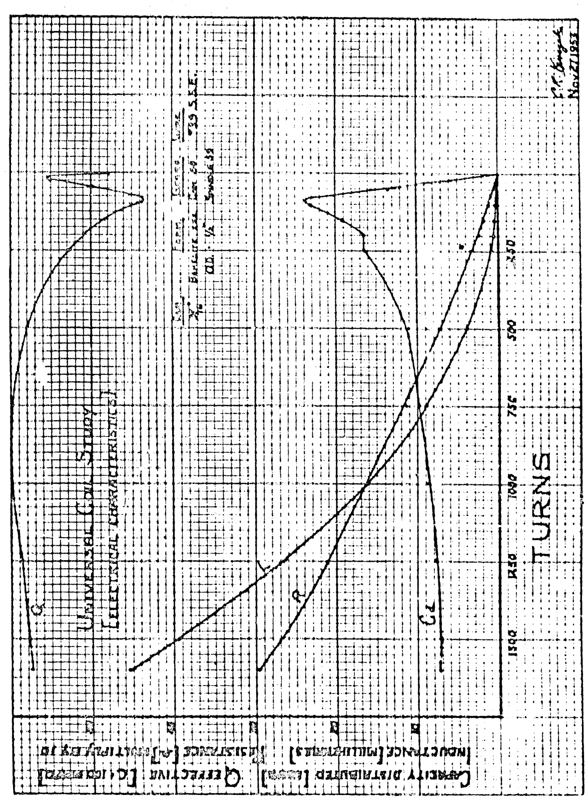


Universal Coil Study
Electrical Characteristics
TURNS

POOR ORIGINAL

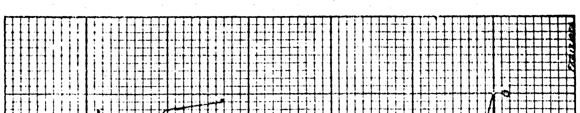


A-8

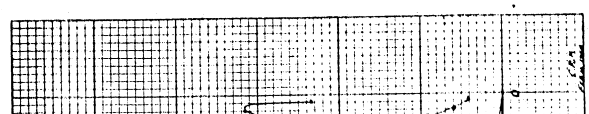


A-9

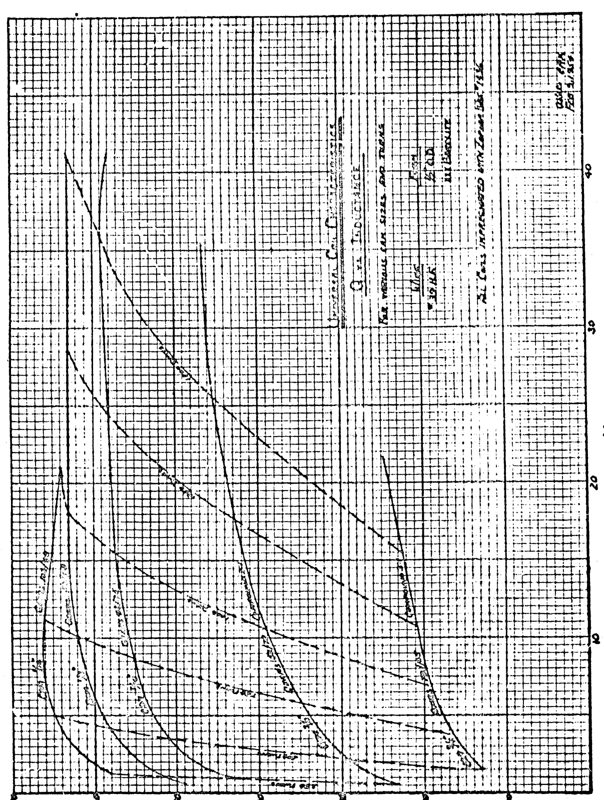
POOR ORIGINAL

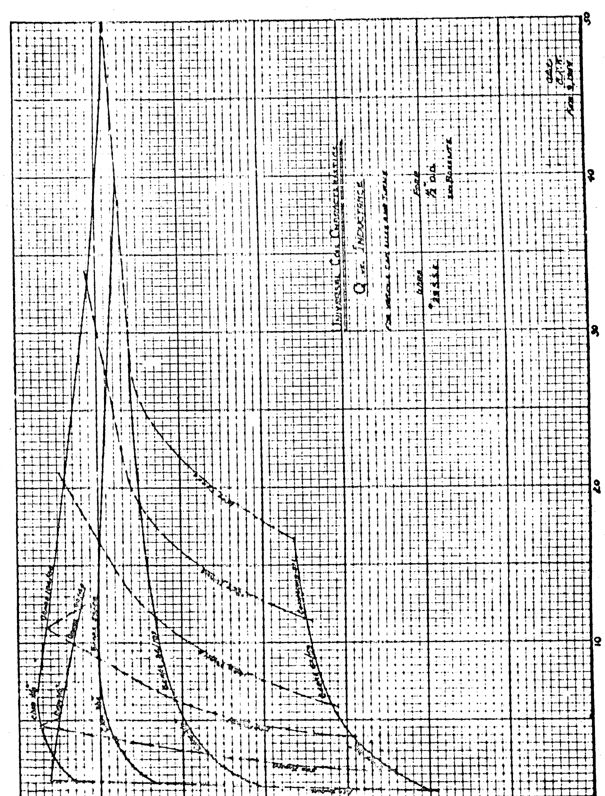


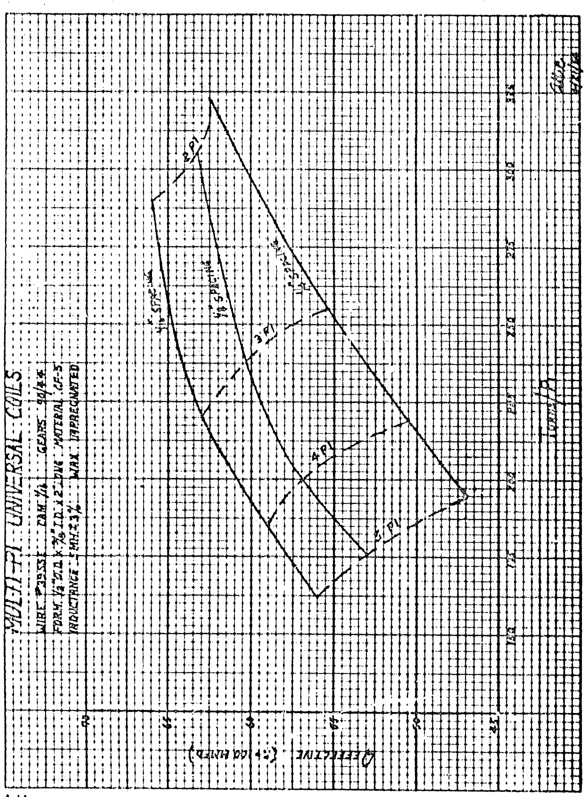
A-10

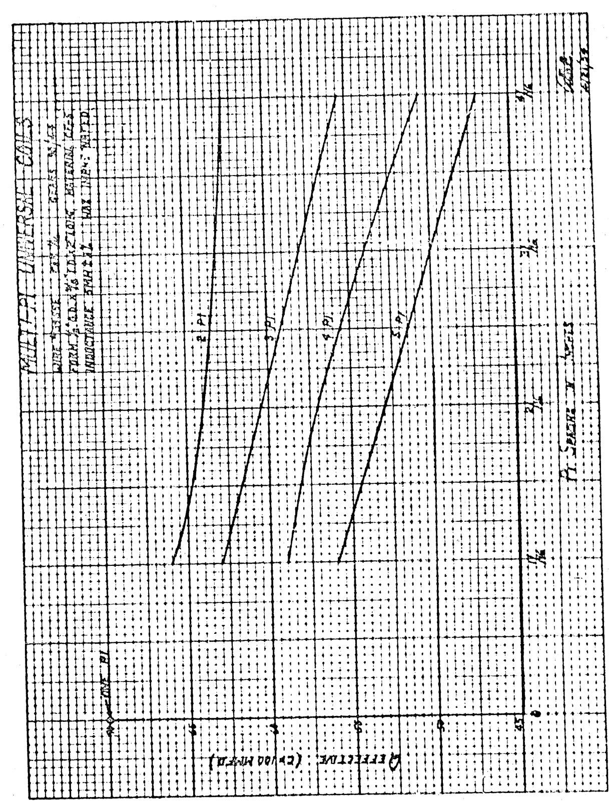


A-11

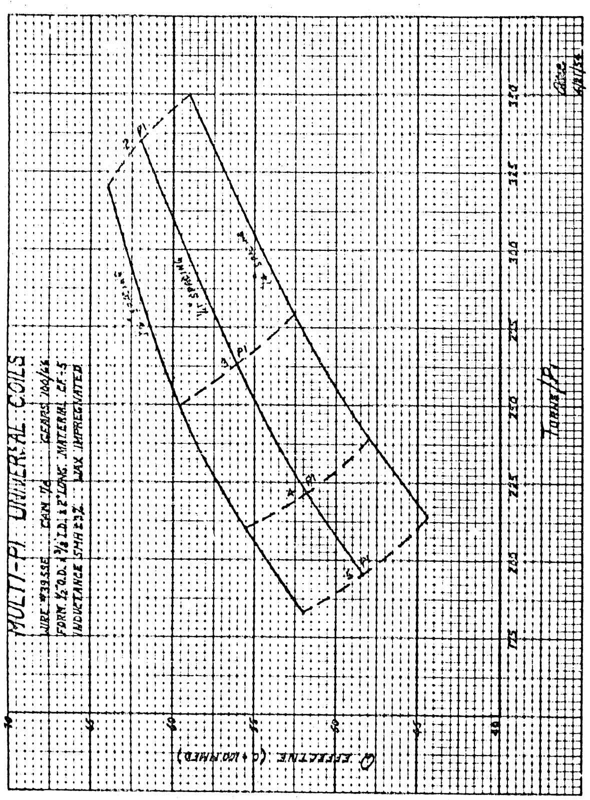


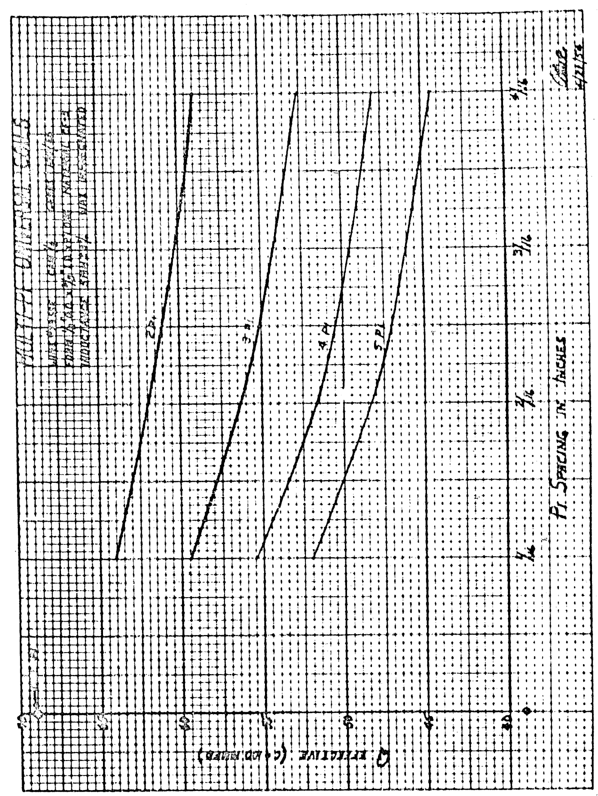



MULTI-PI UNIVERSAL COILS
2 PI
3 PI
4 PI
5 PI


MULTI-PI UNIVERSAL COILS
2 PI
3 PI
4 PI
5 PI

POOR ORIGINAL





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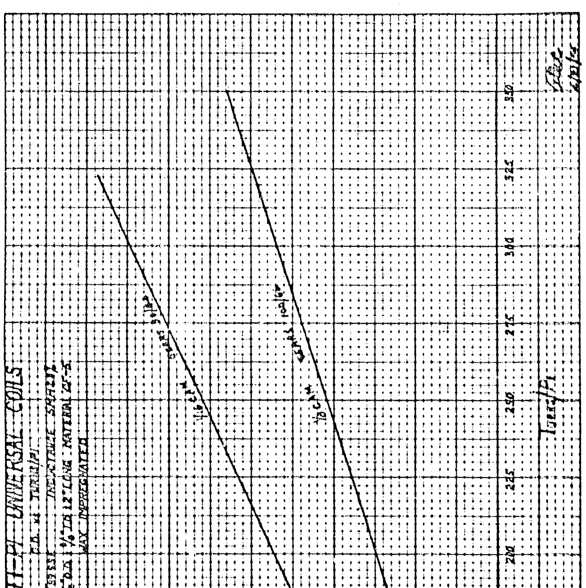

