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IEEE Recommended Practice for Testing Electronics Transformers and Inductors

Sponsor

Electronics Transformer Technical Committee
of the
IEEE Power Electronics Society

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Abstract: A number of tests are presented for use in determining the significant parameters and performance characteristics of electronics transformers and inductors. These tests are designed primarily for transformers and inductors used in all types of electronics applications, but they may apply to the other types of transformers of large apparent-power rating used in the electric power utility industry.
Keywords: common-mode rejection tests, corona tests, current transformer tests, electronic inductors, electronic power transformers, inductance measurements, inrush-current evaluation, insulation tests, large rectifiers, noise tests, product rating, pulse transformers, quality factor, resistance tests, self-resonance, temperature rise tests, terminated impedance measurements, transformer capacitance, voltage-time shielding

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Introduction

(This introduction is not part of IEEE Std 389-1996, IEEE Recommended Practice for Testing Electronics Transformers and Inductors.)

This recommended practice has been prepared to serve as a guide in the design, testing, and specifying of electronics transformers and inductors. This document contains many tests and experimental methods for evaluating almost every aspect of electronics transformer performance, including a number of tests for determining transformer environmental characteristics such as audible-noise generation. The tests and specifications included are aimed primarily at the testing and evaluation of transformers of relatively low apparent-power rating, such as those used in communications, instrumentation, control, small appliances, and computer applications. However, most of these tests are perfectly applicable to transformers of any rating. A useful feature of this recommended practice is the listing, in clause 4, of all standard tests used in the specification of a transformer. This clause will provide a useful starting point for many users of this recommended practice.

MKS units (Standard International or SI units) are used throughout this recommended practice; equivalent CGS units are sometimes given where their usage is still common practice. Definitions and symbols are in accordance with those of the International Electrotechnical Commission (IEC) wherever possible.

The Electronics Transformer Technical Committee (ETTC) wishes to acknowledge its indebtedness to those who have so freely given of their time and knowledge in the development of the original version of this recommended practice. The fellowship of authors of the inaugural publication, IEEE Std 389-1979, includes the following distinguished members:

Paul K. Goethe, *Chair*

J. Adams
R. P. Carey
M. I. Distefano
R. C. Fischer
P. C. Hill
H. W. Lord
H. S. Mitsanas
J. Tardy
H. I. Tillinger
A. A. Toppeto

This recommended practice was prepared by the Working Group on Transformer Tests of the Test Codes Subcommittee of the Electronics Transformer Technical Committee of the IEEE Power Electronics Society, which was comprised of the following membership:

Robert B. Beers, *Chair*

J. S. Andresen
I. D. Bolt
M. Carey
J. Cronk
J. DeCramer
C. J. Elliott
P. K. Goethe
W. D. Goethe
R. R. Grant
H. E. Lee
D. N. Ratliff
R. L. Sell
J. Silgailis
B. Thackwray
M. A. Wilkowski
H. Yarpezeshkan

When the Electronics Transformer Technical Committee balloted and approved this recommended practice, the membership was as follows:

Rohn R. Grant, *Chair*

J. S. Andresen
R. B. Beers
M. Carey
J. Cronk
J. DeCramer
C. J. Elliott
P. K. Goethe
W. D. Goethe
D. N. Ratliff
R. L. Sell
J. Silgailis
B. Thackwray
M. A. Wilkowski*
H. Yarpezeshkan

*Current ETTC Chair

When the IEEE Standards Board approved this recommended practice on 20 June 1996, it had the following membership:

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IEEE Recommended Practice for Testing Electronics Transformers and Inductors

1. Overview

1.1 Scope

This recommended practice presents a number of tests for use in determining the significant parameters and performance characteristics of electronics transformers and inductors. These tests are designed primarily for transformers and inductors used in all types of electronics applications, but they may apply to the other types of transformers of large apparent-power rating used in the electric power utility industry. Some of the tests described are intended for qualifying a product for a specific application, while others are test practices used widely for manufacturing and customer acceptance testing. Clause 4 is intended to serve as a guide for particular application categories.

The tests described in this recommended practice include those most commonly used in the electronics transformer industry: electric strength, resistance, power loss, inductance, impedance, balance, ratio of transformation, and many others used less frequently.

1.2 Transformers and inductors

The following are the specific types of transformers and inductors to which this recommended practice is applicable:

a) Electronic power
    1) Power
    2) Isolating
    3) Current limiting
    4) Rectifier
    5) Combination (rectifier and filament)
    6) Ferroresonant
    7) Converter
    8) Polyphase
    9) Switch mode
    10) Magnetic amplifiers
b) Large rectifiers
c) Pulse

1) Voltage stepdown
2) Voltage stepup
3) Low ratio inverting
4) Low power pulse
5) Square-loop

d) Broadband
1) Impedance matching
2) DC insulating
3) Common-mode rejection
4) Potential transformers
5) Current transformers
6) Filter inductors
7) Charging inductors
8) Hybrid transformers

2. References

This recommended practice shall be used in conjunction with the following publications. When the following standards are superseded by an approved revision, the revision shall apply.

ANSI S1.4-1983, Specification for Sound Level Meters.[1]

IEEE Std 4-1995, IEEE Standard Techniques for High-Voltage Testing (ANSI).[2]

IEEE Std 100-1996, IEEE Standard Dictionary of Electrical and Electronics Terms.

IEEE Std 111-1984, IEEE Standard for Wide-Band Transformers (ANSI).[3]

IEEE Std 119-1974, IEEE Recommended Practice for General Principles of Temperature Measurement as Applied to Electrical Apparatus.[4]

IEEE Std 260.1-1993, IEEE Standard Letter Symbols for Units of Measurement (SI Units, Customary Inch-Pound Units, and Certain Other Units) (ANSI).

IEEE Std 272-1970 (Reaff 1976), IEEE Standard for Computer-Type (Square-Loop) Pulse Transformers.[5]

IEEE Std 280-1985, IEEE Standard Letter Symbols for Quantities Used in Electrical Science and Electrical Engineering (ANSI).

IEEE Std 295-1969 (Reaff 1993), IEEE Standard for Electronics Power Transformers (ANSI).

IEEE Std 315-1975 (Reaff 1993), IEEE Standard Graphic Symbols for Electrical and Electronics Diagrams (ANSI/DoD).

[1]ANSI publications are available from the Sales Department, American National Standards Institute, 11 West 42nd Street, 13th Floor, New York, NY 10036, USA.

[2]IEEE publications are available from the Institute of Electrical and Electronics Engineers, 445 Hoes Lane, P.O. Box 1331, Piscataway, NJ 08855-1331, USA.

[3]IEEE Std 111-1984 has been withdrawn; however, copies can be obtained from Global Engineering, 15 Inverness Way East, Englewood, CO 80112-5704, USA, tel. (303) 792-2181.

[4]IEEE Std 119-1974 has been withdrawn; however, copies can be obtained from Global Engineering, 15 Inverness Way East, Englewood, CO 80112-5704, USA, tel. (303) 792-2181.

[5]IEEE Std 272-1970 has been withdrawn; however, copies can be obtained from Global Engineering, 15 Inverness Way East, Englewood, CO 80112-5704, USA, tel. (303) 792-2181.

IEEE Std 390-1987 (Reaff 1993), IEEE Standard for Pulse Transformers (ANSI).

IEEE Std 393-1991, IEEE Standard Test Procedures for Magnetic Cores (ANSI).

IEEE Std 436-1991, IEEE Guide for Making Corona (Partial Discharge) Measurements of Electronics Transformers (ANSI).

IEEE Std C57.12.91-1995, IEEE Standard Test Code for Dry-Type Distribution and Power Transformers (ANSI).

3. Definitions

Electrical and magnetics terms used in this recommended practice are in accordance with those given in IEEE Std 100-1996[6]. Certain parameters and symbols of particular significance in the evaluation of electronics transformers are given where required in the text.

3.1 burden: That property of the circuit connected to the secondary winding that determines the real and reactive power at the secondary terminals. It is expressed either as total impedance with effective resistance and reactance components or as the total voltamperes and power factor at the specified value of current and frequency.

3.2 current-transformation ratio (as opposed to turn ratio): The ratio of the root mean square (rms) value of the primary current to the rms value of the secondary current under specified conditions.

3.3 phase angle (of a current transformer): The phase displacement between the primary and secondary currents. The phase angle is positive when the secondary current leads the primary current.

4. How to specify electronics transformers

Recommended tests and specifications for specific transformer groups are listed in table 1.

5. Insulation and corona tests

5.1 General

An abnormally high alternating voltage is applied between two (or more) isolated elements of the transformer (windings, shields, core, frame, etc.) to test the integrity of major insulation systems in order to demonstrate that the design, materials, and workmanship are adequate.

Unless otherwise specified, the tests should be made in accordance with IEEE Std 4-1995.

[6]Information on references can be found in clause 2.

Table 1—Recommended tests and specifications for specific transformer and inductor groups

Applicable ratings and performance specifications	Small power transformers										Pulse			Broadband							
	Filament	Isolating	Current limiting	Rectifier	Combination	Ferrorresonant	Converter (HF)	Polyphase	Switch mode	Large rectifier	Voltage stepdown	Voltage stepup	Low ratio inverting	Impedance matching	DC isolating	Common-mode rejection	Potential transformers	Current transformers	Filter inductors	Charging inductors	Hybrid transformers
Power source																					
Voltage and range; ac or dc; polyphase	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x		x	x	x
Frequency and range	x	x	x	x	x	x	x	x	x	x				x	x	x	x	x	x	x	x
Pulse width(s) and shape									x		x	x	x								
Repetition rate and max duty cycle									x		x	x	x								
Effective impedance							x			x	x	x	x	x	x	x					x
Primary ratings																					
Voltages and taps	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x				x
Currents and taps	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x
Polarity		x		x			x	x	x	x	x	x	x	x	x	x	x	x			x
DC resistance	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x
Inductance at rated dc							x		x						x				x	x	x
Exciting current	x	x	x	x	x	x	x	x		x	x	x	x								
Effective impedance under load											x	x	x	x	x	x					x
Distributed capacitance									x		x			x	x	x					x
Insulation voltages	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x		x			x
Secondary ratings																					
Voltage (open circuit or loaded, or both)	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x				x
Currents	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x			x
Voltage waveshape under load						x	x				x	x	x								
Frequency response under load														x	x	x					x
Distortion														x	x	x					x
Voltage ratio and errors																	x				
Current ratio and errors																		x			
Leakage impedances and reactances	x	x	x	x	x	x	x	x	x	x											x

Table 1—Recommended tests and specifications for specific transformer and inductor groups *(Continued)*

Applicable ratings and performance specifications	Small power transformers										Pulse			Broadband							
	Filament	Isolating	Current limiting	Rectifier	Combination	Ferroresonant	Converter (HF)	Polyphase	Switch mode	Large rectifier	Voltage stepdown	Voltage stepup	Low ratio inverting	Impedance matching	DC isolating	Common-mode rejection	Potential transformers	Current transformers	Filter inductors	Charging inductors	Hybrid transformers
Secondary ratings (Continued)																					
Regulation for input voltage range						x															
Insulation voltages	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x			x
Polarities		x		x			x	x	x	x	x	x	x	x	x	x	x	x			x
DC resistances	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x			x
Losses																					
Core	x	x	x	x	x	x	x	x	x	x	x	x	x						x	x	
Conductor	x	x	x	x	x	x	x	x	x	x	x	x	x						x	x	
Efficiency	x	x	x	x	x	x	x	x	x	x	x	x	x								
Load										x											
Capacitance							x		x		x	x	x								
Commutation										x											
Insertion														x	x	x					x
Trans-hybrid														x							x
Temperature rise																					
Case	x	x	x	x	x	x	x	x	x	x	x	x	x						x	x	
Conductor	x	x	x	x	x	x	x	x	x	x	x	x	x						x	x	
Hot spot	x	x	x	x	x	x	x	x	x	x									x	x	
Shielding																					
Electrostatic		x		x	x			x	x	x			x	x	x	x	x				x
Electromagnetic														x		x					
Load																					
Resistance	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x		x
Inductance		x		x	x	x	x	x	x	x							x	x			
Capacitance		x		x	x	x	x	x	x	x		x	x	x	x	x	x		x	x	x
Range	x	x	x	x	x	x	x	x	x	x							x	x	x	x	
Nonlinear (transmitter tube)												x									
Rectifier				x	x	x	x	x	x	x											
Recommended acceptance tests																					
Conductor resistances	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x
Winding terminal polarities	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x

Table 1—Recommended tests and specifications for specific transformer and inductor groups *(Continued)*

Applicable ratings and performance specifications	Small power transformers										Pulse			Broadband							
	Filament	Isolating	Current limiting	Rectifier	Combination	Ferroresonant	Converter (HF)	Polyphase	Switch mode	Large rectifier	Voltage stepdown	Voltage stepup	Low ratio inverting	Impedance matching	DC isolating	Common-mode rejection	Potential transformers	Current transformers	Filter inductors	Charging inductors	Hybrid transformers
Insulation tests																					
Resistance	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x
Applied potential (hipot)	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x
Induced potential	x	x	x	x	x	x	x	x	x	x	x	x	x				x	x	x	x	x
Impulse voltage										x											
Corona extinction voltage above 500 V	x	x	x	x	x	x	x	x	x	x							x	x	x	x	
Electrical characteristics																					
Open-circuit secondary (secondaries)																					
Primary inductance at rated dc									x						x	x			x	x	x
Primary exciting current	x	x	x	x	x	x	x	x		x	x	x	x								
Primary exciting impedance															x	x	x				
Primary excit. power	x	x	x	x	x	x	x	x	x	x											
Primary-to-secondary voltage ratios	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x	x				x
Short-circuit secondary (secondaries)																					
Primary voltage for rated primary current				x	x		x	x		x											
Primary voltage for rated second. current	x	x	x			x															
Primary power for rated primary current		x	x	x	x		x	x													
Primary inductance (leakage inductance)	x	x	x						x		x	x	x	x	x	x	x				x
Current ratios																		x			
Commutating inductance										x											
Primary distributed capacitance											x		x	x	x	x					x
Short-circuit primary																					
Secondary distributed capacitance												x	x	x	x	x					x

5.2 Electric strength test (hi-pot test)

NOTE—See 5.2 item f) for repeated electric strength testing.

Electric strength testing shall always be done with all the windings short-circuited. Windings and shields on one side of the insulation system should be connected to frame and ground, while windings and shields on the other side should be connected together (see figures 1 and 2). An essentially sine-wave voltage with a frequency in the range of 45–65 Hz, having adequate current capacity for the application, is applied to the two sets of terminals. The criterion for passing this test is that no electrical breakdown occurs (refer to annex B). All voltages should be defined in the same terms [e.g., root mean square (rms)], peak, or average.