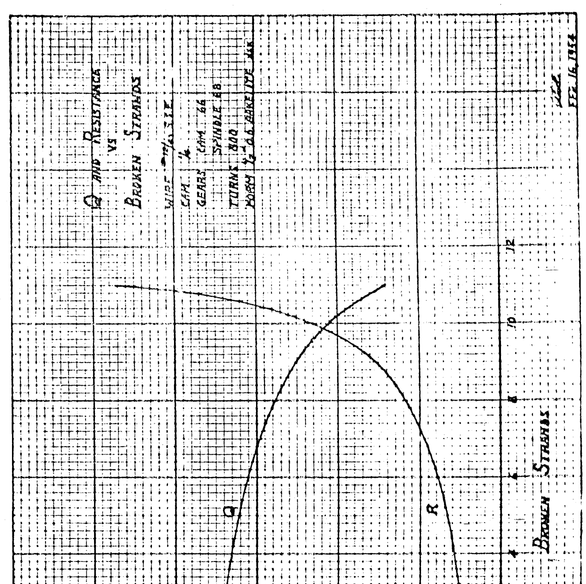

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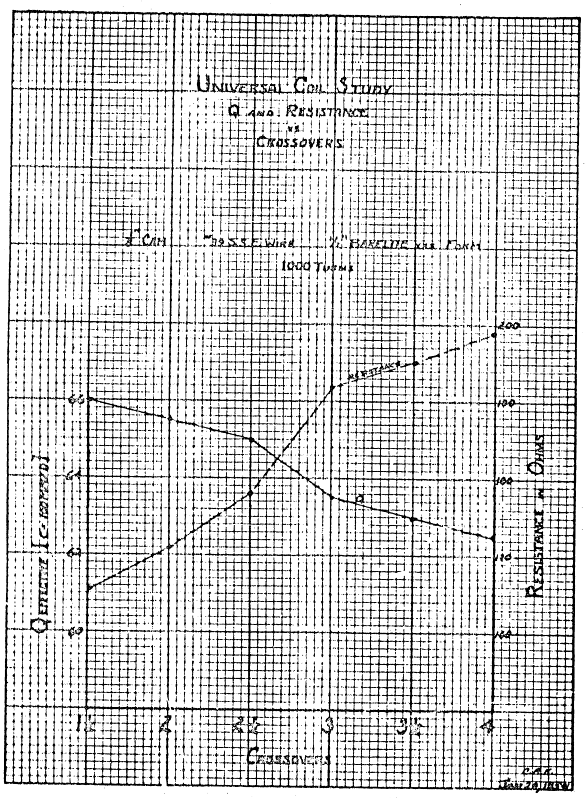
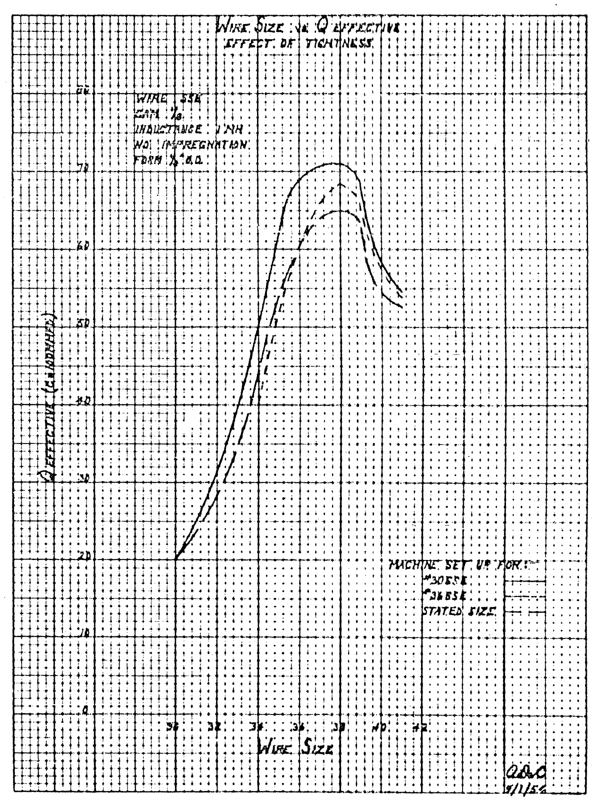
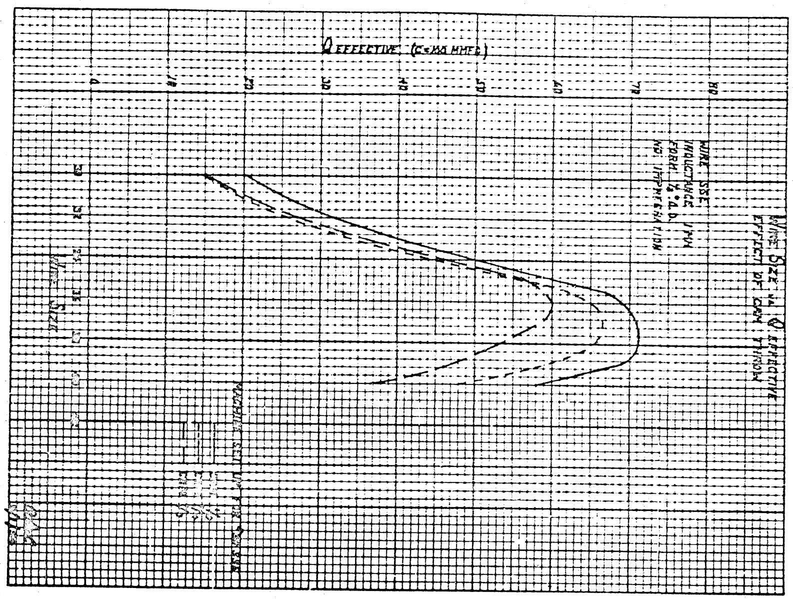






FORM		WIRE		CAM	GEARS			WINDING			L	Q	FREQ.	R	C_d	REMARKS
O.D.	Serial	Type	O.D.		Spindle	Cam	Compound	O.D.	Width	Turns	MH @ 5000 ~	C=100 mmf	Kc	Ω	MMFD	
1 1/2"		39HF	.0043	3/16"	43	85	—	.801	.212	1500	35.9	59.2	80.5	221.9	10.9	SAMPLE DATA SHEET
2 1/2"	xxx	39HF	.0043	3/16"	43	85	—	.804	.212	1500	36.0	59.0	80.1	22[illegible].1	10.3	
3 1/2"	E	39HF	.0043	3/16"	43	85	—	.802	.219	1500	36.1	59.1	80.1	222.1	10.9	
4 1/2"	[illegible]	39HF	.0043	3/16"	43	85	—	.801	.218	1500	35.9	60.0	80.0	221.7	11.0	
5 1/2"	[illegible]	39HF	.0043	3/16"	43	85	—	.802	.219	1500	35.9	59.0	80.0	222.5	10.9	
1 1/2"	[illegible]	39SSE	.0056	3/16"	59	58	—	.992	.200	1500	39.4	57.8	77.9	270.1	7.2	
2 1/2"	E	39SSE	.0056	3/16"	59	58	—	1.004	.201	1500	39.6	57.4	77.8	272.5	9.2	
3 1/2"	K	39SSE	.0056	3/16"	59	58	—	.995	.199	1500	39.6	57.6	77.5	272.[illegible]	7.6	
4 1/2"	A	39SSE	.0056	3/16"	59	58	—	1.002	.198	1500	39.8	57.1	77.3	273.9	7.5	
5 1/2"	B	39SSE	.0056	3/16"	59	58	—	1.024	.202	1500	40.1	57.7	77.1	276.5	7.5	

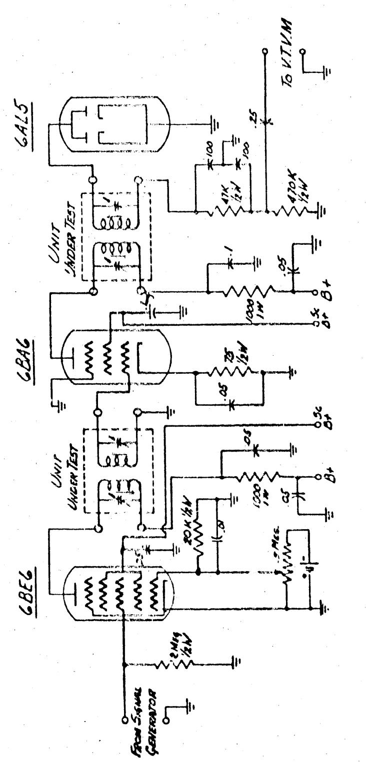
Note: All windings baked for one hour and impregnated in Zophar #1436 wax before readings were taken.

AUTOMATIC MFG. CORP.
Newark, N.J.

Work done by [signature]

Approved [signature]

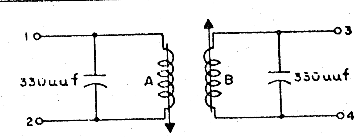
Date 12/10/53



2 STAGE TEST JIG

EXPERIMENTAL 135 KC I.F. TRANSFORMER

WIRE 3/41 S.S.E.
GEARS 42/82
CAM 3/16
TURNS A 480 – B 480
FORM 1/2 O.D 3/8 I.D 3 1/4 LG.
SHIELD CAN 3 1/2 X 2 X 2
CORE 3/8 O.D X 1/2 LG
SK-133, G3

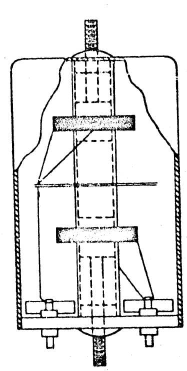


COIL AFTER IMP. WITH CORE

IN CAN	OUT OF CAN
f = 135 KC	f = 135 KC
C = 338	C = 329
Q = 75	Q = 79

COIL AFTER IMP. WITHOUT CORE

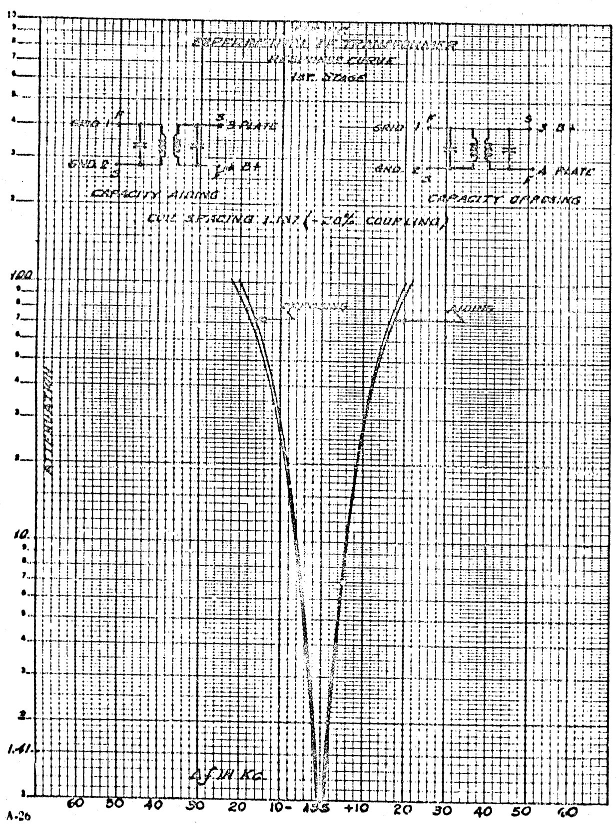
IN CAN	OUT OF CAN
f = 135 KC	f = 135 KC
C = 376	C = 369
Q = 69	Q = 72

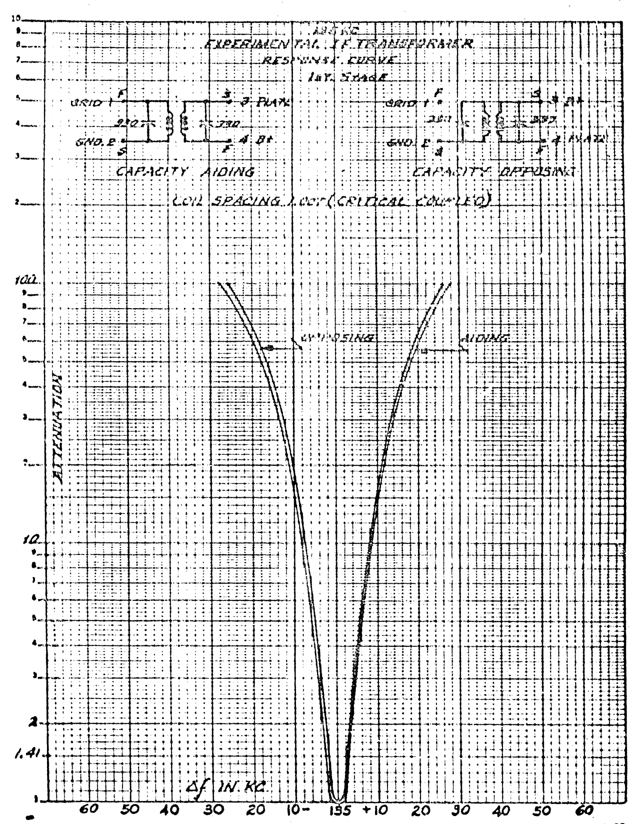


ASSEMBLED TRANSFORMER

NOTE:
TRANSFORMER TUNED ON Q-METER WITH C = 338 uuf i.e.; 330 uuf BASE CAP. + 8 uuf STRAY CAP.
BASE CAP. ARE SILVER MICA

POOR ORIGINAL





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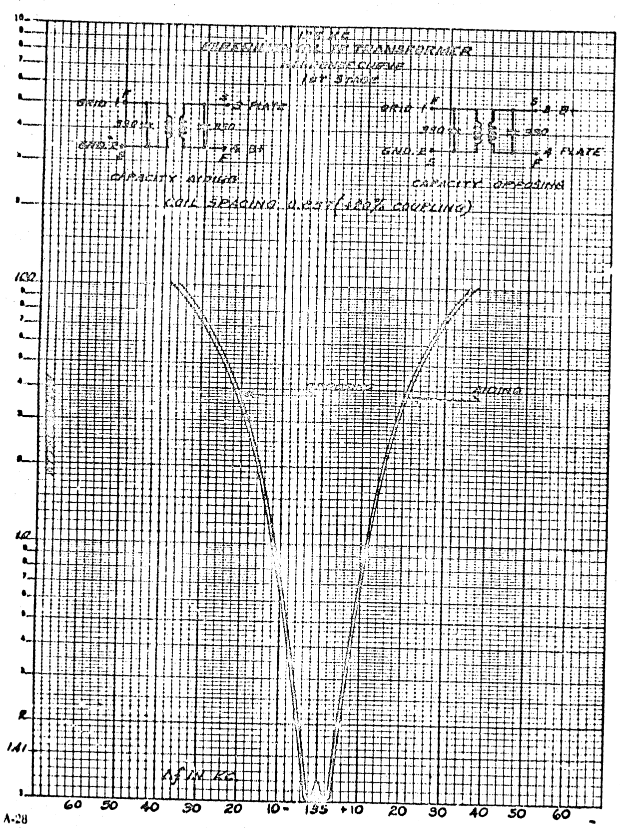
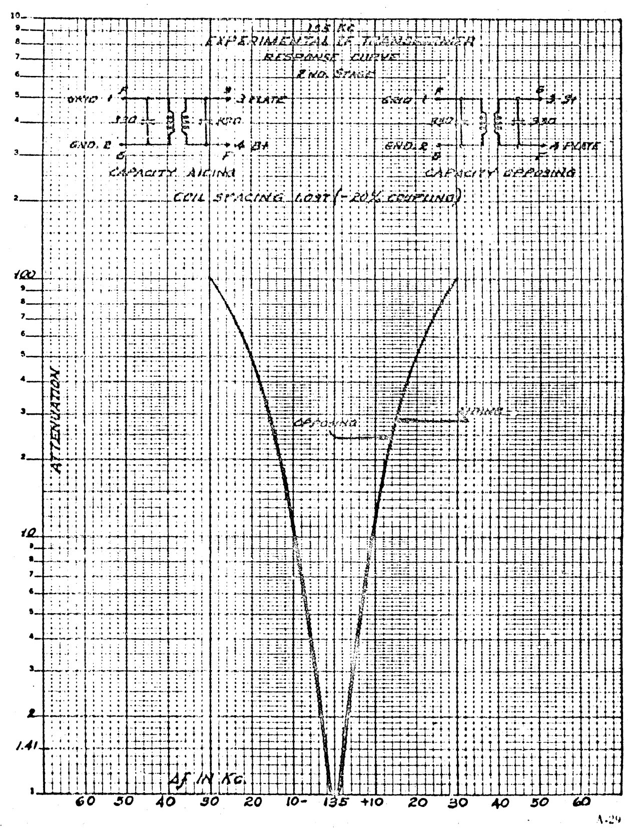
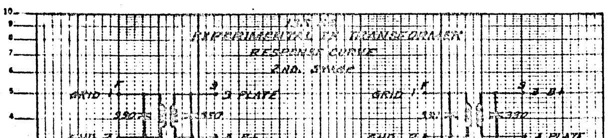
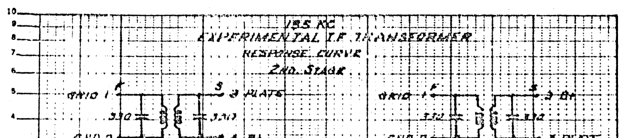

TRANSFORMER
1st STAGE
GRID
PLATE
GND.
CAPACITY AIDING
CAPACITY OPPOSING
AIDING
Δf IN KC.
60 50 40 30 20 10- 135 +10 20 30 40 50 60

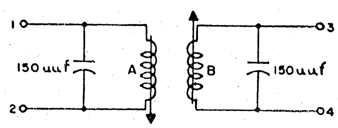

135 KC.
EXPERIMENTAL IF TRANSFORMER
RESPONSE CURVE
2ND. STAGE
GRID
PLATE
GND. 2
CAPACITY AIDING
CAPACITY OPPOSING
COIL SPACING 1.05T (-20% COUPLING)
ATTENUATION
OPPOSING
AIDING
Δf IN KC.
60 50 40 30 20 10- 135 +10 20 30 40 50 60


EXPERIMENTAL I.F. TRANSFORMER
RESPONSE CURVE
2ND. STAGE
GRID 1
3 PLATE
GND. 2
4 B+
CAPACITY AIDING
GRID 1
3 B+
GND. 2
4 PLATE
CAPACITY OPPOSING
COIL SPACING
OPPOSING
AIDING
Δf IN KC
60 50 40 30 20 10 135 +10 20 30 40 50 60
100
10
2
1.41


135 KC
EXPERIMENTAL I.F. TRANSFORMER
RESPONSE CURVE
2ND. STAGE
GRID 1
3 PLATE
330
GND 2
4 B+
CAPACITY AIDING
GRID 1
3 B+
GND 2
4 PLATE
CAPACITY OPPOSING
COIL SPACING 0.762 (+20 % COUPLING)
ATTENUATION
OPPOSING
AIDING
Δf IN KC
60 50 40 30 20 10 135 +10 20 30 40 50 60
100
10
2
1.41

EXPERIMENTAL 262 KC I.F. TRANSFORMER

WIRE 3/41 S.S.E.
GEARS 42/82
CAM 3/16
TURNS A 350-B 350
FORM 1/2 O.D. 3/8 I.D. 3 1/4 LG.
SHIELD CAN 3 1/2 X 2 X 2
CORE 3/8 O.D. X 1/2 LG.
SK-133 G 3

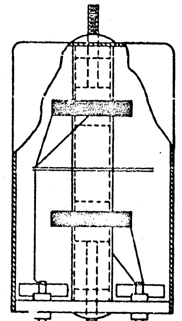


COIL AFTER IMP. WITH CORE

IN CAN	OUT OF CAN
f = 262 KC	f = 262 KC
C = 158	C = 154
Q = 93	Q = 97

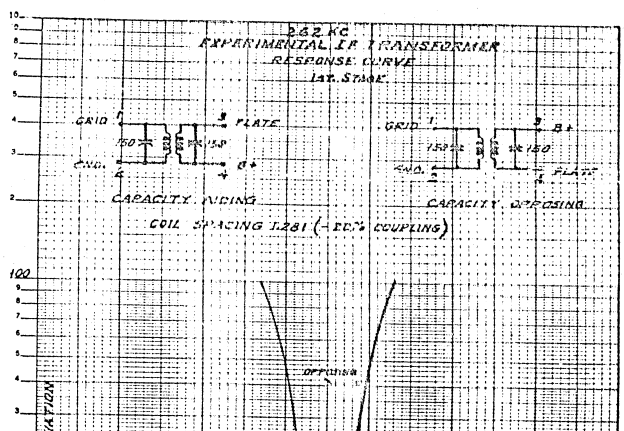
COIL AFTER IMP. WITHOUT CORE

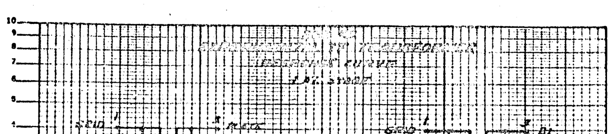
IN CAN	OUT OF CAN
f = 262 KC	f = 262 KC
C = 195	C = 192
Q = 82	Q = 85