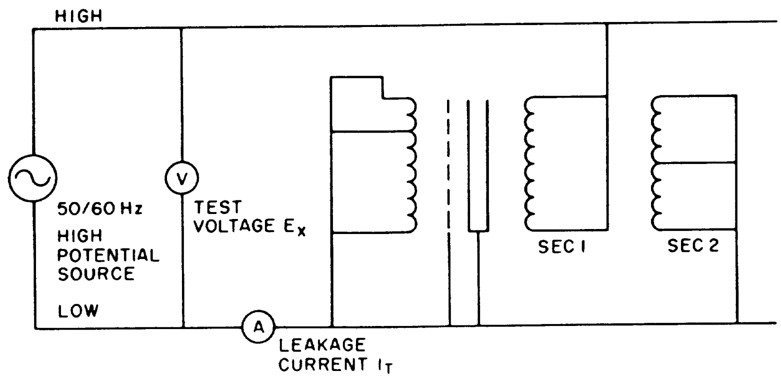


Figure 1—Typical high-potential test, showing section 1 under test

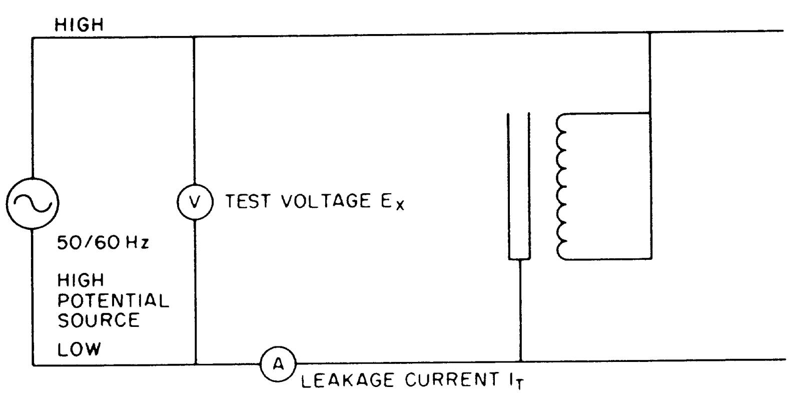


Figure 2—Typical high-potential test of inductor

The voltage should be increased at a convenient rate of not greater than 2000 V/s from zero to the specified value, maintained for 1 min (unless breakdown occurs), and then decreased to zero at the same rate. For production purposes, a voltage 20% higher may be specified for 2 s.

a) Primary windings with rated voltages 600 V or less line-to-line should be tested at an alternating voltage equal to twice the rated voltage of the highest tap plus 1000 V rms, unless otherwise specified.

b) Primary windings with rated voltages over 600 V line-to-line should be tested in accordance with IEEE Std C57.12.91-1995, unless otherwise specified.

c) Secondary windings that have no special test voltage specified should be tested with applied alternating voltage equal to twice the rated voltage of the highest voltage tap plus 1000 V rms, unless otherwise specified.

d) Secondary windings that may have a specific operating direct or alternating voltage derived elsewhere, unless otherwise specified, should be tested at twice the working voltage plus 1000 V rms. High alternating voltage should not be substituted for a direct voltage unless agreed upon between user and manufacturer.

e) Inductors used in line-voltage circuits should be tested in accordance with item a) or b), as applicable. Inductors used in transformer secondary winding circuits should be tested in accordance with item c) or d), as applicable.

f) In case of repeated electric strength testing, since the application of test potentials may impair the strength of the transformer or inductor insulation, any test carried out in accordance with items a) through e) should, if repeated, be made at not more than 80% of the specified test potential for the same time interval.

5.3 Induced potential test

This test applies primarily to insulation systems between layers of windings and between adjacent turns of windings under simulated abnormal functional conditions.

Each secondary winding shall be terminated in an open circuit or into a load resistance that is not less than 2.5 times its normal operating load resistance. All terminals normally grounded shall be grounded during this test and any windings normally biased shall be biased for this test.

5.3.1 General test conditions

Transformers normally driven by a voltage pulse train with pulse duration modulation (or pulse time modulation) should withstand across the primary an induced voltage pulse train for 1 min or 7200 cycles, whichever is less, with each pulse having an amplitude equal to twice the highest normal operating voltage. The pulse should have a duration equal to half the longest normal pulse duration such that the test voltage-time product does not exceed the normal maximum operating voltage-time product.

An alternative test is based on current interruption in an inductance producing a voltage pulse $E = L\frac{\mathrm{d}i}{\mathrm{d}t}$.

5.3.2 General test methods

By means of the induced potential surge test, electrical coils may be conveniently tested for the integrity of turn-to-turn and layer-to-layer insulation by means of high-voltage pulses generated within the coil (figure 3).

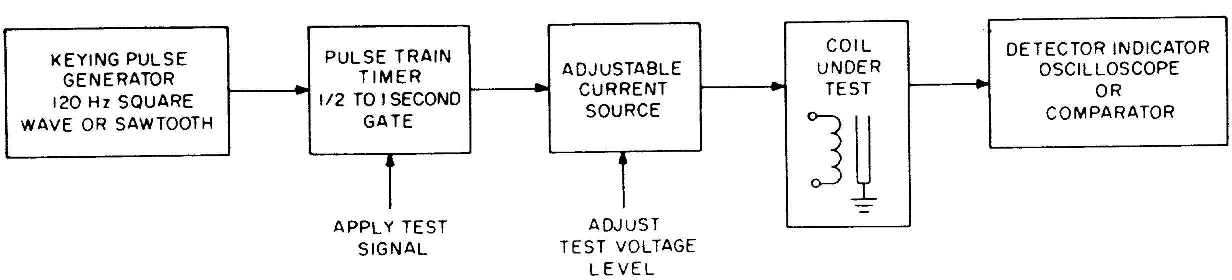


Figure 3—Block diagram of induced voltage surge test

This type of testing is especially applicable to coils wound without interlayer insulation, in a random, controlled random, universal, or layer-wound manner, since the method may cause failure of the coil when one or more of the following faults exist:

— Insufficient crossover lead insulation
— End turns dropped several layers
— Poor-quality wire insulation with breaks or thin spots
— Wire insulation damaged by dereeler or tension device
— Wire insulation damaged by incompatible impregnant or solvent

It should be kept in mind that the test itself is destructive in nature and cannot be applied to a sound coil for an extended period. Therefore, the test duration should be carefully controlled to limit it to the minimum time required to obtain a meaningful indication.

The test consists of applying a controlled train of current pulses to the coil. The leading edge of the pulses may have a moderate slope to allow the current in the coil to rise to a predetermined level. Once this level is reached, the current is disconnected from the coil, and the resultant voltage rise, due to the fast collapse of the current through the inductance of the coil, is observed on an oscilloscope or by means of a comparator to establish the peak value reached $\left(E = L\frac{di}{dt}\right)$.

The coil should be unloaded during this observation so that the voltage is not limited by some unspecified loading. A high-impedance input oscilloscope coupled to the circuit with a frequency-compensated high-impedance voltage divider may be used. Loading should be of the order of 1–10 MΩ.

It is not necessary to have a horizontal sweep operating on the oscilloscope. Prior to the application of the test, the spot on the oscilloscope can be conveniently centered horizontally and can rest on some base level near the bottom of the screen. During the test the excursion of the spot is observed against the calibration of the screen, which acts as a peak reading voltmeter. A 0.5–1.0 s time interval at a 120 Hz rate is sufficient to allow an operator to visually detect any drop or instability in the image caused by internal short circuits that load the coil.

A more sophisticated method uses a comparator consisting of a dc-biased diode bucked against the rising waveform as a detecting device. With the dc bias set to the desired test voltage level, the peak current through the coil is adjusted until the peak exceeds the dc bias by some predetermined amount (10–20 V). The top of the pulse train is clipped and fed to a counter. Pass or fail can be determined by comparing the count with the total number of pulses. If the clipped pulse count is less than the total number of pulses applied by more than, for example, two or three counts, a failure indication is obtained.

In order to ensure the successful generation of high-voltage pulses without the use of excessive current, the test coil should be placed over an iron core during the test. Best results are obtained from cores made up of thin nickel-iron laminations. The core should be grounded for operator protection.

As a precaution against unwarranted damage to the coil, the test current should always be increased from zero until the observation yields the desired voltage level. A test coil may be used for calibration and saved if production of an item is apt to be repeated. The proper current level set on this coil can then be used on the production coils without further adjustment.

Some suggestions as to the determination of the desired voltage level to be used in a specific coil follow.

Assume a coil consisting of 10 000 turns of 36 AWG magnet wire with an insulation rated at 400 V/mil, wound on a form with 250 turns per layer, and therefore having 40 layers. The insulation thickness of the wire is 0.0005 in; therefore two wires laid tightly side by side, having 2 × 0.0005 in separation (0.001 in) should withstand 400 V for an extended time.

The voltage across each layer is obviously 1/40 of the total voltage across the coil. Therefore, the maximum voltage between two wires touching one another, if there is no dropoff at the ends of the coil, occurs between the end turns on adjacent layers and amounts to 1/20 of the voltage across the entire coil.

Suppose our test criterion is that a coil where the end turn drops two layers should be acceptable. The voltage between these turns then can be 1/10 of the total across the coil. If 400 V is a passing level, then 400 times 10, or 4000 V, should be the peak voltage generated in the coil.

It can be seen from the preceding text that long coils with fewer layers are more apt to fail a surge test than short coils of many layers.

In the preceding example, it now becomes obvious that the crossover lead insulation should withstand at least 4000 V peak.

5.3.3 Induced excitation voltage and frequency

Transformers normally driven by a sine-wave or square-wave source should withstand, across the primary, an induced voltage of twice the highest normal operating voltage at a frequency of twice the normal operating frequency.

5.3.4 Repeated induced voltage testing

In case of repeated induced voltage testing, since the application of test potentials may impair the strength of the transformer insulation, any test carried out in accordance with 5.3.1 or 5.3.2 should, if repeated, be made at not more than 90% of the specified test potential for the same time interval.

5.3.5 Excitation current

The input currents should be monitored during the test to check for erratic variations in value. A subsequent normal excitation test should not show a significant change in value from that of a previous test.

5.4 Corona tests

5.4.1 General

Corona is a partial discharge of electrical charges distributed in an insulation system due to the transient ionization of a gas that is part of the insulation system. The gas may be surrounding the conductor as in the case of a terminal, or it may be trapped in a solid or liquid dielectric in the form of a void or bubble in a location where the electric field strength exceeds that necessary to ionize the gas. For detailed treatment of the subject, see [B3][7] and [B7].

Many insulating materials will rapidly fail when in contact with ionized gasses, making the detection and measurement of corona in insulation systems important.

5.4.2 Units of measurement

The unit of test sensitivity of measuring or detecting corona discharges is the picocoulomb, representing the charge transferred due to partial ionization. The quantity affecting the life of the insulation system is the energy that can be transferred by the discharges and is normally expressed in nanojoules. The two units are related:

[7]The numbers in brackets correspond to those of the bibliography in clause 22.

$$W = \frac{1}{2} Q_a V_1 \tag{1}$$

where

W is energy, in nanojoules
Q_a is discharge magnitude, in picocoulombs
V_1 is applied peak voltage, in kilovolts

5.4.3 Detection of corona

The detection of corona at the levels normally encountered in electronics transformers is done by sensing the sudden change of potential distribution in the circuit made up of the insulation system under stress, the coupling impedances, and the test voltage source, caused by the discharge current pulses.

Typical test circuits are shown in figures 4 and 5. For details see IEEE Std 436-1991.

Corona discharges of higher levels may be detected by acoustical means (hissing, cracking sounds, ultrasonic) or by visual means (blue haze).

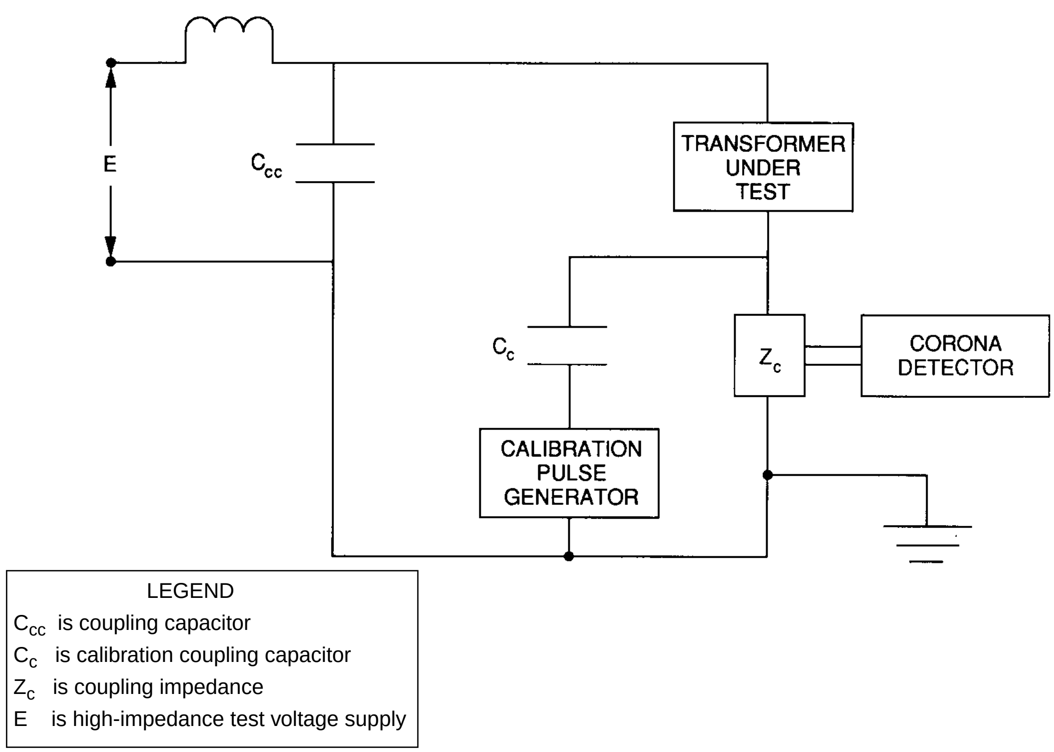
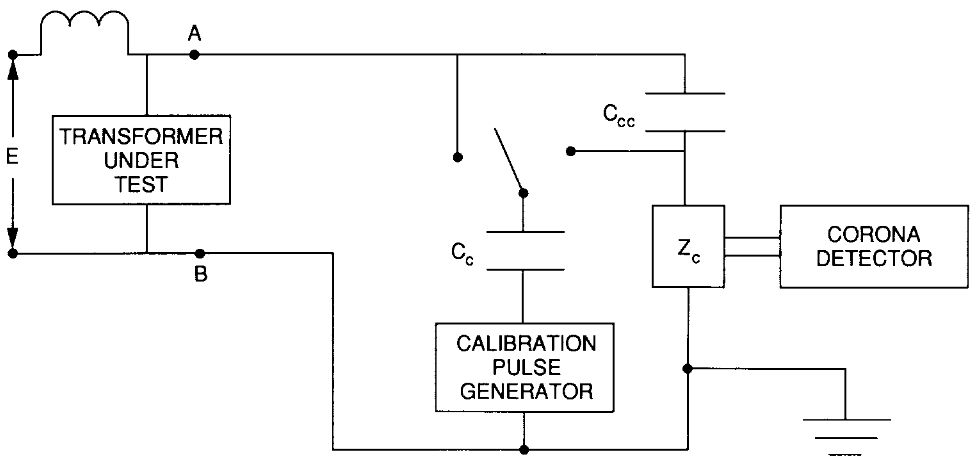


Figure 4—Typical circuit for corona measurement, circuit 1

5.4.4 Analysis of corona

For reliable operation, an insulation system should be free of corona discharges. The insulation system may be characterized by the corona inception voltage, i.e., the voltage where corona discharges of a specified magnitude first appeared; or by the corona extinction voltage, i.e., the lowest voltage where corona discharges were detected as the applied voltage was reduced from a level above the corona inception voltage. The corona extinction voltage is equal to or less than the corona inception voltage.



LEGEND
C_{cc} is coupling capacitor
C_c is calibration coupling capacitor
Z_c is coupling impedance
E is high-impedance test voltage supply

Figure 5—Typical circuit for corona measurement, circuit 2

Where the maximum voltage that can stress the insulation is well defined (including the transients), the insulation quality may be assured by determining that the corona inception voltage is above that level by a specified amount.