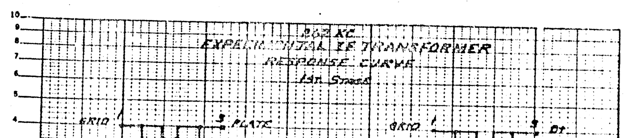


ASSEMBLED TRANSFORMER

NOTE:
TRANSFORMER TUNED ON Q-METER WITH C = 158 uuf ei; 150 uuf BASE CAP. + 8 uuf STRAY CAP.
BASE CAP. ARE SILVER MICA







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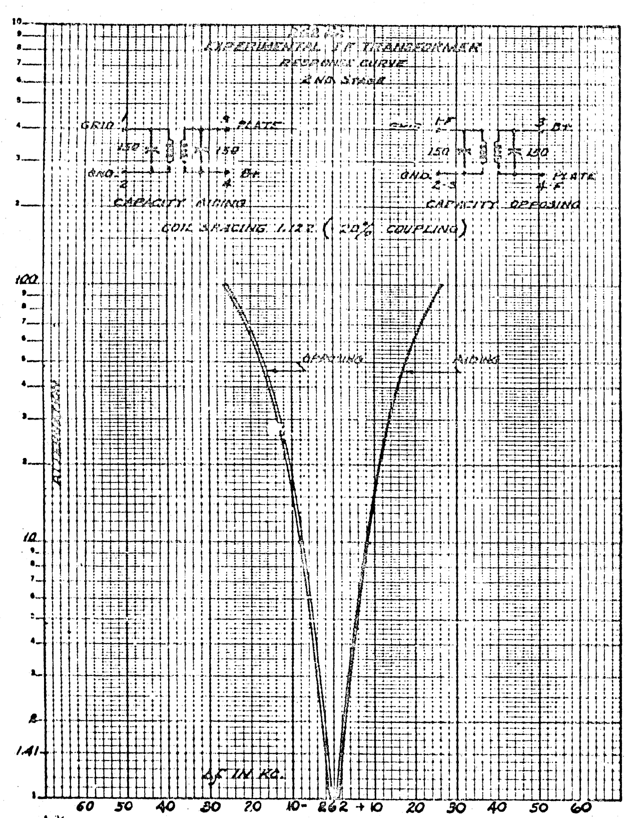
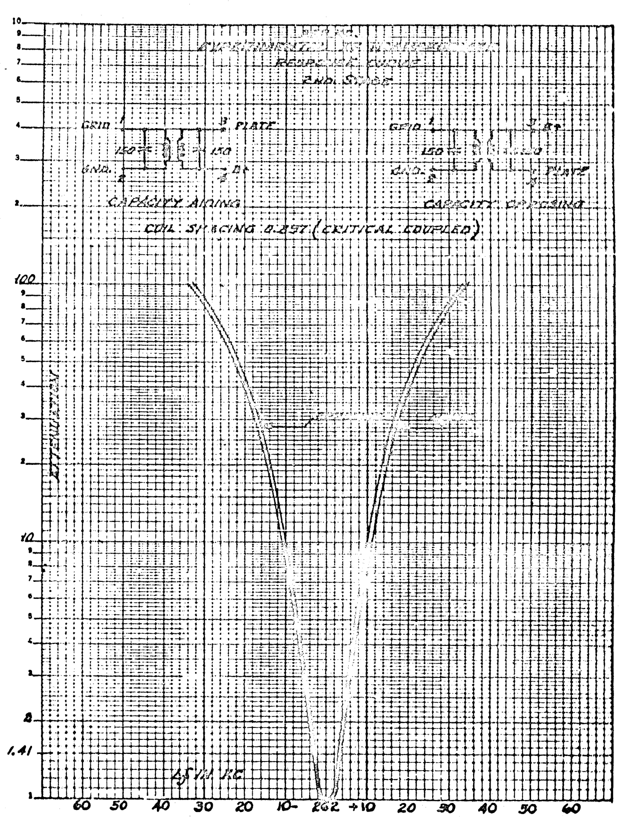
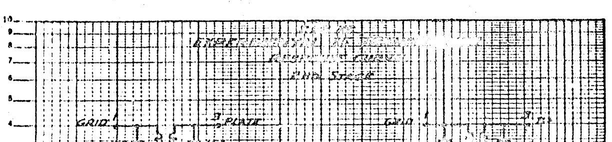
POOR ORIGINAL



A-36

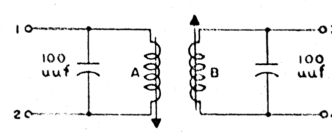


A-37



EXPERIMENTAL 455 KC I.F. TRANSFORMER

WIRE 12/43 S.S.E.
GEARS 81/53
CAM 1/8
TURNS A 215 - B 215
FORM 1/2 O.D. 3/8 I.D. 3 1/4 LG CF-12
SHIELD CAN 3 1/2 X 2 X 2
CORE 3/8 O.D. X 3/8 LG.
SK-133 GRADE G 3

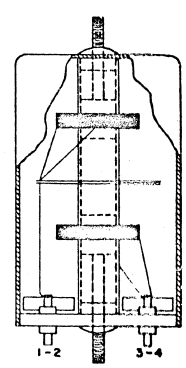


COIL AFTER IMP. WITH CORE

IN CAN	OUT OF CAN
f = 455 KC	f = 455 KC
C = 108.0	C = 108.0
Q = 154	Q = 168

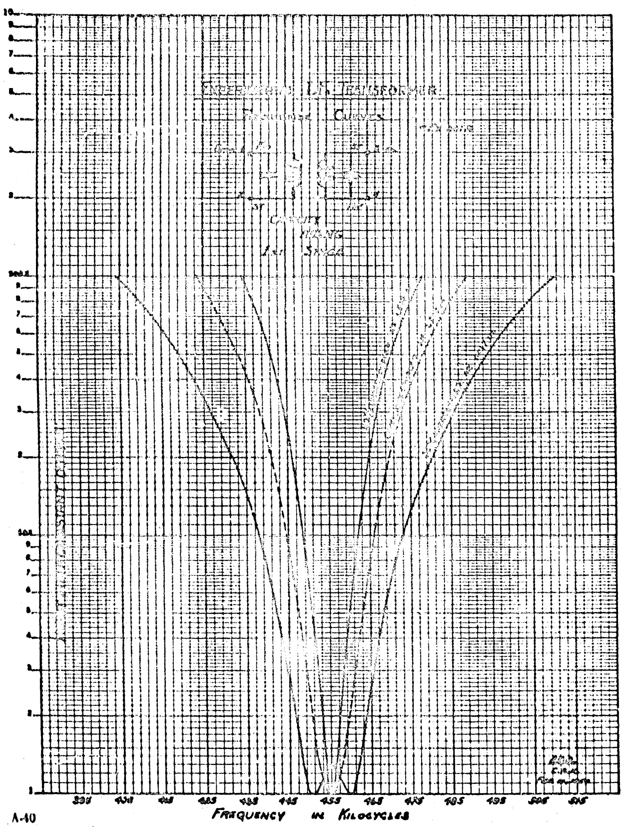
COIL AFTER IMP. WITHOUT CORE

IN CAN	OUT OF CAN
f = 455 KC	f = 455 KC
C = 136.1	C = 131.3
Q = 143	Q = 149



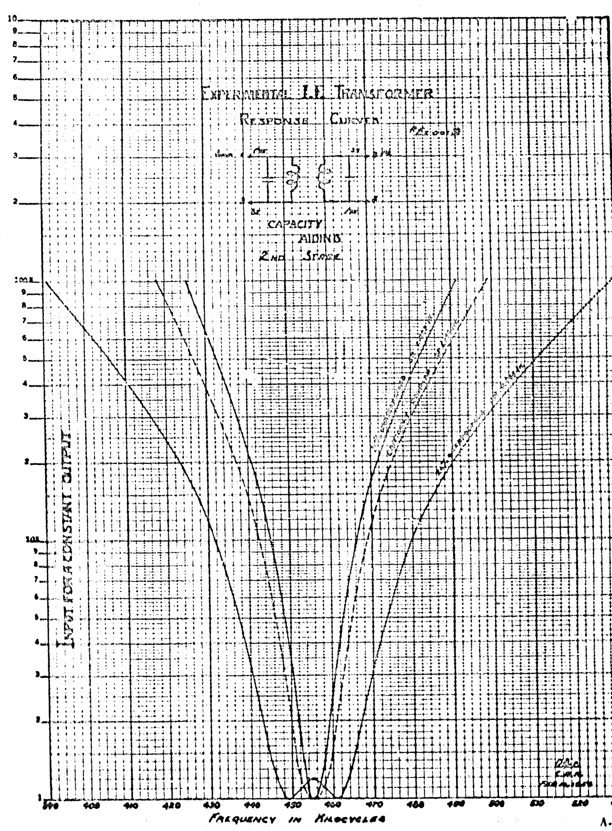
ASSEMBLED TRANSFORMER

NOTE:
TRANSFORMER TUNED ON Q-METER WITH C = 100 uuf
[illegible] 100 uuf BASE CAP.
+ 8 uuf STRAY CAP.
BASE CAP. ARE SILVER MICA



CAPACITY
FREQUENCY IN KILOCYCLES

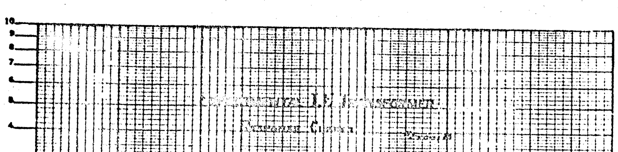
A-40



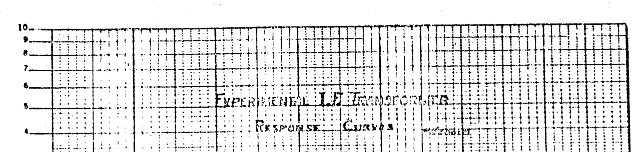
EXPERIMENTAL I.F. TRANSFORMER
RESPONSE CURVES
CAPACITY
AIDING
2ND STAGE
INPUT FOR A CONSTANT OUTPUT
FREQUENCY IN KILOCYCLES

A-41

POOR ORIGINAL

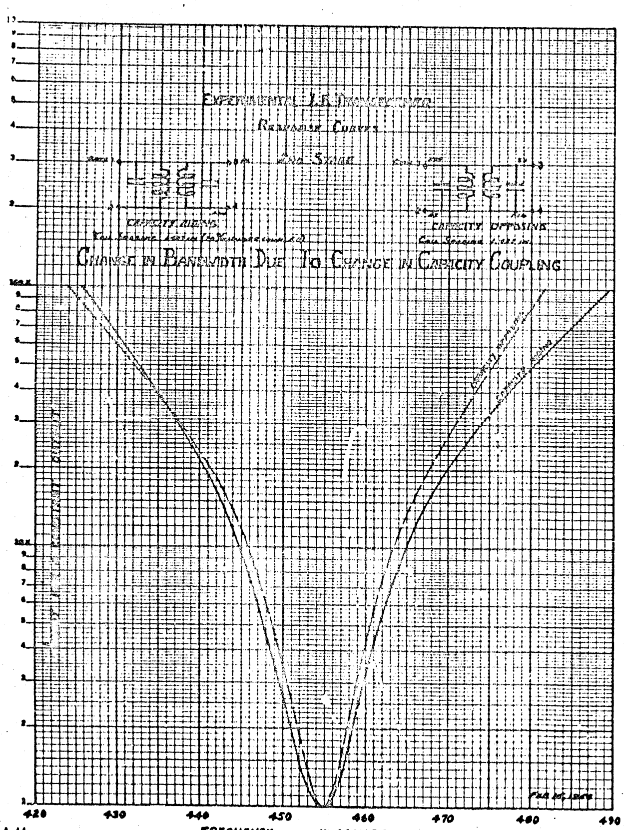


A-42



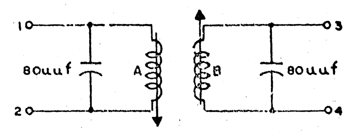
A-43

POOR ORIGINAL



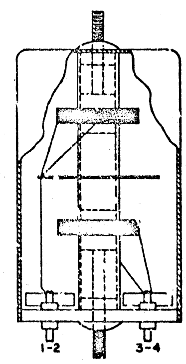
EXPERIMENTAL 1400 KC I.F. TRANSFORMER

WIRE 3/41 S.S.E.
GEARS 28/58
CAM 5/32
TURNS A 75 – B 75
FORM 1/2 O.D. 3/8 I.D. 3 1/4 LG
SHIELD CAN 3 1/2 X 2 X 2
CORE 3/8 O.D. X 1/2 LG
SK-133, G3



COIL AFTER IMP. WITH CORE	
IN CAN	OUT OF CAN
f = 1400 KC	f = 1400 KC
C = 88	C = 87
Q = 84	Q = 84

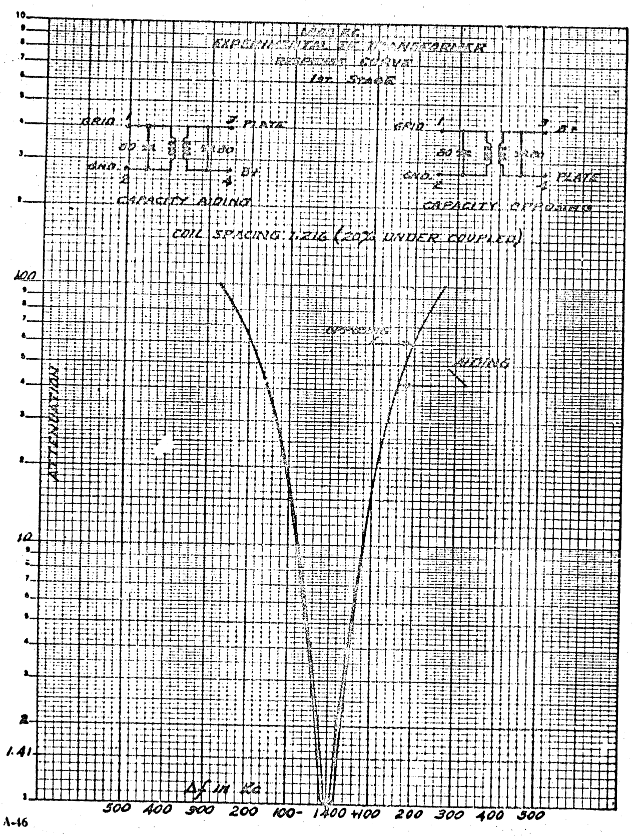
COIL AFTER IMP. WITHOUT CORE	
IN CAN	OUT OF CAN
f = 1400 KC	f = 1400 KC
C = 125	C = 123
Q = 74	Q = 75



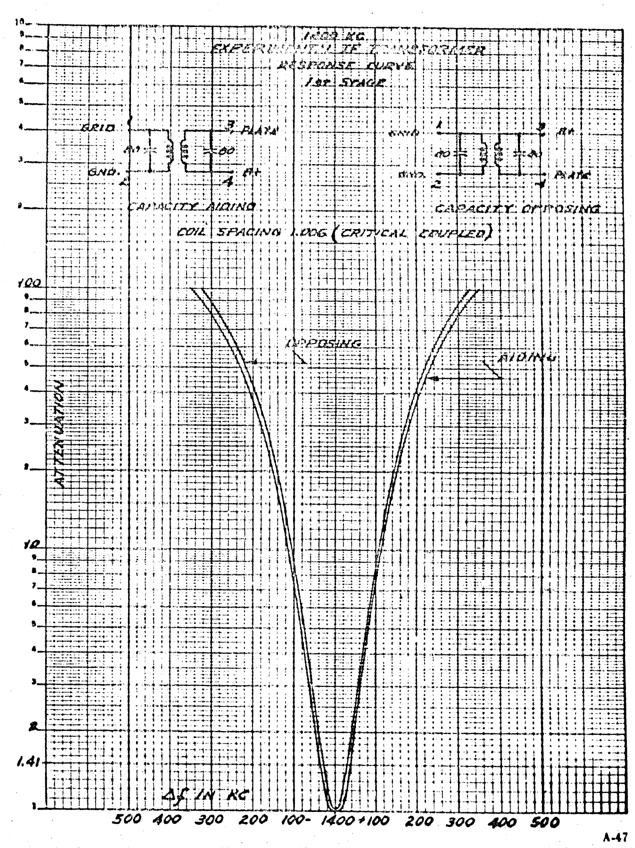
ASSEMBLED TRANSFORMER

NOTE:
TRANSFORMER TUNED ON Q-METER WITH C = 88 uuf.
i.e.; 80 uuf BASE CAP.
+ 8 uuf STRAY CAP.
BASE CAP. ARE SILVER MICA

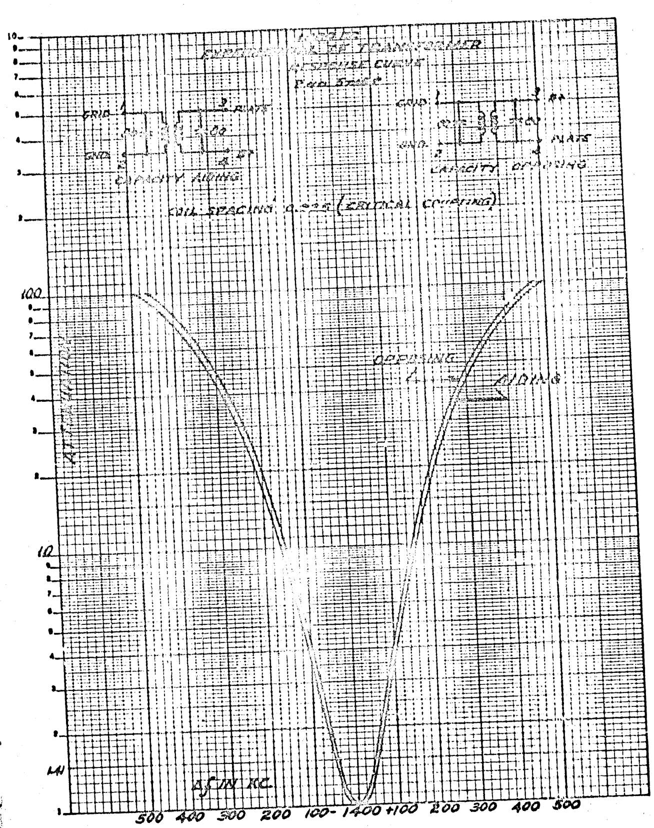
POOR ORIGINAL

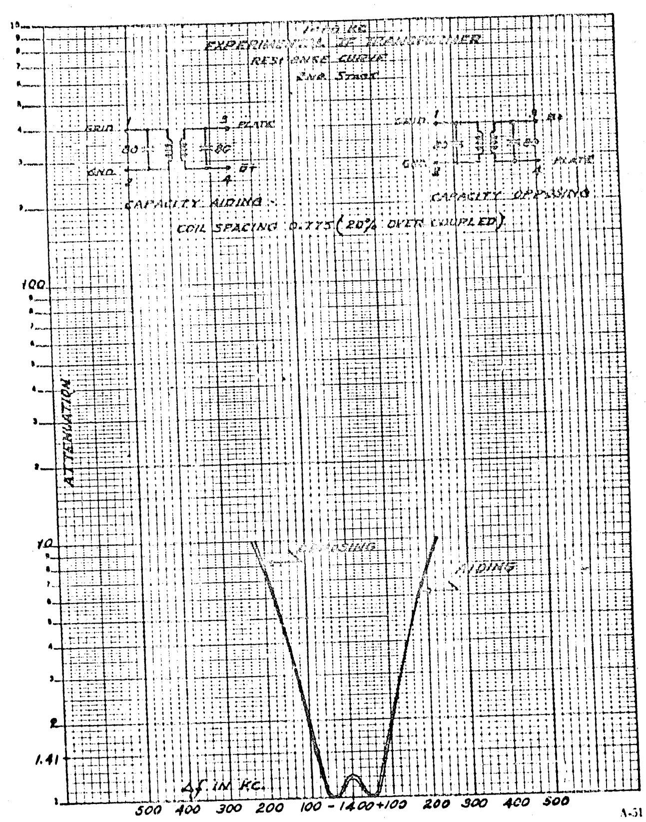


A-46



A-47

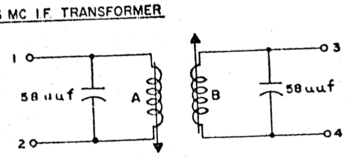

RESPONSE CURVE
GRID
PLATE
GND.
B+
CAPACITY AIDING
CAPACITY OPPOSING
COIL SPACING
(CRITICAL COUPLING)
100
10
1.41
ATTENUATION
OPPOSING
AIDING
Δf IN KC.
500 400 300 200 100 - 1400 +100 200 300 400 500


EXPERIMENTAL IF TRANSFORMER
RESPONSE CURVE
2ND. STAGE
GRID
PLATE
GND.
B+
CAPACITY AIDING
CAPACITY OPPOSING
COIL SPACING 0.775 (20% OVER COUPLED)
100
10
2
1.41
ATTENUATION
AIDING
Δf IN KC.
500 400 300 200 100 - 1400 +100 200 300 400 500

POOR ORIGINAL

EXPERIMENTAL 4.3 MC I.F. TRANSFORMER

WIRE 5/44 SSE
GEARS 101/65
CAM 1/8
TURNS A 26 - B 26
FORM 1/2 O.D. 3/8 I.D. 3 1/4 LG.
SHIELD CAN 3 1/2 X 2 X 2
CORE 3/8 O.D. X 1/2 LG.
SK-133, G 3

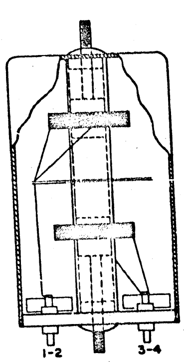


COIL AFTER IMP. WITH CORE

IN CAN	OUT OF CAN
f = 4.3 MC	f = 4.3 MC
C = 66	C = 65
Q = 82	Q = 83

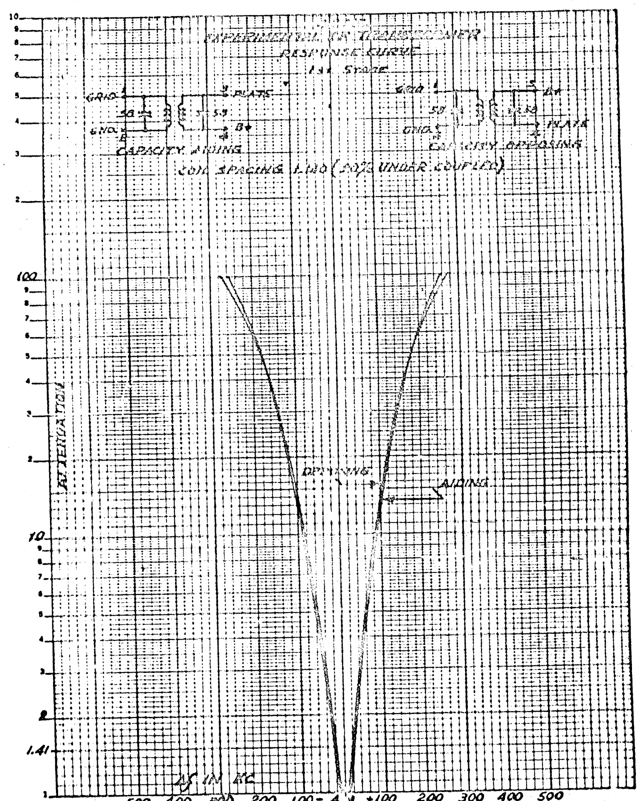
COIL AFTER IMP. WITHOUT CORE

IN CAN	OUT OF CAN
f = 4.3 MC	f = 4.3 MC
C = 97	C = 96
Q = 67	Q = 68

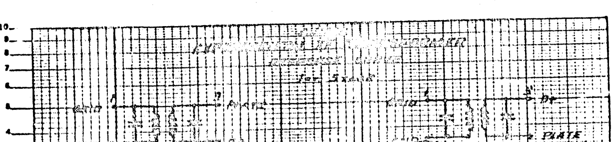


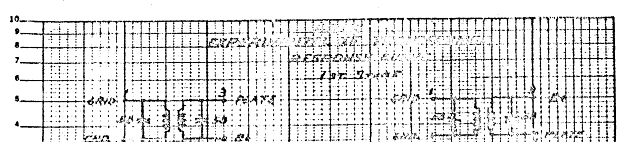
ASSEMBLED TRANSFORMER

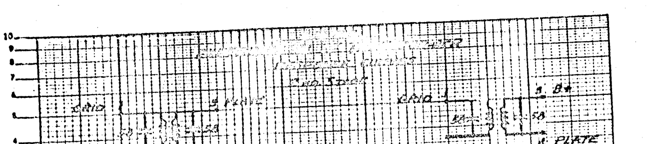
NOTE:
TRANSFORMER TUNED ON Q-METER WITH C = 66 uuf, ie; 58 uuf BASE CAP. + 8 uuf STRAY CAPACITY. BASE CAP. ARE SILVER MICA

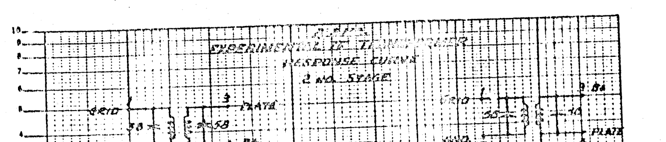


POOR ORIGINAL









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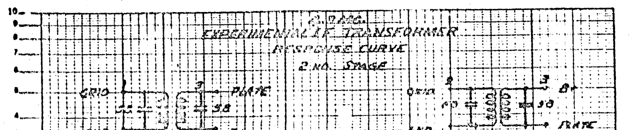
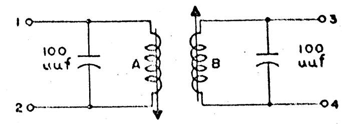