Where this is not possible, the insulation system should have a corona extinction voltage that is above the maximum continuous voltage that can stress the insulation.

The margin of safety should be at least 20% in either case.

5.4.5 Test conditions and specifications

For corona measurements, the following quantities must be defined:

— Test voltage (peak)
— Frequency and waveform
— Discharge pulse detection level
— Discharge energy level
— Special ambient conditions (temperature, altitude, humidity, etc.)

The normal requirement is no detectable corona at the specified test voltage and pulse or energy level.

The pulse energy levels recommended by IEEE Std 436-1991 are as follows:

— *Class I.* 70 nJ maximum for insulation systems with a history of satisfactory life at the voltage stress used

— *Class II.* 700 nJ maximum for insulation with considerable tolerance for corona such as inorganic insulation

— *Class III.* Pulse energy level to be specified

6. DC resistance tests

6.1 General

A number of tests for measuring dc resistance of transformer and inductor coils are presented in this clause. The range of resistance values that can be measured using these tests varies from a few microhms to many thousands of ohms. The recommended methods of testing dc resistance are the kelvin double bridge for resistance values less than 1 Ω and the Wheatstone bridge for resistance values greater than 1 Ω. The digital ohmmeter is recommended for both of these resistance ranges. The ammeter-voltmeter and substitution methods are presented for possible use when the previously mentioned equipment is not available.

Methods of measurement that are suitable for measurements above 1 Ω may be unsuitable for low-resistance measurements chiefly because contact resistance causes serious errors.

It is usually essential with low resistance that the two points between which the resistance shall be measured be very precisely defined. The methods that are specially adapted to low-resistance measurement employ potential connections, i.e., connecting leads that form no part of the circuit whose resistance is to be measured but that connect two points, in this circuit, to the measuring circuit. These two points are referred to as the *potential terminals* and serve to fix, definitely, the length of the circuit under test. In the methods used for the precise measurement of low resistance, the "unknown" resistance is compared with a low-resistance standard of the same order as the unknown, and with which it is connected in series. Both resistances are fitted with four terminals, two current terminals to be connected to the supply circuit and two potential terminals to be connected to the measuring circuit. This arrangement is shown in figure 6.

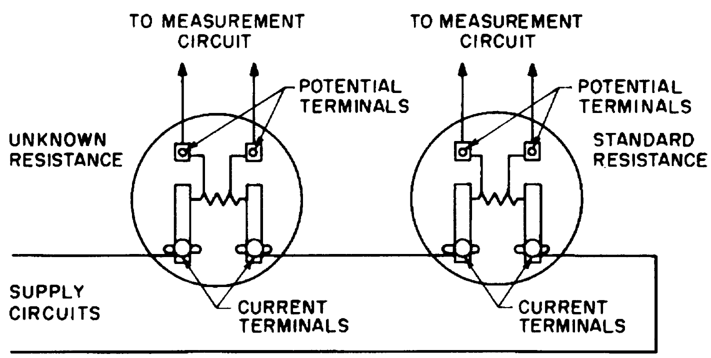


Figure 6—Measurement of low resistance

6.2 Resistance values under 1 Ω—Kelvin double-bridge method

This method is a development of the Wheatstone bridge by which the errors due to contact and load resistances are eliminated. The connections of the bridge are shown in figure 7.

X is the low resistance to be measured and *S* is the standard resistance of the same order of magnitude. These are connected in series with a low-resistance link *r*, connecting their adjacent current terminals. A current is passed through them from a battery supply. A regulating resistance and an ammeter are connected in the circuit for convenience. *Q, M, q,* and *m* are four known noninductive resistances, one pair of which (*M* and *m* or *Q* and *q*) are variable. These are connected to form two sets of ratios as shown, a sensitive galvanometer G connecting the dividing points of *QM* and *qm*. The ratio *Q/M* is kept the same as *q/m*, these ratios being varied until zero deflection of the galvanometer is obtained. Then *X/S* = *Q/M* = *q/m*, from which *X* is obtained in terms of *S, Q,* and *M:*

$$X = \frac{Q}{M} \cdot S \tag{2}$$

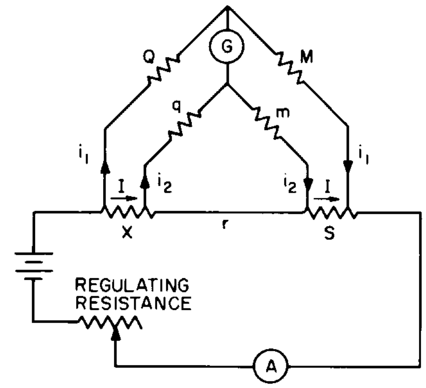


Figure 7—Kelvin double-bridge method of measuring low resistance

In order to take into account thermoelectric electromotive forces (EMF), a measurement should be made with the direction of the current reversed, and the mean of the two readings should be taken as the correct value of *X*.

6.3 Resistance values from 1 Ω to many kilohms

The methods used for this range of measurements are as follows:

— Ammeter and voltmeter
— Substitution method
— Wheatstone bridge
— Ohmmeter
— Digital ohmmeter

6.3.1 Ammeter and voltmeter method

This method, which is the simplest of all, is in common use for the measurement of low resistances when an accuracy of 1% is sufficient. However, this is a relatively rough method, the accuracy being limited by that of the ammeter and voltmeter used, even if corrections are made for the "shunting" effect of the voltmeter. In figure 8, *R* is the resistance to be measured and V is a high-resistance voltmeter of resistance R_V. A current from a steady dc supply is passed through *R* in series with a suitable ammeter.

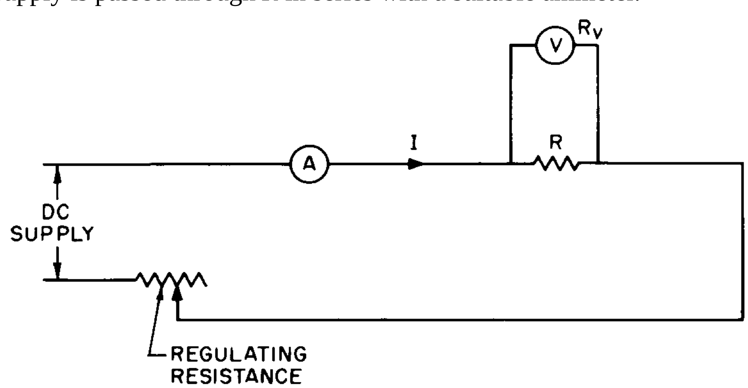


Figure 8—Ammeter and voltmeter method of resistance measurement

Assuming the current through the unknown resistance to be the same as that measured by ammeter A, the former is given by

$$R = \frac{\text{voltmeter reading}}{\text{ammeter reading}}$$

If the voltmeter resistance is not very large compared with the resistance to be measured, the voltmeter current will be an appreciable fraction of current *I*, measured by the ammeter, and a serious error may result.

To correct for the shunting effect of the voltmeter, if the actual value of the unknown resistance is *R*, its measured value R_m, the voltmeter and *R* in parallel is

$$\frac{R \cdot R_v}{R + R_v}$$

and the voltage drop across *R* is

$$I\frac{R \cdot R_v}{R + R_v} = \text{voltmeter reading} \tag{3}$$

Then, assuming voltmeter and ammeter to be reading correctly,

$$\frac{I}{I}\,\frac{R \cdot R_v}{R + R_v} = \frac{R \cdot R_v}{R + R_v} \tag{4}$$

and

$$R = \frac{R_m \cdot R_v}{R_v - R_m} \tag{5}$$

NOTE—The current *I* required to give an acceptable reading for the voltmeter should not cause a heating of the resistor.

6.3.2 Substitution method

The diagram of connections for this method is given in figure 9. *X* is the resistance to be measured while *R* is a variable known resistance. A battery of adequate capacity is used for the supply, since it is important in this method that the supply voltage be a constant. A is an ammeter of suitable range or a galvanometer with a shunt that can be varied as required.

With switch S2 closed and switch S1 on stud a, the deflection of the ammeter or galvanometer is observed. S1 is then closed onto stud b, and the variable resistance is adjusted until the same deflection is obtained on the indicator. The value of *R* that produces the same deflection gives the resistance of the unknown directly.

The resistances *R* and *X* should be large compared with the resistance of the rest of the circuit.

The accuracy of the measurement depends upon the consistency of the supply voltage of the resistance of the circuit excluding *X* and *R* and upon the sensitivity of the indicating instrument as well as the accuracy with which the resistance *R* is known.

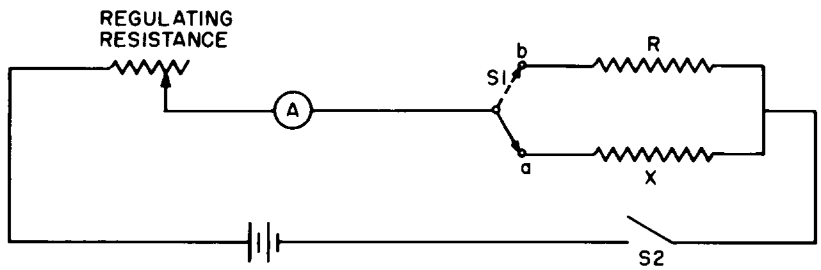


Figure 9—Measurement of resistance by substitution

6.3.3 Wheatstone bridge

The general arrangement is shown in figure 10. *P* and *Q* are known fixed resistors, *S* being a known variable resistance, and *R* the unknown resistance. G is a sensitive galvanometer shunted by a variable resistance *N* to avoid excessive deflection of the galvanometer when the bridge is out of balance. This shunt is increased as the bridge approaches balance so that the shunting is zero, giving full sensitivity of the galvanometer when balance is almost obtained. B is a battery and M is a reversing switch, so that the battery connections to the bridge may be reversed and two separate measurements of the unknown resistance may be made in order to eliminate thermoelectric errors. K_B and K_G are keys fitted with insulating press buttons, so that the hand does not come into contact with metal parts of the circuit, thus introducing thermoelectric EMFs. The battery key K_B should be closed first, followed by the closing of K_G after a short interval. This avoids a sudden (possibly excessive) galvanometer deflection due to self-induced EMFs when the unknown resistance *R* has appreciable self-inductance.

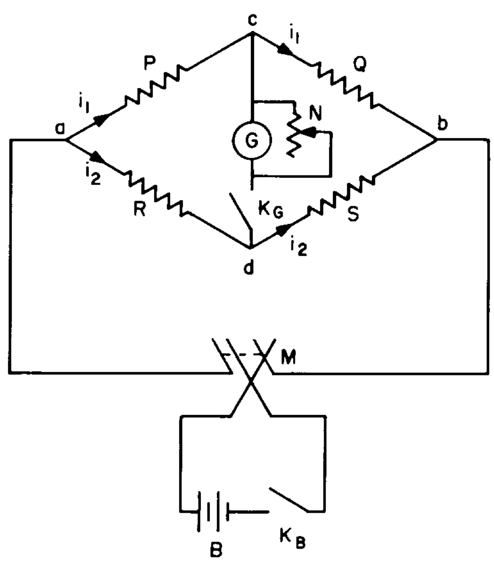


Figure 10—Connections of Wheatstone bridge

At balance (obtained by adjustment of *S*) the same current i_1 flows in both arms *P* and *Q* since the galvanometer takes no current, and the same current i_2 flows in arms *R* and *S*.

The voltage drop across arm *P* equals the voltage drop across arm *R*, and the voltage drop across arm *Q* equals the voltage drop across arm *S*. Then

$$i_1P = i_2R$$
$$i_1Q = i_2S \qquad (6)$$
$$R = \frac{PS}{Q}$$

Arms P and Q are the “ratio arms” of the bridge, and the ratio P/Q may be varied as required to increase the range of the bridge.

Wheatstone bridges are normally constructed with either four or six pairs of ratio coils (tens, hundreds, thousands, and ten thousands in the bridge containing four pairs) and either four or five decades of resistance coils that constitute the variable arm S.

6.3.4 Ohmmeter

The ohmmeter has a moving coil meter, a dry cell, a fixed resistor R_1 and a variable resistance R_2, all mounted within the instrument (figure 11). The unknown resistance R to be measured is connected across the two terminals A and B when the battery circuit is completed. The current flowing through the meter depends on the total resistance of this circuit of which R forms a part, and if R is large the current is small, while if R is small the current is large. The dial of the meter is marked to read directly in ohms. The meter reading would depend to some degree on the state of the battery, but this difficulty is avoided by adjusting the instrument just before each measurement as follows.

Terminals A and B are first short-circuited by a thick piece of copper wire of negligible resistance, and R_2 is adjusted until the meter shows its maximum deflection, indicating zero resistance. The copper wire is then removed and the resistance to be measured is connected across AB when the deflection of the pointer indicates the value of R. The instrument scale is very cramped at the higher resistance end, for the deflections correspond more to values of I/R than of R, and it is most sensitive for low resistances. Usually several ranges of resistance are marked on the dial (e.g., 0–1000 Ω, 0–10 000 Ω, 1–100 000 Ω).

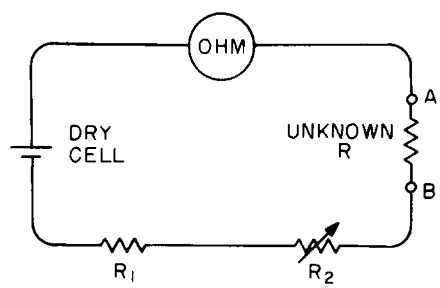


Figure 11—Principle of series ohmmeter

6.4 Digital ohmmeter—Resistance values from under 1 Ω to many kilohms

This method is a development of the ammeter and voltmeter method. The general arrangement of this test method is shown in figure 12. Digital ohmmeters utilize constant current sources, digital electronics, and four terminal connections to provide an accurate indication of resistance values.

Digital ohmmeters provided as dedicated stand-alone instruments incorporating the previously mentioned technologies can provide accurate measurements of resistance from a few microhms to many kilohms. Digital ohmmeters provided as one of the functions of a digital multimeter generally do not incorporate all the previously indicated technologies.

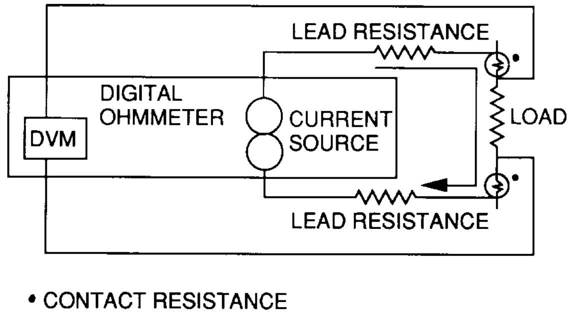


Figure 12—Digital ohmmeter method of resistance measurements

NOTE—Digital ohmmeters that utilize pulsed dc current sources, as opposed to constant dc sources, may introduce errors due to the inductance associated with the coil's winding.

The accuracy of this type of digital ohmmeter is limited at lower values of resistance. Vendor specifications for the accuracy of these devices should be checked against the measurement requirements.

7. Loss measurements

7.1 No-load loss

Transformer no-load loss is defined as the power measured in one transformer winding under specified conditions of excitation and with the voltage-current product equal to zero in all other windings of the transformer. For voltage transformers, the voltage current product is zero when the winding is open-circuited; for current transformers, when the winding is short-circuited. The following discussions refer only to voltage transformers; tests associated with current transformers are presented in clause 21. The term *no-load loss* is often interchanged with the terms *core loss, magnetic loss,* or *excitation loss,* and, under many conditions of operation, there is little difference between the no-load loss and the loss components described by the other three terms. In general there are three components of transformer no-load loss: core loss, ohmic loss in the winding being excited, and (in high-voltage transformers) dielectric loss. In transformers with a magnetic core, the core-loss component is usually much larger in magnitude than the other two components. However, in air-core transformers this component is zero.

The core-loss component of no-load loss is a function of several parameters determined by the source used to excite the transformer. This is a nonlinear functional dependence, which also varies greatly with the type and construction of core material. The principal characteristics of the excitation source affecting transformer core loss are its frequency, waveshape, and voltage (as it affects the magnetic flux density in the core). These parameters, along with the ambient or core temperature, should be specified for all no-load tests.

7.1.1 Excitation waveform

Until recently, most transformer and core testing was specified in terms of the parameters associated with sinusoidal excitation; most core-loss data supplied by the manufacturers of core material are still given only for sinusoidal excitation. Tests using only sinusoidal excitation may not give an adequate measure of the transformer core loss for many applications in which electronics transformers are used, such as in inverter circuits, choppers, switching circuits, and various pulse applications. Therefore the no-load test excitation source should be specified to be as close as possible to that which will excite the transformer in actual operation. Guides for specifying source waveforms are discussed below.

7.1.1.1 Sine-voltage (sine-flux) excitation

The core is excited with a sinusoidal alternating voltage source of very low internal impedance to minimize distortion caused by the nonlinear exciting current. The total harmonic distortion of the induced voltage should be less than 5% in order for this distortion not to affect core loss. This is a very common mode of excitation for core-loss testing and the one by which most manufacturers' loss data are obtained. Magnetic flux density and induced voltage are related by the following well-known relationship:

$$E = 4.44 N f A B_m \quad \text{[volt]} \tag{7}$$

where

N is exciting winding turns
f is excitation frequency, in hertz
A is transformer core effective area, in square meters
B_m is maximum flux density, in tesla
E is rms induced voltage, in volts

Note that the constant in equation (7) results from the values relating average and rms values to maximum values that are peculiar to a sine wave:

$$M_{av} = \frac{2}{\pi} M_m ; \; M_{rms} = \frac{1}{\sqrt{2}} M_m$$

Suggested values for specifying no-load tests for various core materials are given in table 2.

Table 2—Suggested values for specifying no-load tests

Operating flux density (G)	Operating flux density (T)	Material
2500	0.25	Ferrites
4500	0.45	80% Co amorphous
5000	0.50	80% Ni, 20% Fe
10 000	1.00	50% Ni, 50% Fe
14 000	1.40	92% Fe amorphous
15 000	1.50	Si-Fe
15 500	1.55	75% Fe, 20% Co amorphous
20 000	2.00	Cobalt-Iron

Operating frequency (Hz)	Material thickness (in)
60	0.007–0.025
400	0.004–0.006
1000	0.002–0.003
5000	0.0001–0.001
10 000 and above	Ferrites, 80% Co amorphous
—	—
—	—
—	—

7.1.1.2 Sine-current excitation

The core is excited with a sinusoidal current provided by a high-impedance source to minimize the distortion of the current wave by the voltage induced across the core winding. This induced voltage and the associated flux density in the core will generally be nonsinusoidal due to the nonlinear *B-H* characteristic of the core material. There is no simple relationship between flux density and voltage as in equation (7) or in 7.1.1.1. Peak flux density can be determined graphically from the core material *B-H* characteristic and the specified input current, if desired. One method for specifying the current source for this type of excitation is given in 6.4.1 of IEEE Std 393-1991.

7.1.1.3 Square-wave voltage excitation

The core is excited with an alternating square-wave voltage source of very low output impedance to minimize distortion caused by the nonlinear exciting current. The auxiliary-commutated silicon controlled rectifier (SCR) bridge inverter, or McMurray-Bedford inverter [B5], with lead-acid battery energy source, is a simple circuit for achieving such a square-wave excitation source, although many commercial inverters are readily available. With ideal square-wave excitation from a low-impedance source, the flux-density variation in the transformer core is triangular in shape as shown in figure 13, and the maximum flux density in the core is related to the induced voltage by the relationship

$$E = 4NfAB_{\mathrm{m}} \quad \text{[volt]} \tag{8}$$

where E represents the maximum, rms, and half-wave average of the square-wave-induced voltage, and the other symbols have the same meaning as in equation (7).

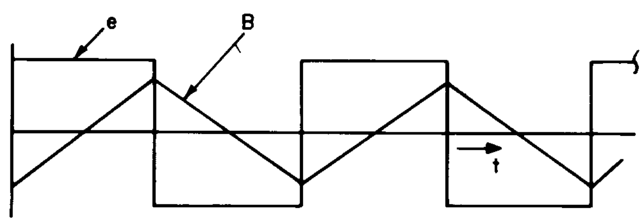


Figure 13—Triangular flux-density variation in transformer core

In practical circuits the ideal square-wave voltage source can seldom be realized. Therefore, for the purposes of no-load transformer loss testing, a source may be considered a square-wave source if it contains the fundamental frequency component and measurable values of all odd harmonics of the fundamental through the 15th harmonic. Suggested values for specifying flux densities and frequencies may be taken from table 2.

7.1.2 Test method and instrumentation

Transformer no-load losses can be measured easily by means of instrumentation available in the average laboratory. The accuracy of the measurement is a function of the accuracy of the instruments used, and instruments with rated accuracy of 1% of full-scale deflection or less are recommended. For measurements in systems with irregular or very nonsinusoidal waveshapes, wattmeters with accuracies of even 1% will be difficult to find. The simplest test circuit is shown in figure 14. The measurement may be performed on any of the transformer windings. Care should be taken to isolate the other windings, which shall be left open-circuited, from test personnel and other equipment. In this circuit an electrodynamometer type wattmeter is illustrated. This the most common type of wattmeter, available in most laboratories. The requirements of the wattmeter and other instruments will now be discussed.

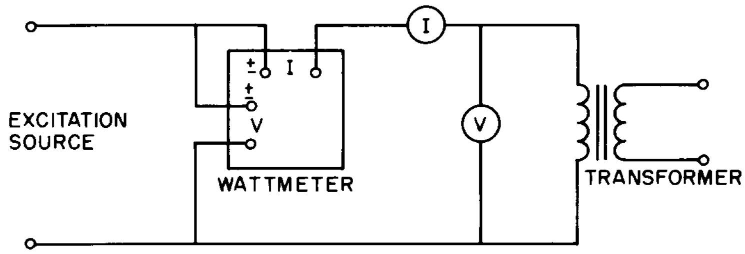


Figure 14—Test circuit for transformer no-load losses

7.1.2.1 Wattmeter

The wattmeter used to measure no-load transformer losses should be of the low-power-factor type, should have a frequency response equal to the highest frequency components in the voltage and current signals being measured, and should have a certified accuracy. The power factors in transformer no-load tests may easily be 0.1 or less, and the wattmeter should be designed and *calibrated* for power measurements in this power-factor range.

The most common type of wattmeter is the electrodynamometer (dynamometer) type, which indicates average power. Low-power-factor dynamometer wattmeters are generally calibrated for power factors up to 0.2, which means that the full-scale value on the wattage scale is 0.2 times the product of current-coil rating and voltage-coil rating. The calibrated accuracy of this type of wattmeter is achievable, and an upscale reading is obtained only when current and voltage coils are connected to the proper voltage points in the circuit. Each coil has one terminal designated by a polarity mark (±), and these terminals should be connected as shown in figure 14. With connections as shown, the power loss in the voltage coil is not included in the wattmeter reading. Standard dynamometer wattmeters have a frequency range up to 800 Hz and accuracies of 0.25% to 1% of full scale, and can be used on all sine-voltage or sine-current excitation tests in this frequency range. Specially calibrated low-power-factor instruments for use at excitation frequencies up to 3200 Hz are also available. Dynamometer wattmeters may also be used with square-wave excitation if the inductive reactance of the voltage-coil circuit is minimized (or compensated for) so as to be 1/1000 of the voltage-coil circuit resistance at the fundamental frequency of the square wave. However, even with this precaution, the wattmeter reading will be less accurate with square-wave excitation than with sinusoidal excitation due to phase-angle errors unless specifically recalibrated with square-wave excitation by means of a calorimeter. In using the circuit of figure 14, the sum of the power losses in the voltmeter and ammeter should be less than the wattmeter probable error times the wattmeter scale reading.

NOTE—The meter *I* should be shorted out and meter *V* should be opened to ensure that they do not affect the wattmeter reading.

There are other types of wattmeters that can be used to measure no-load losses in transformers excited by highly nonsinusoidal signals, such as those associated with inverters, SCR choppers, pulse generators, etc. These include thermal, Hall-effect, and electronic-multiplier wattmeters. These devices generally have a much higher frequency response than the dynamometer type; however, their accuracies seldom achieve 1%, and most have not been calibrated for low-power-factor service and may have large phase-angle errors. These wattmeters and their applications are discussed in more detail in 6.3.2 of IEEE Std 393-1991.

7.1.2.2 Ammeters

A thermocouple (or true rms responding) meter is suggested for all types of excitation since current during the no-load test is nonsinusoidal except with sine-current excitation. Commercial thermocouple ammeters have frequency response capabilities into the megahertz range and accuracies of 0.5–1% of full scale, which are adequate for no-load tests. Thermocouple meters indicate true rms current or voltage regardless of waveshape.

NOTE—Care must be exercised to avoid introducing errors due to the self heating effect.

7.1.2.3 Voltmeters

A thermocouple voltmeter (or true rms responding) meter is suggested for all the same reasons as discussed in 7.1.2.2.

NOTE—Care must be exercised to avoid introducing errors due to the self heating effect.

7.1.3 Test specifications and results

The transformer no-load loss varies with voltage, frequency, and waveform. These parameters and ambient temperature should be specified for each no-load loss measurement and should be stated when presenting or listing the value of the loss. Specifications for various types of transformer excitation are discussed in 7.1.1. The wattmeter reading represents the total no-load loss of the transformer; the core loss (or excitation loss) is found by subtracting ohmic loss in the exciting winding from the no-load loss. The ohmic loss should be calculated using the no-load current measured during the no-load test. In transformers with high-voltage windings, part of the no-load loss may also be composed of dielectric and corona losses. These may be separated from the other no-load test. For more complete analysis of corona loss, see IEEE Std 436-1991.

7.2 Excitation apparent-power measurements

The excitation apparent-power also is found from the no-load test as described in 7.1. The average excitation voltamperes are

$$P_{ex} = \sqrt{(VI)^2 - W^2} \quad \text{[voltampere]} \tag{9}$$

where

V	is no-load test rms voltage, in volts
I	is no-load test rms current, in amperes
W	is no-load test average power, in watts
P_{ex}	is excitation apparent power

The excitation apparent power is often termed *voltamperes reactive* and abbreviated as VAR.

7.3 Stray-load losses

Stray-load loss is a third component of transformer loss (in addition to the no-load and the ohmic or I^2R loss) and results from the flow of load current in the transformer windings. The two principal sources of the stray-load loss are the increased ohmic loss in the transformer windings due to skin effect and eddy currents in the windings, and the induced or eddy-current losses in the transformer case, mounting brackets, etc., due to leakage fluxes. The stray-load loss is a component of the transformer load loss, so termed because it is a function of the transformer load current. The other component of the load loss is the ohmic or I^2R loss in the windings, always much larger than the stray-load component. Both components are obtained from the short-circuit power test of 7.4. The stray-load loss component may be separated out from the total load loss by the method shown in 7.3.2.

7.3.1 Temperature variations

The stray-load loss decreases with increasing temperature, in contrast to the ohmic component of the load loss. This is due to the fact that the stray-load loss is an electromagnetically induced or eddy-current type loss that varies inversely with the magnitude of the resistance in the eddy paths. The stray-load loss may be referred to another temperature by means of the relationship

Copper conductors: $P'_{SL} = P_{SL}\frac{(234.5 + \theta')}{(234.5 + \theta)}$ [watt]

Aluminum conductors: $P'_{SL} = P_{SL}\frac{(226 + \theta')}{(226 + \theta)}$ [watt] (10)

where P_{SL} and P_{SL} are the stray-load losses at θ′ and θ, in °C, respectively.

7.3.2 Calculation of stray-load loss

The stray-load loss can be calculated from the results of the short-circuit power test (see 7.4) and the dc resistance test (see clause 6) by means of the following steps:

a) Specify the winding temperature and load current at which the stray-load loss is desired.

b) Obtain voltage, current, power, and temperature measurements from the short-circuit power test at the specified load current and temperature as outlined in 7.4.

c) Obtain the dc resistance of each transformer winding by means of the methods outlined in clause 6, noting the winding temperature at the time of measurement.

d) Refer the dc resistances measured in step c) to the winding in which the short-circuit test measurements were taken in step b) by means of the transformation ratios listed on the transformer nameplate (or as measured by the technique outlined in clause 8). The sum of these referred resistances is the transformer equivalent dc resistance R_{DCe}.

e) Refer the equivalent dc resistance R_{DCe} obtained in step d) to the specified temperature by the relationship

Copper conductors: $R'_{DCe} = R_{DCe}\frac{(234.5 + \theta')}{(234.5 + \theta)}$ [ohm]

Aluminum conductors: $R'_{DCe} = R_{DCe}\frac{(226 + \theta')}{(226 + \theta)}$ [ohm] (11)

where R_{DCe} is the equivalent dc resistance at the temperature at which the resistance measurements were performed, θ (°C), and R'_{DCe} is the equivalent dc resistance at the temperature specified in step a), θ′ (°C).

f) Calculate the ohmic loss:

$$P_{DC} = (I_{SC})^2 \cdot R'_{DCe} \quad (12)$$

where $I_{SC}^2$is the specified load current used in step b).

g) The stray-load loss is

$$P_{SL} = I^2 \cdot R_{eff} - P_{DC} \quad \text{[watt]} \quad (13)$$

where $I^2 \cdot R_{eff}$ is the total load loss obtained in 7.4.

h) If a negative number results from equation (13), assume that the stray-load loss P_{SL} is zero.

7.4 Short-circuit power test

7.4.1 Purpose

The purpose of the short-circuit power test is to determine the losses due to the effective ac resistance of the transformer, excluding the hysteresis losses of the core, at rated current.

7.4.2 Test circuit

The simplified diagram is shown in figure 15. The circuit consists of a power source of specified frequency, an ammeter to monitor the primary current of the transformer, and a wattmeter to measure the losses in the transformer's primary circuit. The secondary winding is short-circuited with the least impedance practical.