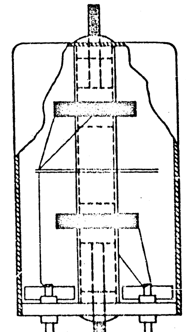
EXPERIMENTAL 455 KC I.F. TRANSFORMER

WIRE 12/43 S.S.E.
GEARS 81/53
CAM 1/8
TURNS A 215 - B 215
FORM 1/2 O.D. 3/8 I.D. 3 1/4 LG. CF-12
SHIELD CAN 3 1/2 X 2 X 2
CORE 3/8 O.D. X 3/8 LG.
SK-133 GRADE G 3



COIL AFTER IMP. WITH CORE	
IN CAN	OUT OF CAN
f = 455 KC	f = 455 KC
C = 108.0	C = 108.0
Q = 154	Q = 168

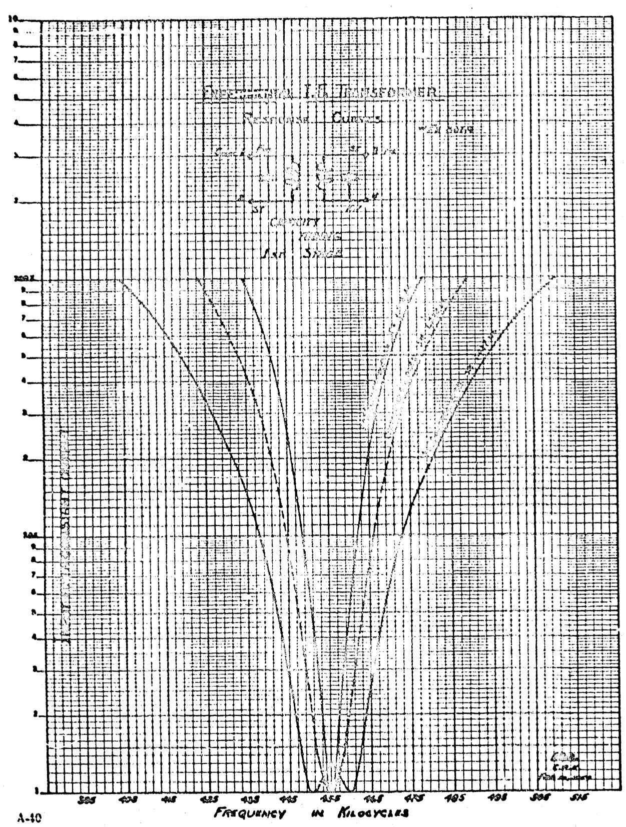
COIL AFTER IMP. WITHOUT CORE	
IN CAN	OUT OF CAN
f = 455 KC	f = 455 KC
C = 136.1	C = 131.3
Q = 143	Q = 149



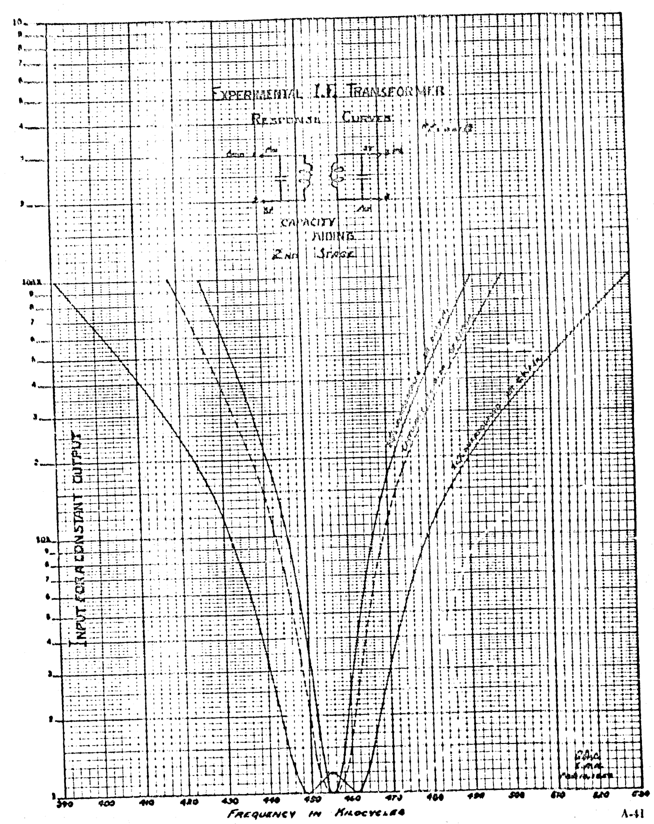
ASSEMBLED TRANSFORMER

NOTE:
TRANSFORMER TUNED ON
Q-METER WITH C = 100 uuf
μ; 100 uuf BASE CAP.
+ 8 uuf STRAY CAP.
BASE CAP. ARE SILVER MICA

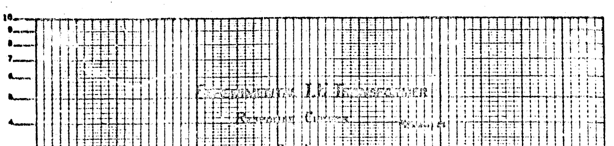
POOR ORIGINAL



A-40

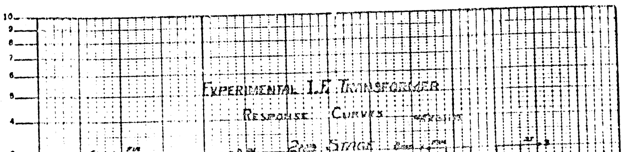


A-41



Change in Bandwidth Due To Change in Capacity Coupling
Frequency in Kilocycles

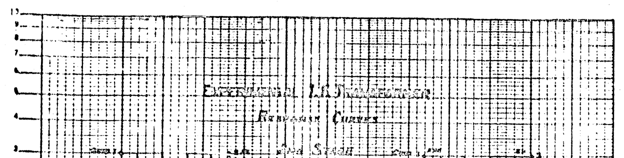
A-42



Experimental I.F. Transformer
Response Curves
2nd Stage
Change in Bandwidth Due To Change in Capacity Coupling
Input for a Constant Output
410
425
440
455
470
485
500
Frequency in Kilocycles

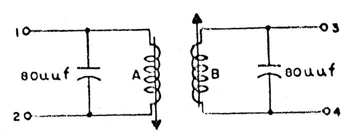
A-44

POOR ORIGINAL



EXPERIMENTAL 1400 KC I.F. TRANSFORMER

WIRE 3/41 S.S.E.
GEARS 28/58
CAM 5/32
TURNS A 75 – B 75
FORM 1/2 O.D. 3/8 I.D. 3 1/4 LG
SHIELD CAN 3 1/2 X 2 X 2
CORE 3/8 O.D. X 1/2 LG
SK-133,03

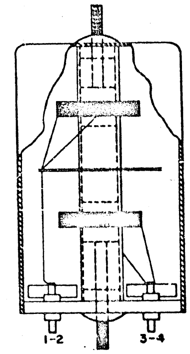


COIL AFTER IMP. WITH CORE

IN CAN	OUT OF CAN
f = 1400 KC	f = 1400 KC
C = 88	C = 87
Q = 84	Q = 84

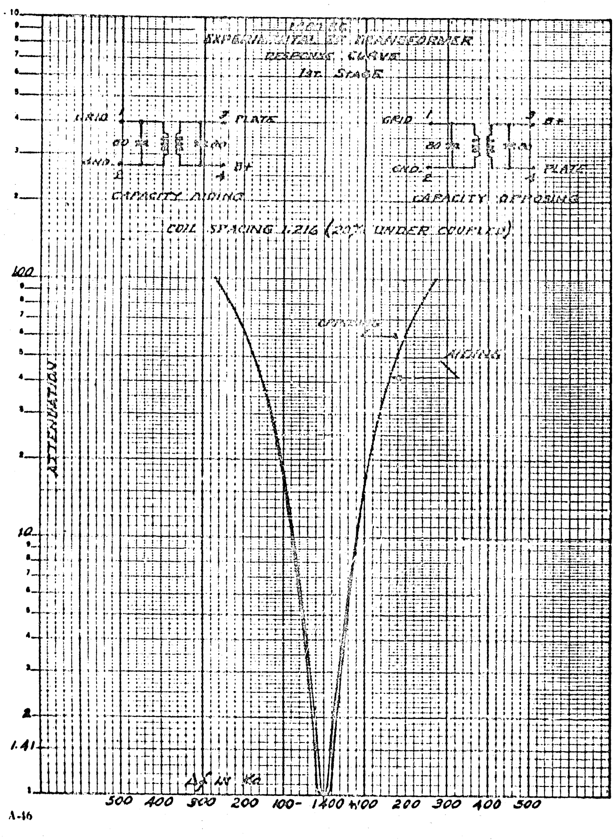
COIL AFTER IMP. WITHOUT CORE

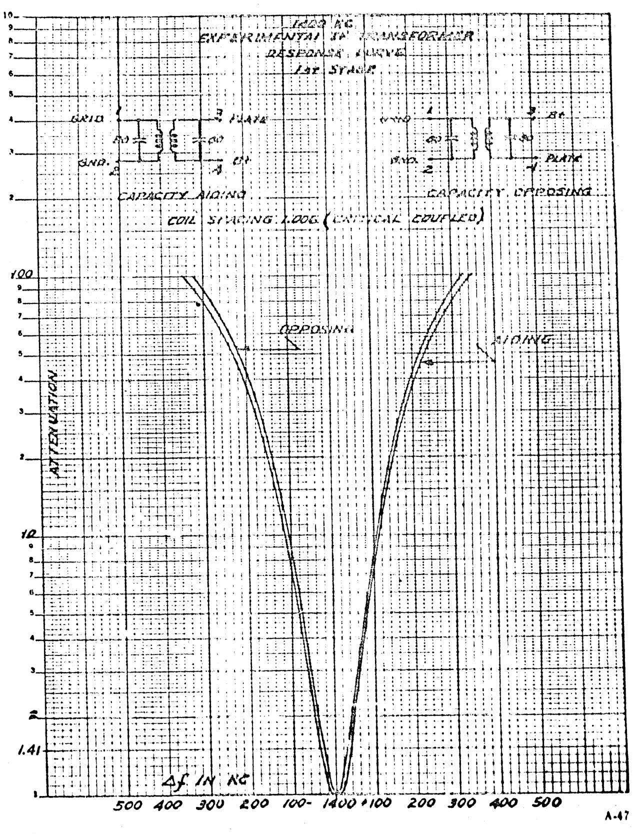
IN CAN	OUT OF CAN
f = 1400 KC	f = 1400 KC
C = 125	C = 123
Q = 74	Q = 75