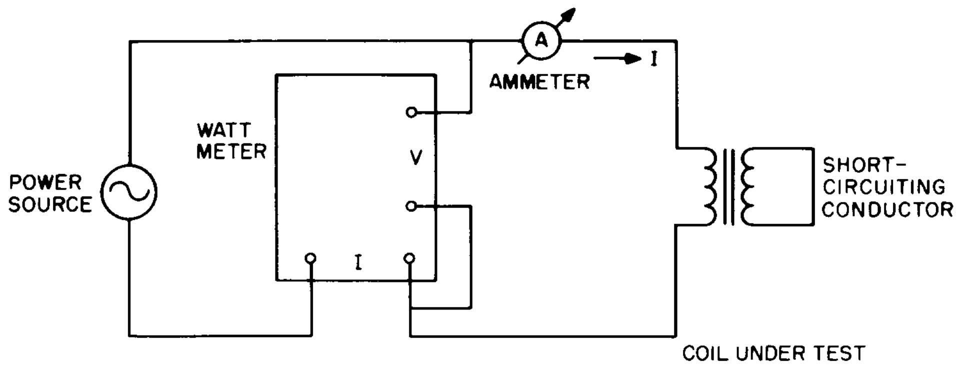


Figure 15—Simplified diagram for short-circuit power test

a) The power source shall furnish the required voltage and current with a specified limit on harmonic distortion.

b) The wattmeter shall indicate the real power lost in the transformer under test with the specified accuracy.

c) The ammeter shall indicate the rms current in the primary winding with a specified accuracy.

d) The short-circuiting conductor and its connections to the terminals of the secondary winding shall have negligible impedances with respect to the secondary winding.

7.4.3 Test procedure

The transformer is connected in the test circuit. The input voltage should be raised from 0 V until the input current, as read on the ammeter, shall equal the rated current. The power indicated by the wattmeter then is

$$P = (I^2 \cdot R_{\mathrm{eff}} + P_{\mathrm{m}}) \quad [\mathrm{watt}] \tag{14}$$

where

I is rated rms current in the primary winding
R_{eff} is effective resistance of the coil including the reflected resistance values of all eddy-current loss elements in and around the winding structure, plus the power losses in the short-circuiting device
P_{m} is power losses in the ammeter and wattmeter
P is indicated power

NOTE—The measurement should be made as quickly as possible after excitation is applied to minimize the effects of self heating.

7.5 Efficiency and power factor

7.5.1 Efficiency

The efficiency of a device is the ratio of the power output from it to the total power input to the device, expressed in percent:

$$\eta = \frac{\text{power output}}{\text{power input}} \cdot 100 \tag{15}$$

The efficiency of transformers can be more accurately calculated when the core and copper losses are known from independent measurements or estimates:

$$\begin{aligned}\eta &= \frac{\text{input power–losses}}{\text{input power}} \cdot 100 \\ &= \frac{\text{output power} \cdot 100}{\text{output power + losses}}\end{aligned} \tag{16}$$

For a two-winding transformer supplying a load with a power factor (PF), the efficiency can be calculated as

$$\eta = \frac{V_2 I_2(\mathrm{PF})}{V_2 I_2(\mathrm{PF}) + \text{core loss} + \text{copper losses}} \cdot 100 \tag{17}$$

$$\eta = \frac{V_2 I_2(\mathrm{PF})}{V_2 I_2(\mathrm{PF}) + \text{core loss} + I_1^2 R_{e1} + I_2^2 R_{e2}} \cdot 100 \tag{18}$$

where

V_2 is rms output voltage
I_2 is rms output current
PF is load power factor, per unit
I_1 is rms input current
R_{e1} is primary ac resistance
R_{e2} is secondary ac resistance

The core loss should be measured or estimated at the flux level required to induce V_2 volt at the output under operating conditions.

For multiple-winding transformers the power output to each load shall be added to get the total, and the losses in all the windings shall be summed to obtain the total copper loss.

7.5.2 Power factor

The power factor (PF) of a reactive circuit component is the ratio of the power input to the device to the total voltampere product flowing into it:

$$\mathrm{PF} = \frac{P}{V \cdot I} \tag{19}$$

where

V is rms input voltage
I is rms input current
P is actual power input (“real” power)
$V \cdot I$ is apparent-power input

Alternately,

$$P = V \cdot I(\mathrm{PF}) \quad [\text{watt}] \tag{20}$$

Since the power factor is also the cosine of the phase angle θ between V and I

$$P = V \cdot I \cos\theta \quad [\text{watt}] \tag{21}$$

If the device can be described by its component parameters, the resistive component R, the reactive component X, and the total impedance Z, its power factor is

$$\mathrm{PF} = \cos\theta = \frac{R}{\sqrt{R^2 + X^2}} = \frac{R}{Z} \tag{22}$$

In an inductive device the current will lag in phase behind the applied voltage, thus the inductor is said to have a lagging power factor. Accordingly, a capacitor has a leading power factor. In transformers the power factor is determined by the reflected load impedances unless the magnetizing inductance is very low, or the leakage inductances are very high with respect to the load (e.g., ferroresonant transformers).

8. Ratio of transformation

8.1 General

The ratio of transformation of two magnetically coupled windings is the transformer parameter determined primarily by the ratio of the number of turns in each winding. By convention it is measured as the ratio of voltages induced across each winding by a common exciting current, with the windings connected series aiding (see 7.2). Thus it can be described as the forward voltage transfer ratio of the transformer with the currents in the two windings being identical.

8.1.1 Ideal transformation ratio

For an ideal transformer with windings of N_1 and N_2 turns, the ratio of transformation equals the turns ratio:

$$a_{12} = \frac{N_1}{N_2} \tag{23}$$

Due to the relationship between the number of turns and the inductance of the windings in an ideal transformer, the ratio of transformation is also equal to the square root of the ratio of the winding self-inductances:

$$a_{12} = \sqrt{\frac{L_1}{L_2}} \tag{24}$$

These relationships are valid only where the coefficient of coupling is unity and the resistive components of the winding impedances have the same ratio as the square of the turns ratio. They may also be used in equivalent circuits.

8.1.2 Transformation ratio by impedance

The ratio of transformation of a practical transformer with a first winding having N_1 turns to a second winding magnetically coupled to it and having N_2 turns may be determined as the ratio of the impedances of the two windings, including the mutual impedances:

$$a_{12} = \frac{Z_{10} + K\sqrt{Z_{10} \cdot Z_{20}}}{Z_{20} + K\sqrt{Z_{10} \cdot Z_{20}}} \tag{25}$$

where

a_{12} is ratio of transformation
Z_{10} is open-circuit impedance of winding 1, $= R_{10} + j\omega L_{10}$
Z_{20} is open-circuit impedance of winding 2, $= R_{20} + j\omega L_{20}$
K is coefficient of coupling, with the term $K\sqrt{Z_{10} \cdot Z_{20}}$ representing the mutual impedance between the two windings

8.1.3 Transformation ratio by inductance

In the special case when the phase angles of Z_{10} and Z_{20} are equal, the ratio of transformation is equal to the ratio of the inductances of the two windings, including the mutual inductances:

$$a_{12} = \frac{L_{10} + K\sqrt{L_{10} \cdot L_{20}}}{L_{20} + K\sqrt{L_{10} \cdot L_{20}}} \tag{26}$$

where

a_{12} is ratio of transformation
L_{10} is open-circuit impedance of winding 1
L_{20} is open-circuit impedance of winding 2
K is coefficient of coupling, with the term $K\sqrt{L_{10} \cdot L_{20}}$ representing the mutual impedance between the two windings

This condition is established in most conventional ratio-of-transformation bridges.

8.1.4 Coefficient of coupling

The coefficient of coupling is the ratio of the impedance of the coupling to the square root of the product of the total impedances of similar elements in the two meshes. In the case of transformers it refers to inductive coupling:

$$K = \frac{M}{\sqrt{L_{10} \cdot L_{20}}} \tag{27}$$

where

M is mutual inductance

The coupling coefficient may be determined from the relation

$$K = \frac{L_{sa} - L_{so}}{4\sqrt{L_{10} \cdot L_{20}}} \tag{28}$$

where

L_{sa} is inductance of the two windings connected series aiding
L_{so} is inductance of the two windings connected series opposing
L_{10} is open-circuit impedance of winding 1
L_{20} is open-circuit impedance of winding 2

8.2 Measurement methods

The measurement of the ratio of transformation is performed under the conditions of 8.1.2; that is, the phase angles of Z_1 and Z_2 are equalized. This may be accomplished by adding compensating resistance to one of the windings until the difference between phase angles is eliminated, or by using a method where only the impedance components that are in phase are compared. The two windings can be connected series aiding to permit the current to pass through both, thus simplifying the equalization of phase. Then the voltages appearing across the two windings can be accurately compared using potentiometric methods. The current through the windings should be limited to where the core and copper losses are negligible.

8.2.1 Resistive ratio bridge (figure 16)

$$a_{12} = \frac{R_1}{R_2} \cong \frac{N_1}{N_2} \tag{29}$$

where

R_2 is fixed bridge arm, with a value that is sufficiently high not to load the signal generator significantly
R_1 is precision variable resistor
R_c is compensating resistor
D is detector

NOTE—R_1, R_2, and R_c shall be noninductive.

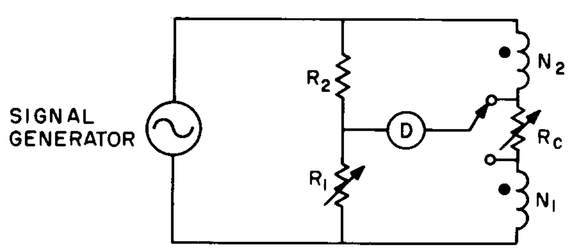


Figure 16—Resistive ratio bridge

8.2.2 Bridge with ratio transformer (figure 17)

$$a \cong \frac{N_1}{N_1 + N_2} \tag{30}$$

In this circuit the potentiometer of the bridge described in 8.2.1 is replaced by a precision inductive voltage divider (ratio transformer).

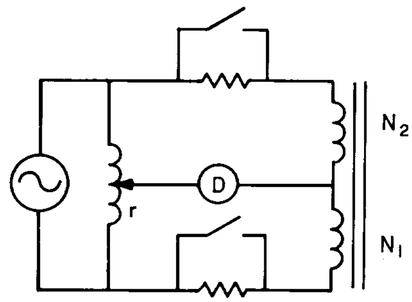


Figure 17—Bridge with ratio transformer

8.2.3 Ratio of transformation from voltage measurements

The recommended method of connection is shown in figure 18. When the conditions of 8.1.1 apply, the ratio of transformation may be determined from the ratio of voltages across the windings:

$$a_{12} \cong \frac{E_1}{E_2} \tag{31}$$

This method may not indicate reverse polarity unless special precaution is taken.

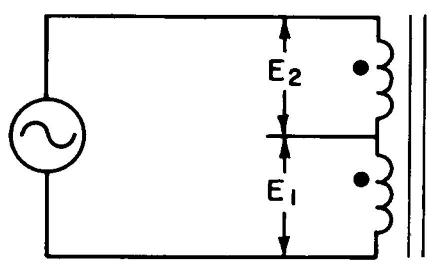


Figure 18—Ratio of transformation from voltage measurements

8.3 Impedance unbalance

The impedance unbalance U between windings 1 and 2 is defined as

$$U = \frac{Z_{20} - Z_{10}}{Z_{10} + K\sqrt{Z_{10} \cdot Z_{20}}} \cdot 100 \tag{32}$$

where

U is impedance unbalance, in percent
Z_{10} is open-circuit impedance of winding 1

Z_{20} is open-circuit impedance of winding 2
K is coefficient of coupling

For the special case where the phase angles of Z_{10} and Z_{20} are equal, the impedances may be replaced with the corresponding inductances, giving the inductance unbalance

$$U = \frac{L_{20} - L_{10}}{L_{10} + K\sqrt{L_{10} \cdot L_{20}}} \cdot 100 \tag{33}$$

If N_1 and N_2 are the number of turns of the windings indicated by the corresponding subscripts, and if the winding impedances are proportional to the squares of the turns in the respective windings, then

$$U = \left[\frac{N_2(N_2 + KN_1)}{N_1(N_1 + KN_2)} - 1\right] \cdot 100 \tag{34}$$

For the special case when the coefficient of coupling $K = 1$, the inductance unbalance is

$$U = \frac{N_2 - N_1}{N_1} \cdot 100 \tag{35}$$

The inductance unbalance between windings 1 and 2 may be expressed in terms of the ratio of transformation a_{12} between the windings:

$$U = (a_{12} - 1) \cdot 100 \tag{36}$$

The measurement of inductance unbalance may be accomplished with a bridge circuit as shown in figure 19, where the resistors R_1 are ratio-arm resistors and R_2 is a calibrated ratio-arm potentiometer. The maximum unbalance for which this bridge can be balanced is

$$U_{\max} = \frac{R_2}{R_1} \cdot 100 \tag{37}$$

R_2 may be calibrated in ±% unbalance.

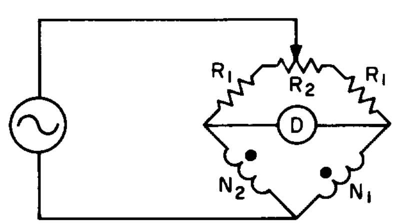


Figure 19—Bridge circuit for measurement of impedance unbalance

The null-detector constraints for the ratio of transformation and inductance unbalance testing are as follows:

a) *Frequency range.* As required for intended use.

b) *Sensitivity.* Should be able to indicate unbalances smaller than one-half the accuracy limit of the intended measurement.

c) *Harmonic suppression.* Sufficient to prevent errors or lack of precision due to the harmonics present in the output signal of the bridge.

d) *Input impedances.* Input to ground impedances should be sufficiently large and so arranged that errors due to stray currents will not be introduced. An impedance matching transformer with appropriate shielding may be used to match the bridge output to the detector input. The differential impedance between the bridge output terminals is not critical since under null conditions both the voltage and the current approach zero.

e) *Phase-sensitive detectors.* May be used to reduce the need for compensating resistance for windings of differing phase angles. They will also reduce the time required to perform tests against minimum or maximum limits.

8.4 Balance tests

8.4.1 General

The test circuits shown in figure 20 should be used where specified for determining the balance of the transformer windings. Matched resistors of the precision type should be used for the resistor pairs R_1, R_2, and R_3, R_4. The sums R_1 plus R_2 and R_3 plus R_4 should equal, within 5%, the terminating impedance specified for the associated windings.

The test may be made at any frequency, but frequencies near (1) the low-frequency 1 dB point, (2) the mid-band frequency, and (3) the high-frequency 1 dB point should be used unless other frequencies are specified. In general, at the low frequencies the elements contributing to unbalance will usually be winding resistance and turns; in the middle frequencies, winding turns; and at the high frequencies, leakage inductances and winding capacitances. The test is not limited to two-winding transformers, and the test circuit may be modified as required. For additional information on balance measurements see [B8].

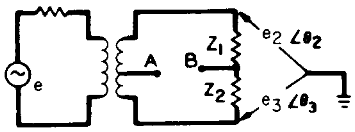
8.4.2 Method of measurement

Select the appropriate test circuit from figure 20, depending on the anticipated use. The various conditions of grounding and the appropriate methods of testing are described in the table within figure 20. The accuracy of measurement will depend on the ability to measure the voltage ratios with sufficient accuracy. Since for good balance conditions the voltage ratio e_4/e_3 will be quite large, it is advantageous to use as high a voltage as possible for the signal e_3. This voltage is not limited by core loss since the energized windings are connected flux opposition. Thus this voltage is limited only by the dielectric strength of the insulation between windings and shield. The voltage e_1 is applied directly to the winding (or windings); therefore it is necessary to limit its level to the normal maximum operating level of the transformer. The impedance level of the signal source is of no importance since only the ratios of the voltages, namely, e_2/e_1 and e_4/e_3, are required. The impedance of the voltmeter used for the measurement of e_2 and e_4 must be high compared to the termination R_3 [or R_1 in figure 20 (g) and (h)] to avoid improper termination of the transformer under test. Since these measurements will be sensitive to the values of termination, R_1 through R_4, it is essential that the resistance terminations be those appropriate to the circuit in which the transformer is used, even though they might not be the values that would produce the least in-band loss.

8.5 Polarity tests

The polarity of a transformer winding is determined by the polarity of the voltage induced in the winding with respect to a voltage applied to a reference or primary winding. For an alternating reference voltage the winding is either in-phase (with near 0° phase shift) or out-of-phase (with near 180° phase shift) with the primary or reference winding.

A. BALANCE DEFINITION

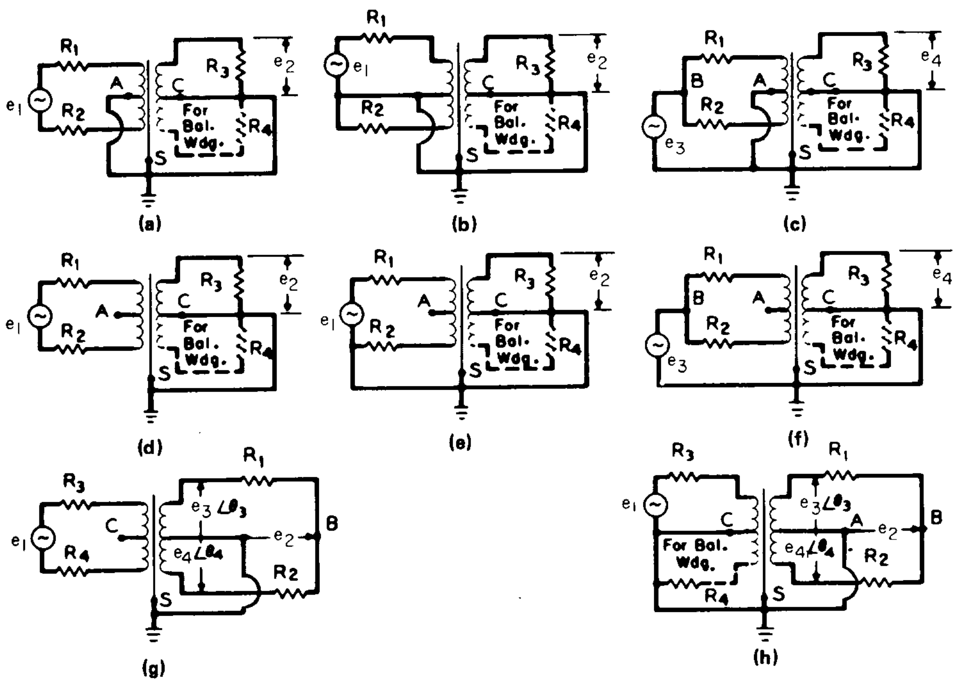


$$\text{TRANSFORMER BALANCE} = 20\log_{10}\left|\frac{e_2\angle\theta_2 - e_3\angle\theta_3}{e_2\angle\theta_2 + e_3\angle\theta_3}\right| \text{ dB}$$

RESTRICTIONS:
A,B OR BOTH MUST BE GROUNDED
$Z_1 = R_1 + jX_1 = Z_2 = R_2 + jX_2$

B. BALANCE TEST METHODS

GROUND CONDITIONS		TEST METHOD	TEST CIRCUIT	BALANCE EQUATION	NOTES
TERMINAL A	TERMINAL B				
GROUNDED	GROUNDED	I	(a) or (b), (c)	$20\log_{10} e_2/e_1 - 20\log_{10} e_4/e_3 + 6.02$ dB	1,2,3,4,6,7
FLOATING	GROUNDED	II	(d) or (e), (f)	$20\log_{10} e_2/e_1 - 20\log_{10} e_4/e_3 + 6.02$dB	1,3,4,6,7
GROUNDED	FLOATING	III	(g) or (h)	$20\log_{10} \left\|(e_3\angle\theta_3 - e_4\angle\theta_4)/e_2\right\|$ dB	1,3,5,6,7



NOTES
1—Winding with terminations R_1 and R_2 is the winding under test.
2—Test method I is the same as that for longitudinal balance; transformer balance is 6.02 dB greater than longitudinal balance.
3—Where both windings are balanced, center tap C should be grounded or floating as required to simulate use condition.
4—Where $e_1 = e_3$, balance = $20 \log_{10} 2e_2/e_4$.
5—Ideally, $e_3 = -e_4$; only a small approximation gives balance $\cong 20 \log_{10} e_3/e_4$.
6—(a), (d), and (g) require a balanced source. If the transformer balance is 20 dB or better, an unbalanced generator may be used as in (b), (e), and (h) with less than 1 dB error.
7—In all cases, $R_1 = R_2$ and $R_3 = R_4$.

Figure 20—Test circuits for balance tests

Polarity tests for transformer windings may be combined with the ratio-of-transformation tests. The ratio bridges in accordance with 8.2.1 and 8.2.2 will not balance properly due to the negative coefficient of coupling when the windings are connected series opposing.

The relative winding polarities may be observed by using an oscilloscope to compare the phases of input and output waveforms. The polarity of single pulses may also be detected by galvanometers or similar instruments.

Since the induced voltages can be added as vectors, the winding polarities can be determined from voltage measurements. For windings properly assembled with the shown polarity (figure 21),

$$|E_3| = |E_1| + |E_2| \tag{38}$$

For windings with opposite polarity,

$$|E_3| = |E_1| - |E_2| \tag{39}$$

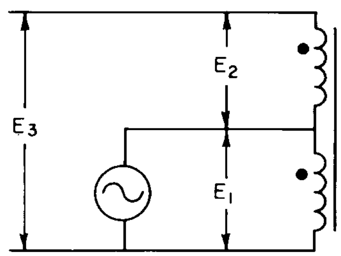


Figure 21—Polarity test using voltage measurements

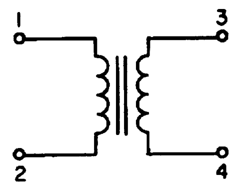
9. Transformer capacitance

9.1 General

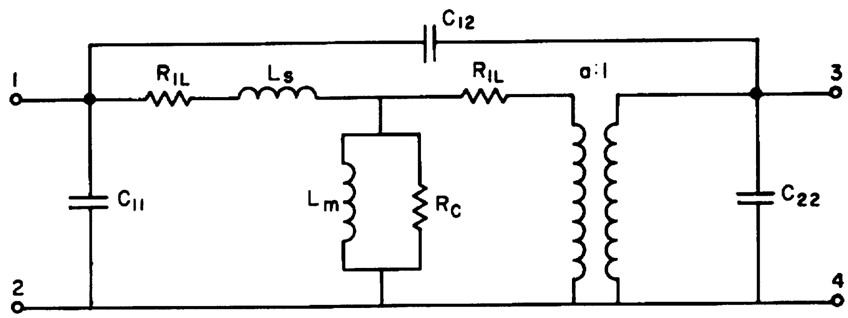
Transformer capacitance is the result of energizing the transformer. When the transformer is energized, different voltage gradients arise almost everywhere. In the pervasive electric field there are a large variety of capacitances that correspond to each of these gradients:

— Between turns
— Between layers
— Between windings
— Between terminals
— Between the core and the end turn of a layer
— Between the core and each terminal

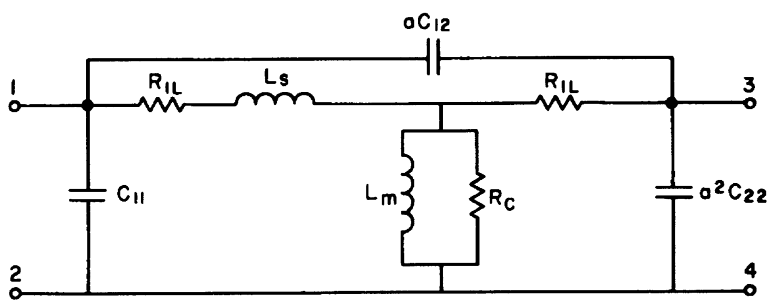
Only the lumped rather than the distributed parameter will be considered and only three capacitances (C_{12}, C_{22}, and C_{11}) will be used (see [B6]). The measuring techniques for the three capacitances are as follows. Given a two-winding transformer [figure 22 (a)] whose equivalent circuits are as shown in figure 22 (b), and transferring C_{12} and C_{22} to the left-hand side and eliminating the ideal transformer, we have the circuit shown in figure 22 (c). Since the resistance has no effect in measuring the inductance and capacitance, the circuit can be simplified further as shown in figure 23.



(a) Two-winding transformers



(b) Equivalent circuit of (a)



(c) With ideal transformer eliminated

Figure 22—Measuring techniques for transformer capacitance

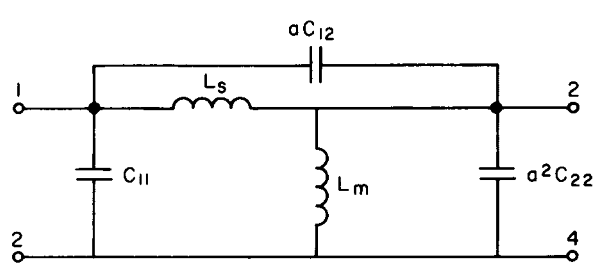


Figure 23—Simplified circuit for transformer capacitance measurement

9.2 Interwinding capacitance

The interwinding capacitance may be determined by connecting the shorted primary terminals and the shorted secondary terminals, respectively, to the two unknown terminals of a capacitance bridge or digital

LCR meter. The core, shield (if any), and additional secondaries should be appropriately guarded or connected to ground in order to eliminate them from the measurement, as specified by the equipment manufacturer. Figure 24 shows a typical measurement setup. The measured values of interwinding capacitance can vary with frequency or due to stray inductances. It is therefore recommended that, for design verification purposes, more than one test frequency is used. For production testing, test frequencies below 10 kHz generally are chosen.

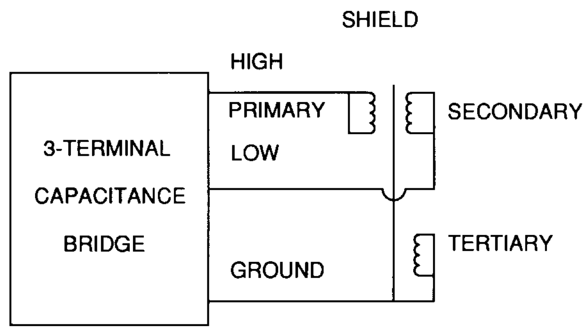


Figure 24—Test circuit for interwinding capacitance measurement

9.3 Distributed capacitance

The distributed capacitance (or self-capacitance) may be determined by either of the following methods:

a) Finding the resonant frequency of the winding under test

b) Measuring the apparent capacitance of the winding under test at a frequency significantly (i.e., five times) above the resonant frequency

The distributed capacitance so measured is actually a lumped capacitance made up of multiple capacitances all reflected to the particular winding under test. The separation of this one lumped capacitance into its constituents is very difficult and is not within the scope of this recommended practice.

Figure 25 shows a measurement circuit with a transformer having a balanced primary, an unbalanced secondary, and an electrostatic shield.

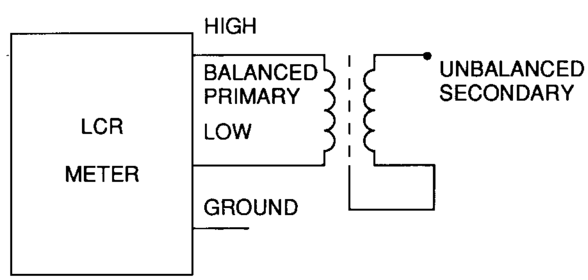


Figure 25—Test circuit for distributed capacitance measurement

9.3.1 Resonant frequency method

The two terminals of the winding under test are connected to a digital LCR meter or equivalent. The test frequency is varied until the winding impedance appears to be purely resistive. The distributed capacitance can be calculated from the equation

$$C_{\mathrm{dist}} = \frac{1}{4\pi^2 f_r^2 L_m} \tag{40}$$

where L_m is the open-circuit inductance of the winding under test and f_r is the resonant frequency.

NOTE—L_m usually varies with frequency and core flux density, so care should be taken to account for these effects.

9.3.2 Above-resonance method

The distributed capacitance can be measured on a digital LCR meter at a frequency where the phase angle approaches –90°. In actual practice, phase angles less than –89° may be the best that can be achieved, in which case the accuracy is suspect.

Another way to judge the best test frequency is to set it near the point of inflection (minimum capacitance) in the curve of effective capacitance versus frequency.

9.4 Bridge methods

Bridge methods can be used for transformer capacitances. Typical capacitance bridge circuits are shown in figure 26.

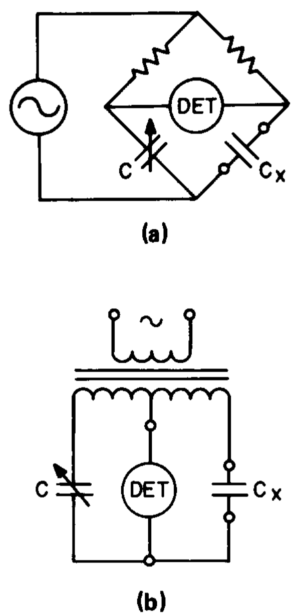


Figure 26—Circuit for bridge method

10. Inductance measurements by impedance bridge method

10.1 General

Unless careful attention is paid to the measurement of impedance, serious inaccuracy may result. It is the purpose of this clause to point out possible ways in which inductance and effective resistance or capacitance and conductance may be represented and measured.

In a wideband transformer, the core is usually of high-permeability material wound as a continuous strip or with a very small air gap. Under these conditions the Q, or the ratio of effective series reactance to effective series resistance, is small. Both inductances L and Q are dependent on the voltage level and frequency used. Winding resistance ordinarily has a much smaller value than the core-loss equivalent series resistance in this kind of transformer. Short-circuit inductance also is very small compared to mutual inductance. Therefore, it is more representative of actual performance to regard open-circuit inductance L_{sh} and core-loss resistance R_{sh} as parallel elements.

As frequency increases above the minimum, open-circuit reactance increases proportionately, so that at mid-band and higher frequencies open-circuit reactance becomes so large as to be virtually negligible. Core-loss resistance does not change nearly as fast with frequency. Both inductance L_{sh} and equivalent parallel resistance R_{sh} are most significant at the lowest frequency, but at midband the open-circuit reactance may be ordinarily neglected, whereas the parallel equivalent resistance may contribute appreciably to midband transformer loss. For the parallel representation, Q equals $R_{sh}/2\pi f L_{sh}$. At any frequency it is possible to convert from series to parallel equivalence by using the two expressions for Q. This conversion is unnecessary if the proper bridge configuration is used.

Bridge measurements are applicable not only to measurements of high Q but also to open-circuit measurements on low-score materials, capacitance between windings, low Q; to measurements on lossy materials, and measurements of terminated impedance to determine the reflection coefficient. The accuracy of measurement for each type can be optimized by the method in which the initial measurement or zero balance is made.

10.2 Method of measurement

As noted, the method in which the initial balance is made is of significance. Four appropriate methods are listed in 10.2.1 through 10.2.4, depending on the type of measurement to be made. See figure 27 for typical bridge circuits.

10.2.1 Series bridge, low impedance

In order to obtain highest accuracy, it is necessary to use rather heavy straps from the bridge terminals to the terminals of the apparatus. A shorting strap comparable in length to the distance between terminals of the apparatus should be used and the straps to the bridge should be disturbed as little as possible.