ASSEMBLED TRANSFORMER

NOTE:
TRANSFORMER TUNED ON Q-METER WITH C = 88 uuf. i.e. 80 uuf BASE CAP. + 8 uuf STRAY CAP.
BASE CAP. ARE SILVER MICA





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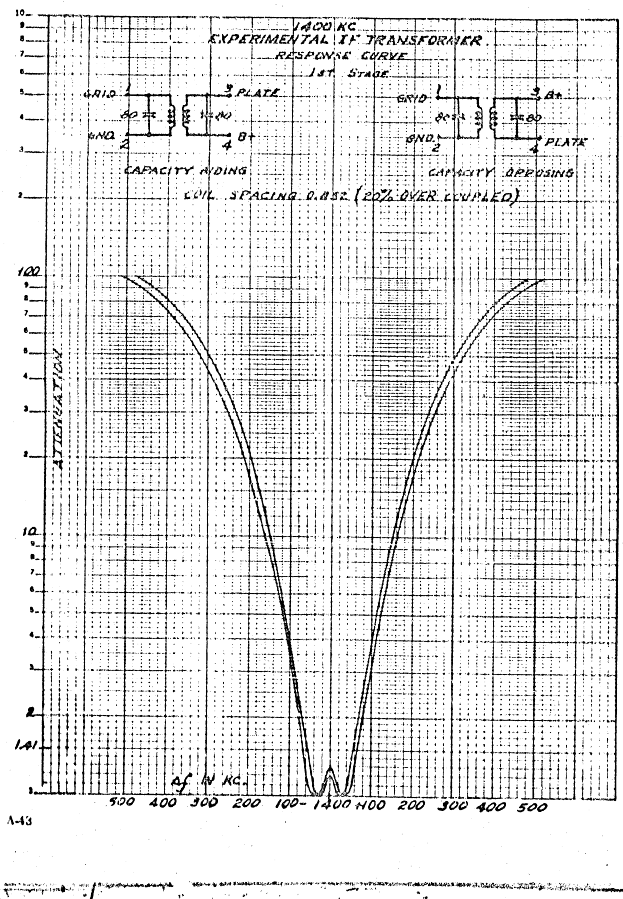
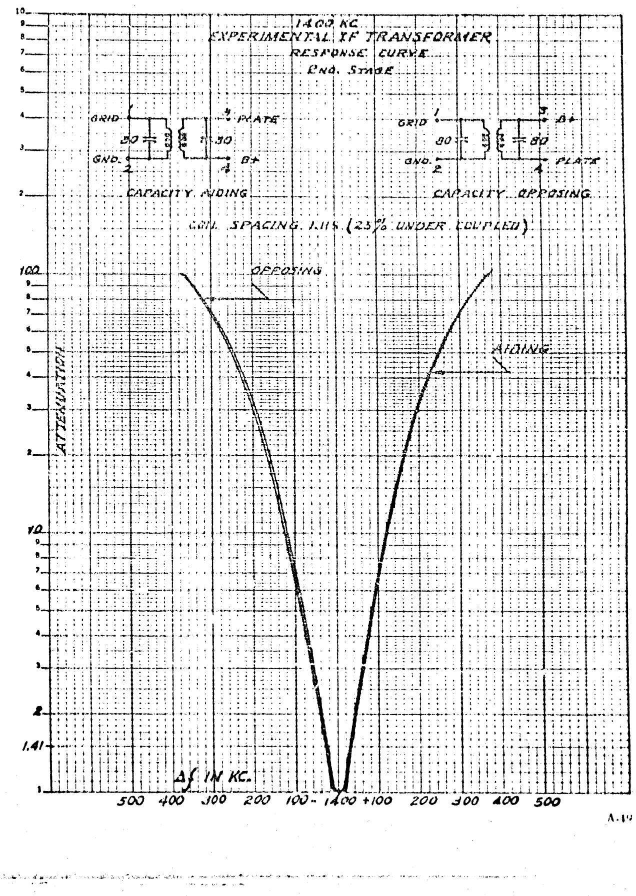


1400 KC
EXPERIMENTAL IF TRANSFORMER
RESPONSE CURVE
1ST. STAGE
GRID
PLATE
GND.
B+
CAPACITY AIDING
CAPACITY OPPOSING
COIL SPACING 0.85" (20% OVER COUPLED)
ATTENUATION
Δf IN KC.
500 400 300 200 100- 1400 +100 200 300 400 500

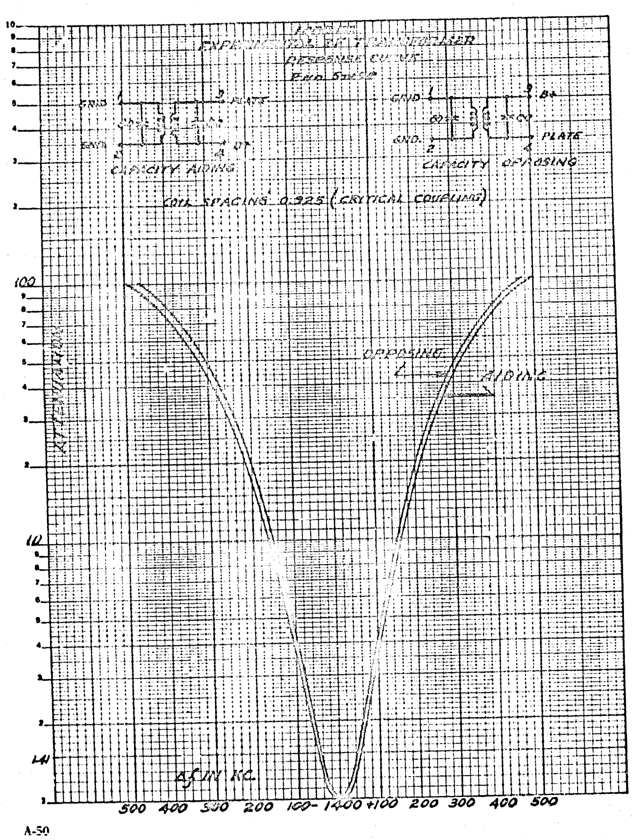
A-43

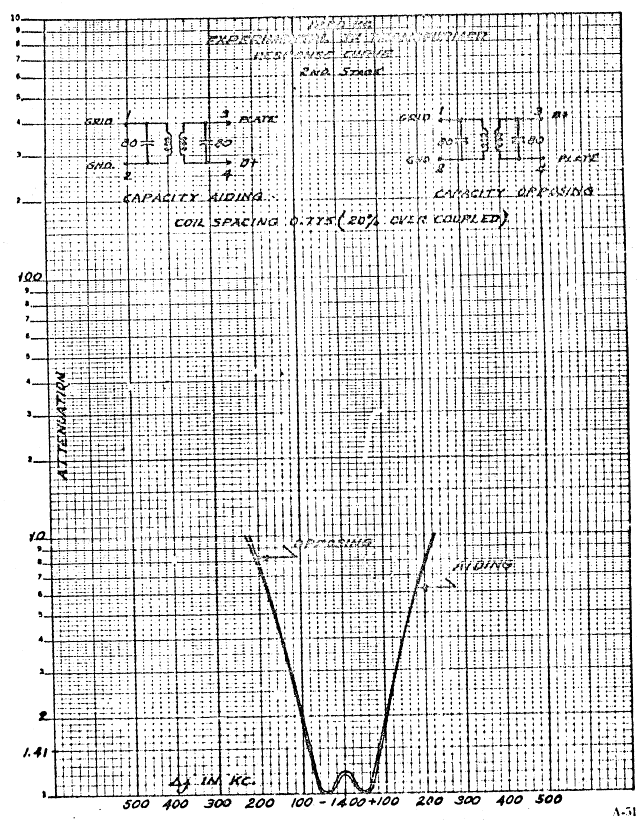


1400 KC
EXPERIMENTAL IF TRANSFORMER
RESPONSE CURVE
2ND. STAGE
GRID
PLATE
GND.
B+
CAPACITY AIDING
CAPACITY OPPOSING
COIL SPACING 1.115 (25% UNDER COUPLED)
OPPOSING
AIDING
ATTENUATION
Δf IN KC.
500 400 300 200 100- 1400 +100 200 300 400 500

A-44

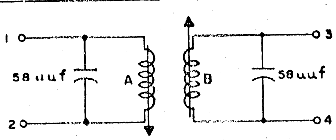
POOR ORIGINAL





EXPERIMENTAL 4.3 MC I.F. TRANSFORMER

WIRE 5/44 SSE
GEARS 101/66
CAM 1/8
TURNS A 26 - B 26
FORM 1/2 O.D. 3/8 I.D. 3 1/4 LG.
SHIELD CAN 3 1/2 X 2 X 2
CORE 3/8 O.D. X 1/2 LG.
SK-133, G3

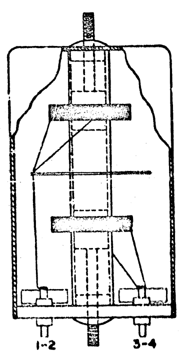


COIL AFTER IMP. WITH CORE

IN CAN	OUT OF CAN
f = 4.3 MC	f = 4.3 MC
C = 66	C = 65
Q = 82	Q = 83

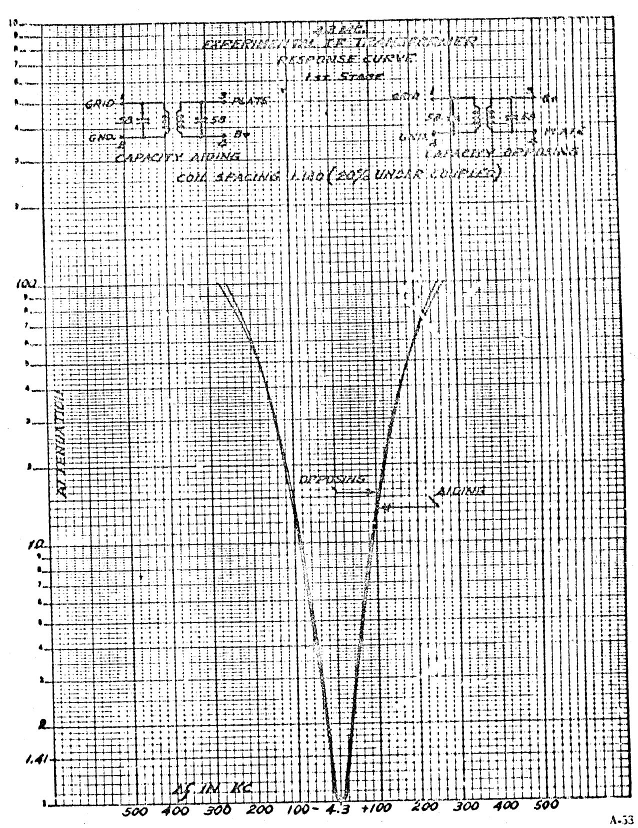
COIL AFTER IMP. WITHOUT CORE

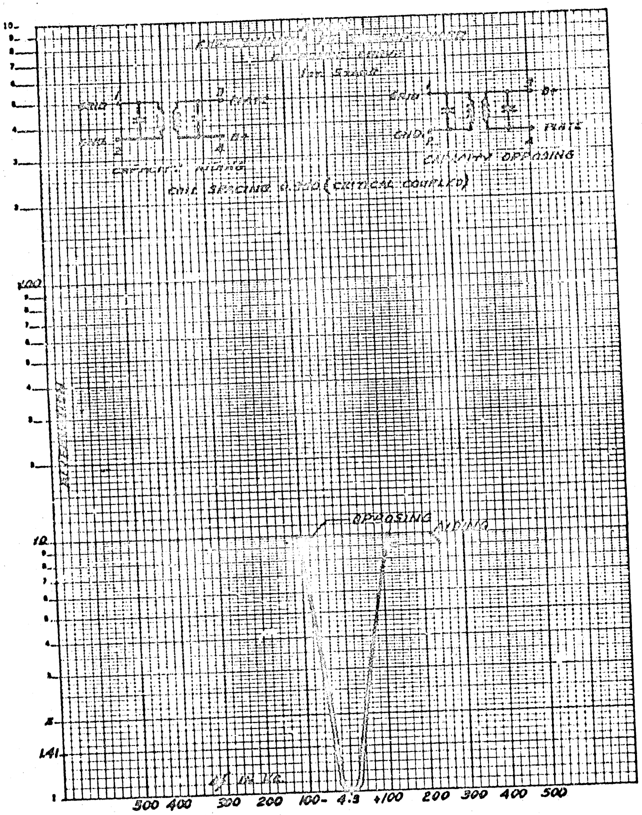
IN CAN	OUT OF CAN
f = 4.3 MC	f = 4.3 MC
C = 97	C = 96
Q = 67	Q = 68