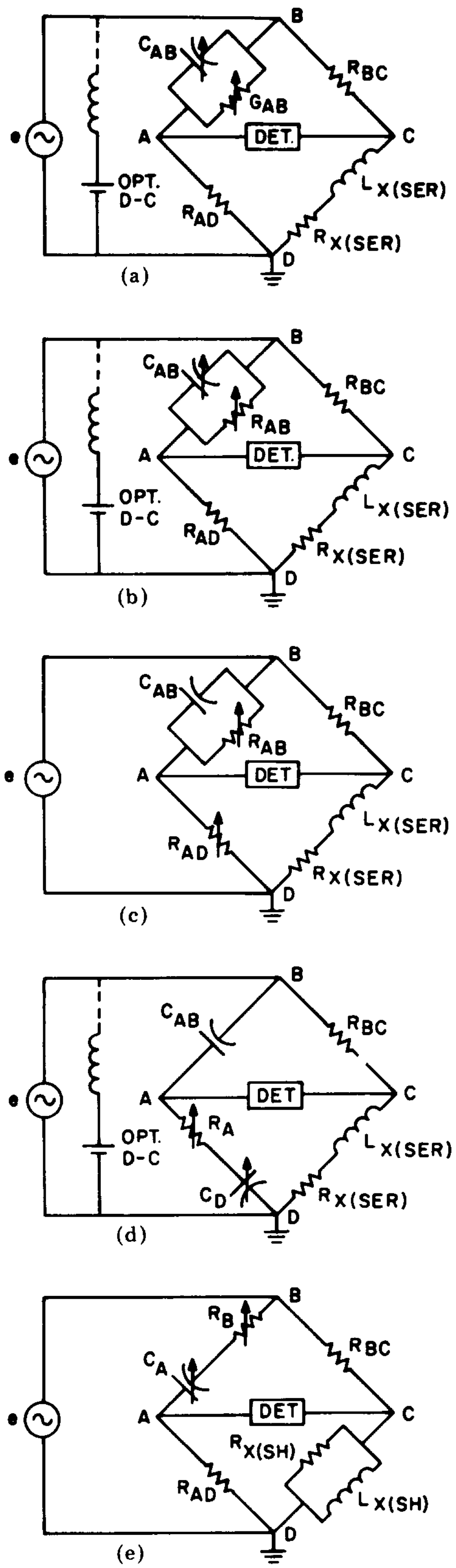
The impedance being measured may be determined from the formulas

$$\begin{aligned} L_x &= L_1 - L_0 + L_c \\ R_x &= R_1 - R_0 + R_c \end{aligned} \tag{41}$$

where

L_x, R_x	are inductance and effective resistance of components being measured
L_1, R_1	are bridge readings (corrected in accordance with the bridge instructions where necessary) for the component in place
L_0, R_0	are zero-balance measurements when a shorting strap is used in place of the component

L_c and R_c are corrections for the inductance and effective resistance of the strap. The value of the strap inductance L_c is given by



Maxwell *L—R* Bridge Using Conductance Standard

$L_X = (R_{AD}R_{BC})C_{AB}$
$R_X = (R_{AD}R_{BC})G_{AB}$

Reads series components directly requires a conductance standard.

NOTE: Will balance on dc like a wheatstone bridge and will read accurately over a wide frequency range. Best accuracy at low and medium-high Q values. For superimposed dc and high power measurements, R_{AD} is made large and R_{BC} small.

Maxwell *L—R* Bridge Using Resistance Standard

$L_X = (R_{AD}R_{BC})C_{AB}$
$R_X = (R_{AD}R_{BC})/R_{AB}$

Does not require a conductance standard but R_X must be computed.

(Same NOTE as under (a).)

Maxwell *L—Q* Bridge

$Q_X = \omega L_X/R_X$
$L_X = (R_{BC}C_{AB})R_{AD}$
$Q_X = (\omega C_{AB})R_{AB}$

Can be made to read Q directly at a single frequency.

NOTE: Balances for L_X and Q_X are dependent. Balance is slow and difficult when Q_X is less than 3.

Owen *L—R* Bridge

$L_X = (C_{AB}R_{BC})R_A$
$R_X = (C_{AB}R_{BC})/C_D$

Will not balance on dc but will read accurately over a limited low and audio frequency range.

NOTE: Good for superimposed dc measurements because C_{AB} blocks dc from standard.

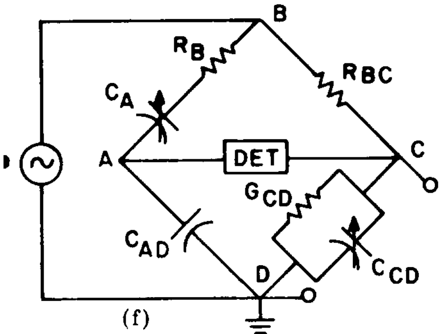
Hay *L—R* Bridge

$L_{X(SH)} = (R_{AD}R_{BC})C_A$
$R_{X(SH)} = (R_{AD}R_{BC})/R_B$

Particularly accurate for high Q measurements. (R_B approaches zero.)

NOTE: If series inductance is desired it must be computed. The *L—Q* version of this bridge is difficult to balance for low Q values.

Figure 27—Typical bridge circuits for inductance measurements

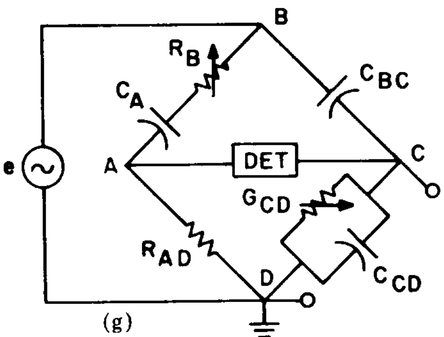


(f)

Schering Bridge Using Two-Capacitance Standards

$$C_X = \Delta C_{CD}$$
$$G_X = \frac{C_{AD}}{R_{BC}}\left(\frac{1}{C_1} - \frac{1}{C_2}\right)$$
C_1 = value of C_A for initial balance
C_2 = value of C_A for final balance

Good for fairly high frequencies.
Capacitance standards may be accurately calibrated.
Not direct reading in G_X.
Limited range of C_X.
L_X determined by resonance method.

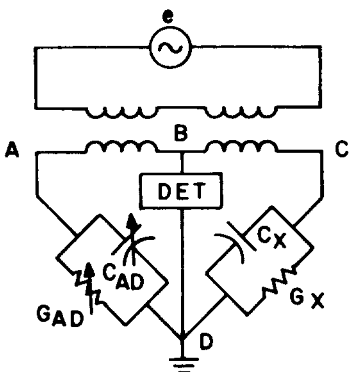


(g)

Schering Bridge Using Two-Resistance Standards

$$C_X = \frac{C_{BC}}{R_{AD}}(\Delta R_B)$$
$$G_X = \Delta G_{CD}$$

Limited to moderate frequencies.
Not direct reading in C_X.
L_X determined by resonance method.



(h)

High-Frequency Admittance Bridge

$$C_X = \Delta C_{AD}$$
$$G_X = \Delta G_{AD}$$

Uses grounded standards.
Direct reading.
Useful over wide frequency range.
L_X determined by resonance method.

Figure 27—Typical bridge circuits for inductance measurements (continued)

$$L_c = 0.005l\left[\log_e\left(\frac{4l}{d} + \frac{d}{2l} - 0.75\right)\right][\text{microhenry}] \tag{42}$$

where

l is length of shorting wire, in inches
d is diameter of shorting wire, in inches

Assuming the shorting wire is copper,

$$R_c = R_{dc} = [l + 28.5 f^2 d^4 \cdot 10^6] \tag{43}$$

where

R_{dc} is dc resistance of shorting wire
f is frequency of test, in megahertz

NOTE—This method may also be used for high impedance, in which case the correction terms may be negligible.

10.2.2 Series bridge, low *Q*

This method is applicable for terminated measurements to determine the reflection coefficient when the impedance to be measured is essentially resistive. In place of the shorting strap used in the method described in 10.2.1, a noninductive, low-capacitance resistor of value as close as possible to the ideal nominal value looking into the terminals of the apparatus is selected. If the resistor has low parasitic inductance and capacitance, the impedance being measured may be determined from the formulas

$$\begin{aligned} R_x &= R_2 - R_a + R_R \\ L_x &= L_2 - L_a + L_R \end{aligned} \tag{44}$$

where

R_x, L_x are effective resistance and inductance of components being measured
R_2, L_2 are bridge readings (corrected in accordance with the bridge instructions where necessary) for the component in place
R_a, L_a are initial measurements with the standard resistor in place of the apparatus
R_R, L_R are dc resistance and inductance of resistor

10.2.3 Parallel bridge, high *Q*

To avoid errors when connecting the bridge to the apparatus under test, fairly rigid leads should be used. The grounded terminal of the bridge should be connected to the ground terminal of the apparatus to be measured, and the high terminal lead should be near but not touching the terminal or terminals of the apparatus. The value of impedance is then merely the difference between the zero (open) measurement and the value measured with the terminals of the apparatus connected to the bridge (corrected in accordance with the bridge instructions where necessary). If a balanced to ground or direct capacitance measurement is to be made, only the grounded terminal of the apparatus should be connected to the grounded terminal of the bridge for the zero or open-circuit measurement with both high-potential terminals of the apparatus.

10.2.4 Parallel bridge, low *Q*

A parallel bridge may also be used for terminated impedance measurements as in the method described in 10.2.2. As before, a low-capacitance, low-inductance resistor should be used for the zero measurement. For this case the admittance being measured may be determined from the formulas

$$\begin{aligned} R_p &= R_3 - R_B + R_R \\ C_p &= C_3 - C_B + C_R \end{aligned} \tag{45}$$

where

R_3, C_3 are measured parallel resistance and capacitance of component
R_B, C_B are initial measurements with resistor in place of component
R_R, C_R are dc resistance and capacitance of standard resistor

11. Transformer response measurements

11.1 Transformer frequency response

When transformers are designed to operate over a fairly wide range of frequencies and serve the function(s) of matching a load to a power source with a linear response over that range, to provide voltage isolation between source and load, or to provide rejection of common-mode voltages between source and load, such transformers are usually tested to determine their frequency response characteristic over a specified range when operated in an amplifier circuit for which they were designed. Also, where transformers are used in control systems, the transformer frequency response is often valuable in evaluating the stability and other dynamic properties of the control system.

The frequency response of a simple two-winding transformer can be measured in a relatively simple circuit consisting of a variable frequency low-impedance sine-wave source of low harmonic content, a linear and noninductive load resistor of a value equal to the rated secondary load resistance, a linear capacitor equal in capacitance to the rated load shunt capacitance, and a linear noninductive resistor in series between the variable-frequency source and the transformer primary, which is equal in value to the internal resistance of the power source from which the transformer under test is designed to operate. Figure 28 shows such a test circuit.

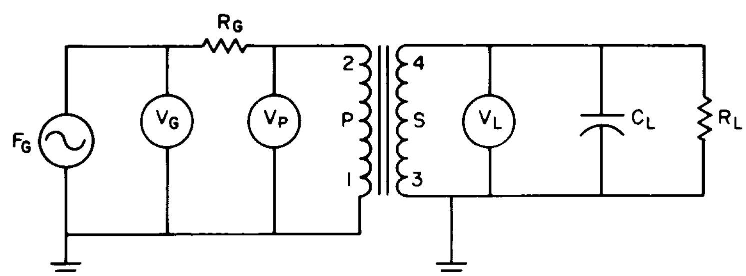


Figure 28—Test circuit for frequency response measurements

The voltage and power levels used to make frequency response tests should never exceed the rated values of the transformer. Voltage and power levels less than rated values may be used, but they should not be less than about 1% power (10% voltage) at the low-frequency end of the rated frequency range in order to avoid the effects of low initial permeability of core materials. Also, a full voltage test at the lowest rated frequency should be included to ensure that core saturation is not a problem.

If the primary (or secondary, or both) is required to carry unbalanced components of direct current, means should be provided to inject the required values of direct current into the windings. Figure 29 shows one method of accomplishing this.

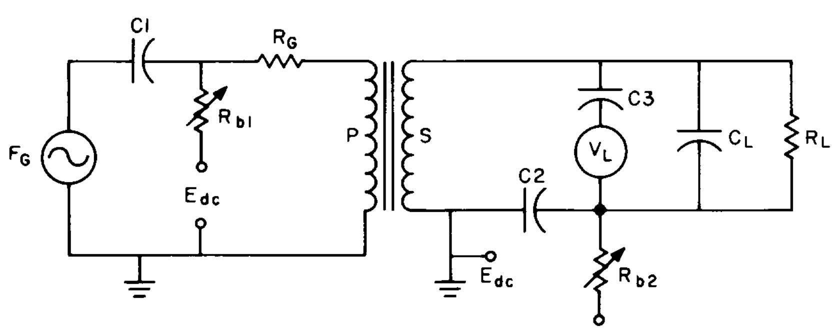


Figure 29—Test circuit for frequency response measurements with dc current(s)

In figure 29 the direct-current supply voltage E_{dc} is chosen to be high enough to supply a sufficient bias current to primary P when the resistance of R_{b1} is made high enough that it will not unduly load the variable-frequency generator F_G. Capacitor C1 should have negligible reactance at the lowest test frequency. If the secondary winding must carry a direct current, the value of E_{dc} must be high enough to supply the required bias current through the secondary winding resistance plus the load resistance. Capacitor C2 should have a low reactance at the lowest test frequency relative to the resistance of R_L. Capacitor C2 is required if voltmeter V_L has no internal blocking capacitor. The frequency response of a single-ended to push-pull transformer should include separate response curves for each half-secondary winding to show the extent, if any, of differences in response at the high-frequency end of the response curves. Differences in response at that end of the characteristic may be due to differences in capacitance couplings from the primary to each half-secondary or differences in leakage inductances from the primary to each half-secondary, or both.

Likewise, for push-pull to single-ended transformers, frequency response tests include tests made with first one half-primary excited and then the other half-primary excited for the same reasons as given for single-ended to push-pull transformers. These frequency response tests are in addition to tests of the frequency response with 180° out-of-phase balanced input voltages to the two primary windings. Separate resistors should be used to simulate the driver internal resistances. Here again, if negative feedback from the secondary of 10 dB or more is provided in the circuit in which the transformer is to be used, the internal resistance of the driver stage is thereby made so low that it can be neglected in a test circuit.

When the driver stage is of the push-pull amplifier type known as class AB or is a class B amplifier, the variable-frequency generator circuit should be modified to simulate the type of amplifier stage in which the transformer is to be used. A class B type amplifier may be simulated by inserting a diode in series with each resistor, which simulates the internal resistance of the driver stage. The rectifier polarities should be the same as they would be to form a full-wave rectifier circuit. For class AB operation *without* negative feedback from the output terminals, the combination of the diodes and a direct-voltage bias may be used to simulate the driver amplifier. However, if the application is a class AB amplifier with negative feedback from the output terminals, the best way to make frequency response tests is to use a low-power-level amplifier of approximately one-tenth the voltage output of the rated amplifier for the high-frequency response tests and then rely upon the single-ended response tests at rated power for determining any core saturation effects at the lowest rated operating frequency.

An alternate method of measuring transformer response is by using the circuit shown in figure 30.

NOTE—By keeping the generator voltage constant when the frequency varies, the generator impedance is not critical. Also, to eliminate voltmeter errors, it is recommended that the same voltmeter for E and V_{out} readings is used.