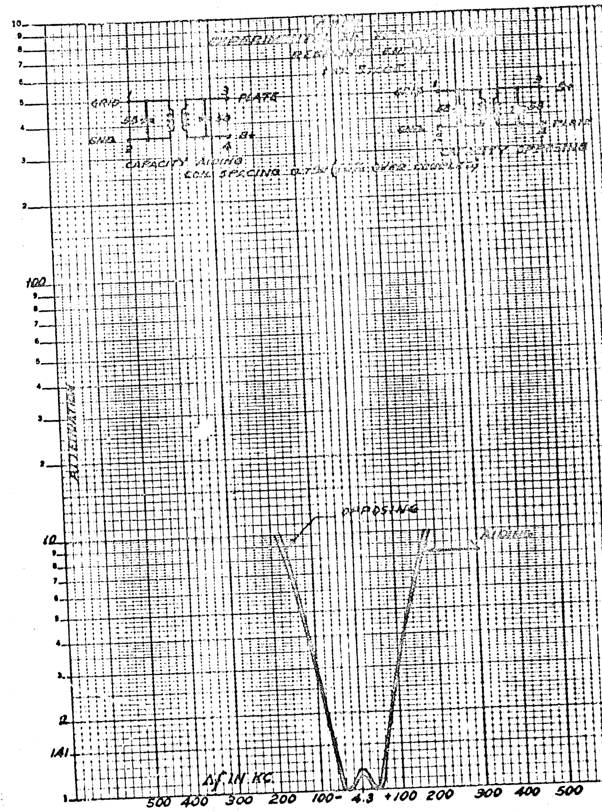


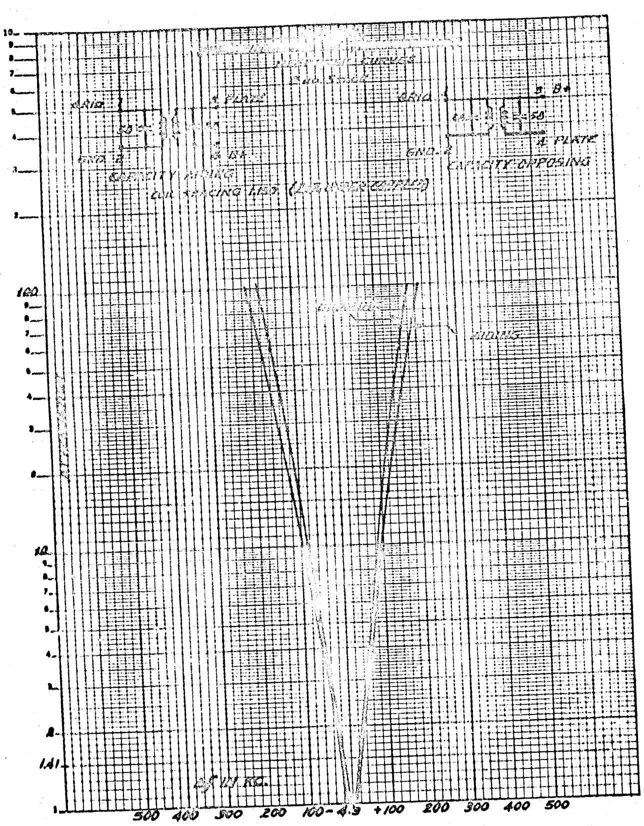
ASSEMBLED TRANSFORMER

NOTE:
TRANSFORMER TUNED ON Q-METER WITH C = 66 uuf, ai; 58 uuf BASE CAP. + 8 uuf STRAY CAPACITY. BASE CAP. ARE SILVER MICA



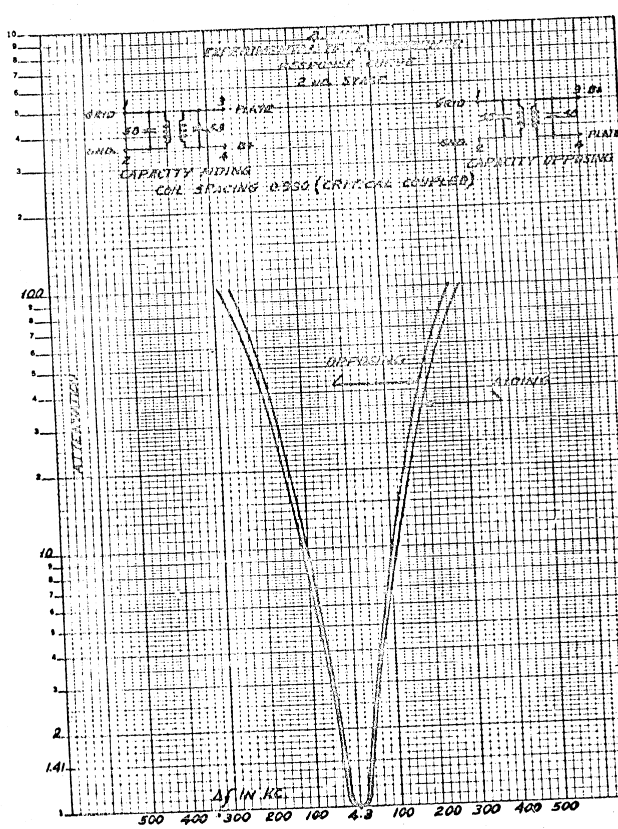

1st STAGE
GRID
PLATE
B+
GND
CAPACITY OPPOSING
COIL SPACING
(CRITICAL COUPLED)
OPPOSING
AIDING
100
10
1
500 400 300 200 100- 4.3 +100 200 300 400 500


GRID
PLATE
B+
GND
CAPACITY AIDING
COIL SPACING
ATTENUATION
OPPOSING
AIDING
100
10
1
ΔF IN KC
500 400 300 200 100- 4.3 +100 200 300 400 500



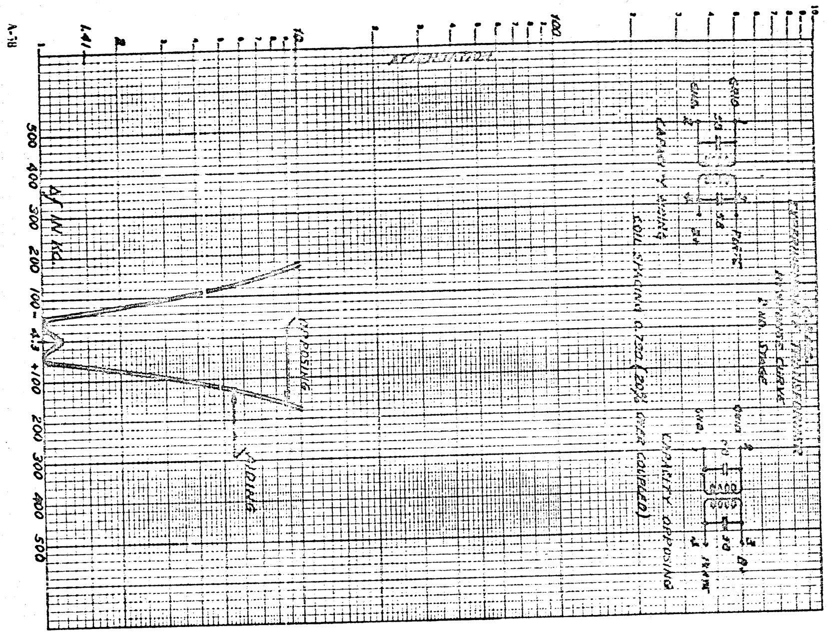
CURVES
GRID
PLATE
B+
GND.
CAPACITY AIDING
COIL SPACING
CAPACITY OPPOSING
AIDING
100
10
1.41
Δf IN KC.
500 400 300 200 100 -4.3 +100 200 300 400 500

A-36



RESPONSE CURVE
2 nd. STAGE
GRID
PLATE
B+
GND.
CAPACITY AIDING
COIL SPACING .0350 (CRITICAL COUPLED)
CAPACITY OPPOSING
OPPOSING
AIDING
ATTENUATION
100
10
1.41
Δf IN KC.
500 400 300 200 100 4.3 100 200 300 400 500

A-37



A-58

WIRE TABLE

AWG SIZE	Diameter, Inches: Min.	Diameter, Inches: Nom.	Diameter, Inches: Max.	Cross Sectional Area: Minimum Cir. Mils	Cross Sectional Area: Nominal Cir. Mils	Cross Sectional Area: Maximum Cir. Mils	Weight, Nominal Wire: Lbs. per 1000 Ft.	Weight, Nominal Wire: Ft. per Lb.	Resistance, Nominal Wire: Ft. per Ohm	Resistance, Nominal Wire: Ohms per Lb.	Resistance, Ohms per 1000 Ft.: Min.	Resistance, Ohms per 1000 Ft.: Nom.	Resistance, Ohms per 1000 Ft.: Max.	AWG SIZE
14	.0634	.0641	.0647	4024.	4107.	4190.	12.44	80.39	396.2	.20290	2.560	2.524	2.478	14
15	.0565	.0571	.0576	3192.	3257.	3322.	9.87	101.32	314.4	.32226	3.249	3.181	3.115	15
16	.0503	.0508	.0513	2531.	2583.	2635.	7.912	128.0	248.8	.5145	4.099	4.018	3.940	16
17	.0448	.0453	.0457	2007.	2048.	2090.	6.213	161.0	197.9	.8135	5.167	5.054	4.943	17
18	.0399	.0403	.0407	1592.	1624.	1657.	4.914	203.6	156.6	1.3001	6.514	6.386	6.263	18
19	.0355	.0359	.03623	1262.	[illegible]	1314.	3.900	256.4	124.3	2.0628	8.231	8.046	7.869	19
20	.0316	.0320	.0323	[illegible]	[illegible]	1042.	[illegible]	322.7	[illegible]	[illegible]	[illegible]	10.128	9.943	20
21	[illegible]	.0285	[illegible]	[illegible]	[illegible]	[illegible]	2.45?	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	21
22	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	22
23	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	1.544	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	23
24	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	412.	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	24
25	.0177	.0179	.0181	[illegible]	320.	327.	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	25
26	.0158	.0159	.0161	249.	254.	259.	.7650	1307.	24.38	53.61	42.07	[illegible]	[illegible]	26
27	.0140	.0142	[illegible]	197.6	202.	205.7	.6101	1639.	19.44	84.31	52.91	51.44	50.71	27
28	.0125	.0126	.0128	156.6	160.	162.9	.4806	2081.	15.31	135.92	66.35	65.31	63.32	28
29	.0111	.0113	.0114	124.3	126.7	129.3	.3866	2587.	12.31	210.15	82.70	81.21	79.78	29
30	.0099	.0100	.0101	98.6	100.5	102.6	.3025	3306.	9.64	342.95	105.83	103.71	101.67	30
31	[illegible]	[illegible]	[illegible]	77.9	79.7	81.5	.2394	4170.	7.639	545.9	133.9	130.9	[illegible]	31
32	[illegible]	[illegible]	[illegible]	[illegible]	63.2	64.8	[illegible]	5263.	6.273	[illegible]	[illegible]	[illegible]	158.1	32
33	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	6553.	4.442	[illegible]	[illegible]	[illegible]	[illegible]	33
34	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	34
35	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	35
36	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	36
37	.0044	.0045	[illegible]	[illegible]	[illegible]	16.5	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	37
38	.0039	.0040	.0041	14.9	15.7	[illegible]	.04944	20644.	2.543	[illegible]	[illegible]	648.2	[illegible]	38
39	.0034	.0035	.0036	11.8	12.5	13.2	.03708	26969.	1.181	22836.	837.1	844.6	800.2	39
40	.0030	.0031	.0032	9.27	9.89	10.5	.02910	34364.	.927	37070.	1152.3	1079.2	1012.8	40
41	.0027	.0028	.0029	7.39	7.84	8.41	.02374	42123.	.7559	55726.	1423.	1323.	1233.	41
42	.0024	.0025	.0026	5.73	6.22	6.73	.01892	52854.	.6028	87691.	1801.	1659.	1534.	42
43	.00212	.00222	.00232	4.50	4.93	5.39	.01465	68259.	.4667	146259.	2352.	2143.	1960.	43
44	.00188	.00198	.00208	3.53	3.91	4.32	.01210	82645.	.3857	214273.	2873.	2593.	2352.	44
45		.00176												45
46		.00157												46
47		.00140												47
48		.00124												48
49		.00111												49
50		.00099												50

Resistance values based on copper of 100% conductivity - 10.371 Ohms (mil ft.) at 20° C.
Values for diameter are those used in commercial practice.

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