To measure the transformer loss, also referred to as insertion loss at f_m (figure 31), set the generator level E at a convenient decibel value and then switch the voltmeter to read V_{out}:

$$\text{transformer loss at } f_m = 20 \log \frac{E}{V_{out}} + 10 \log \frac{R_L}{4R_g} \quad \text{[dB]}$$

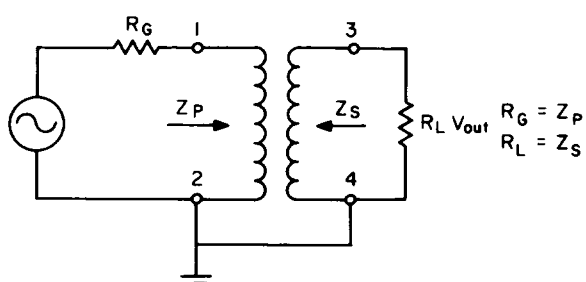


Figure 30—Alternate method for frequency response measurements

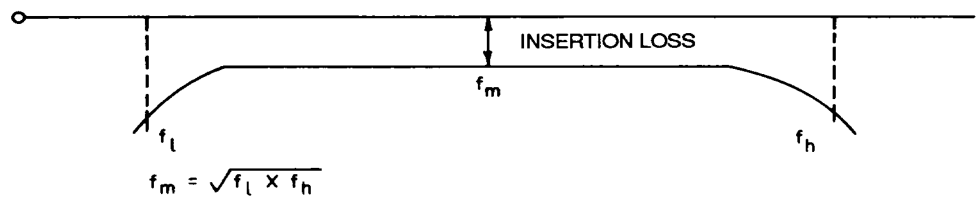


Figure 31—Measurement of transformer loss

To measure transformer loss at any other frequency with respect to the loss at frequency f_m, set V_{out} at a convenient voltage reference, switch voltmeter to check generator voltage *E*, then change frequency and adjust generator voltage for the voltage reading *E*. Switch voltmeter to the output of the transformer and read the voltage V'_{out}. The transformer response at that frequency is the difference between V_{out} and V'_{out}, or

$$\text{response} = 20\ \log\frac{V_{out}}{V'_{out}} \quad [\text{dB}]$$

When conducting all frequency-response tests, an oscilloscope should be used to observe the waveform of the output (secondary) voltage across the load. Any distortion of this voltage wave should be noted. The frequency at the high-frequency end and that at the low-frequency end at which the distortion sets in should be recorded.

If the phase angle of the output (load) voltage with respect to the input (generator) voltage is required, an *X-Y* oscilloscope display may be used to obtain data for plotting a curve of frequency versus phase performance of the transformer in the amplifier circuit for which it is designed to operate.

11.2 Transformer pulse response

When transformers are designed to operate between a source of pulse voltages and a load, they should be designed to be capable of operating over a wide range of frequencies, but they are usually tested in the time-domain mode, i.e., by determining their pulse responses, as this kind of information is most useful to the system designers. Pulse transformers may perform one or more of the following functions:

— Isolation between the source of pulse power and the load
— Impedance matching of the load to the pulse power source
— Phase (polarity) inversion
— Carrying low-frequency power to a cathode heater from a low-voltage insulated transformer

For more detailed evaluation of pulse transformers and of magnetic materials under pulsed excitation, consult IEEE Std 390-1987 and IEEE Std 393-1991.

Although much information regarding the ability of a pulse transformer to provide its rated performance in a specified system can be had by low-level tests in a circuit simulating the operations conditions, a final evaluation of the design and construction may require a full-scale test in the system in which it is to be used. The performance of the pulse transformer is evaluated by displaying the output (secondary) voltage of the transformer under rated load and driven by a specified power source. In those applications where the nonlinearity of the load impedance versus voltage curve exerts a large effect upon the waveshape of the output voltage, the load versus current waveshape should also be displayed. The test results consist of photographic recordings of the calibrated oscilloscope display or curves plotted point by point from that display.

Pulse transformers usually are driven by either a so-called hard-tube type pulse generator (amplifier), which also includes transistor-type amplifiers, or by a delay-line type pulse generator consisting of a pulse-forming network in which energy is stored during the interpulse period and discharged into the primary of the pulse transformer during the power pulse period through an electronic switch such as a thyratron or a thyristor.

The load on a pulse transformer may be the control element of an electronic device such as an electron tube grid, a transistor base, a thyristor gate, a magnetron, a klystron, a traveling wave tube, an amplitron, or an X-ray tube. All those listed require a unidirectional pulse voltage and many have negative-slope impedance versus voltage characteristics; i.e., as the voltage impressed upon them increases, the load impedance they present to the transformer secondary decreases.

For the purposes of low-level testing of pulse transformers, the pulse generator can be simulated by a unidirectional pulse generator of low internal impedance, the rise time of which can be controlled from a fast rise time of about 20 ns over a range to a slower rise time of about 1 µs, or even longer for long-pulse applications. The overshoot of the rise time should be less than 2%. The top of the pulse should be flat with a ripple and droop of less than 0.5%. The pulse width should be continuously adjustable from 0.1 µs to 100 µs. The pulse output voltage should be at least 10 V into a 50 Ω load and preferably 50 V into a 50 Ω load. The fall time at the end of the pulse should be no more than double the rise time. The rated internal resistance of the system generator is then simulated by a series resistor between the test generator and the transformer primary. A fast-recovery diode in series with this resistor is required to simulate hard-tube and transistor-type pulse generators.

The various characteristics of loads into which pulse transformers operate can be simulated reasonably well at low levels by a shunt capacitor, a fast-recovery diode in series with a noninductive linear resistor, and, for magnetrons and negative-biased vacuum tube grids, a direct voltage bias supply.

Figure 32 depicts a low-level pulse response test circuit. Here diode D_G would be used to simulate a hard-tube or transistor-type switch, and for that type of pulse generator R_G is relatively low and is determined by the voltage-saturation portion of the switch device. For a so-called line-type pulse generator, resistor R_G is made equal to the characteristic impedance Z_0 of the pulse-forming network.

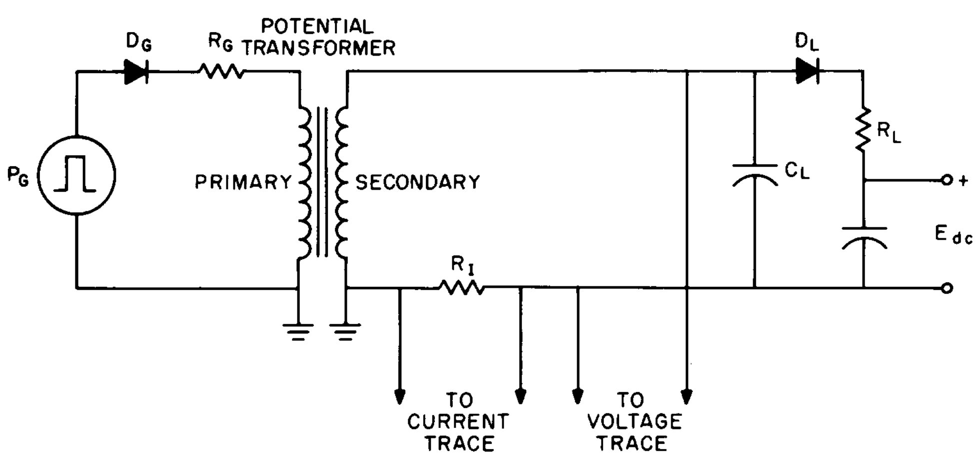


Figure 32—Test circuit for low-level pulse response

Referring again to figure 32, but to the simulated load portion, R_I is a noninductive resistor of the order of 0.1 Ω for supplying a voltage to the current trace of the oscilloscope. Capacitor C_L represents the shunt capacitance of the rated load device. Diode D_L simulates the rectifying properties of most pulse-type loads. Resistor R_L simulates the dynamic resistance of the load during the pulse. To simulate a magnetron of a biased-off grid load, the direct voltage E_{dc} represents the knee (Hartree) voltage of the magnetron or the grid bias voltage of an amplifier tube.

Low-level pulse tests using simulated pulse generator characteristics and load characteristics are quite accurate for measuring the rise time, overshoot, and ripple of the pulse voltage supplied to the load at full power levels. However, for determining the voltage droop, pulse fall time, and back-swing characteristics, the non-linearity of the magnetization characteristics of low pulse voltage levels are not the same as they are at high voltage levels, hence low-level tests for those characteristics may be subject to substantial errors.

12. Noise tests

12.1 Test conditions for audible noise

The transformers shall be mounted in an enclosure having a sound level of at least 4 dB, and preferably 7 dB or more, lower than the sound level of the transformer and the ambient combined. The ambient sound level shall be the average of the measurements taken immediately before and immediately after the transformer is tested at each of the locations as indicated in 12.2 item b). Corrections shall be applied in accordance with table 3.

The enclosure should be free of any noise-reflecting surface. Whenever possible, the transformer should be bolted on the chassis or other mechanical structures on which it is to be permanently mounted during operation.

Table 3—Sound-level corrections for noise tests

Difference between sound level of transformer and ambient combined and sound level of ambient (dB)	Correction to be applied to sound level of transformer and ambient combined to obtain sound level of transformer (dB)
4	–2.2
5	–1.7
6	–1.3
7	–1.0
8	–0.8
9	–0.6
10	–0.4
Over 10	0.0

The transformer is to be energized at rated voltage and frequency, and at full load or with no load as specified.

12.2 Measurement of audible noise

a) Sound levels shall be measured with an instrument that is in accordance with ANSI S1.4-1983. Response curve A (for 40 dB sound level) shall be used.

b) Measurements shall be taken with the probe of the sound-level meter located not more than 30 cm from the surface being measured. The readings shall be taken at the center of each of the vertical planes of the transformer and at the center of the top horizontal plane.

c) The average sound level is defined as the arithmetic mean of the sound levels measured according to item b).

d) Electromagnetic interference (EMI) testing shall be performed on the total system utilizing a transformer or inductor, and cannot be effectively performed on a transformer or inductor as a component.

13. Terminated impedance measurements

13.1 General

Terminated impedance measurements may be required when it is necessary to limit reflections on a transmission line to a minimum value. The degree of mismatch is given in terms of the reflection coefficient between the impedance of the line and the impedance looking into the transformer when it is terminated in its nominal termination. The reflection coefficient may be determined from series impedance measurements, from parallel measurements using an appropriate impedance bridge, or by means of return-loss measurements described in 13.2.

13.2 Return-loss method

A rapid and convenient method of making return-loss or reflection coefficient measurements requires an appropriate return-loss bridge. The bridge may take the form of the examples shown in figure 33.

To obtain highest accuracy, an impedance match should exist at each port except for the unknown port. The output voltage at the detector is directly related to the mismatch and therefore the reflection coefficient at this port.

The two-core hybrid bridge of figure 33 (b) is the most complex but permits matching of arbitrary impedance at input, detector, standard, and unknown ports. The single-core hybrid bridge of figure 33 (c) permits matching of arbitrary impedances at input and at standard and unknown ports. Unless an additional transformer is used at the detector port, the detector impedance should be half the standard impedance. In the resistance bridge of figure 33 (d) and the reflectometer of figure 33 (e) the balance is derived from the resistors of the resistance bridge rather than the balance of the transformer. The transformers in the latter two cases are unbalanced-to-balanced devices to permit grounding both the signal source and the detector. In the case of the resistance bridge, the transformer permits arbitrary match at the input. The reflectometer requires equal impedance at all ports.

The test method is similar for all devices of figure 33. Figure 34 shows that the test procedure requires four transmission voltage ratios to be measured at each test frequency.

Let

$$\begin{aligned} T_1 &= 20\ \log\frac{E_0}{E_1}\quad \text{[dB] for position 1} \\ T_2 &= 20\ \log\frac{E_0}{E_1}\quad \text{[dB] for position 2} \end{aligned} \tag{46}$$

For a resistive bridge or resistance reflectometer the average, $T_0 = \frac{(T_1 + T_2)}{2}$, will be approximately 12 dB if the value R_s is the design value and will be greater as R_2 deviates from the ideal value.

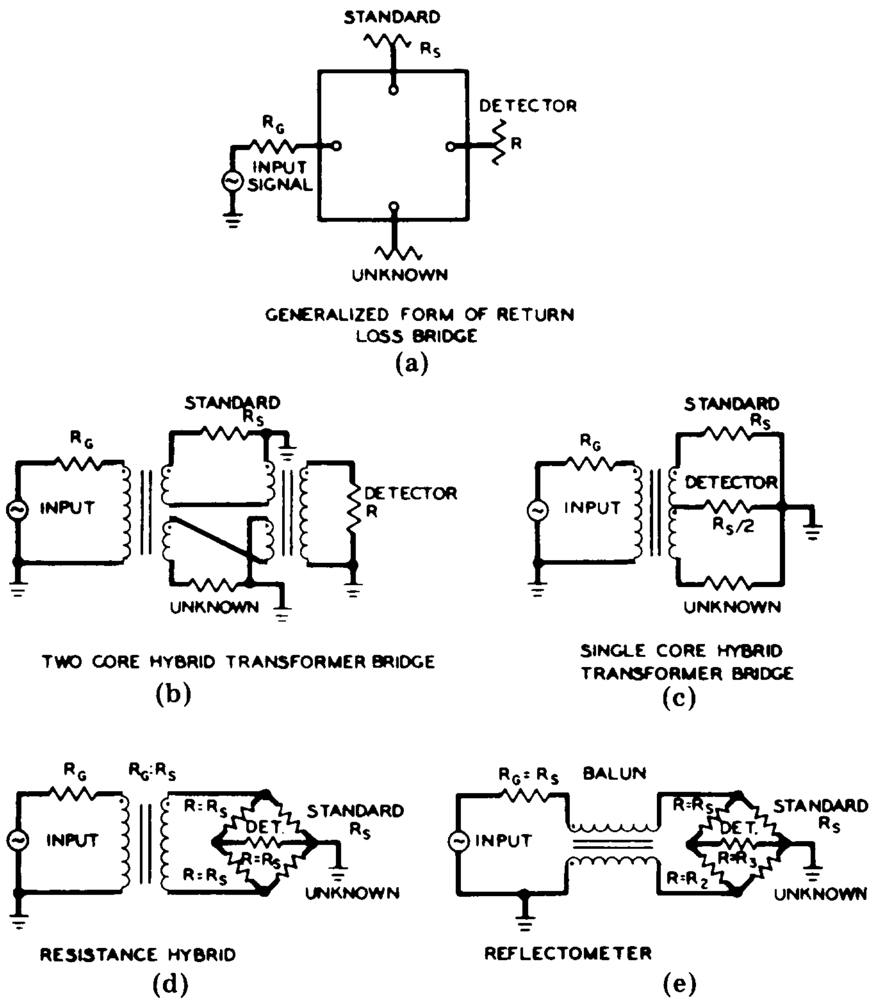


Figure 33—Test circuit for frequency response measurements with dc current(s) return-loss bridges

For hybrid transformer bridge, T_0 will be about 6 dB ideally, and will also increase as R_s is changed from the design value. T_3 in position 3 is a check measurement to test the balance of the bridge. The difference $(T_3 - T_0)$ is a measure of the precision of the bridge. This difference should be 20 dB greater than the requirement of return loss imposed on the transformer to obtain a reasonable accuracy of measurement.

T_4 is obtained when the transformer under test is connected to the "unknown" terminal of the bridge. The return loss RL is given by

$$\mathrm{RL} = T_4 = T_0 \quad [\mathrm{dB}] \tag{47}$$

for the case where R_1, R_2, and R_s are the values for which the bridge was designed.

It is possible to obtain satisfactory accuracy in certain cases merely by measuring the output voltage E_0, which for the four conditions will be designated e_1, e_2, e_3, and e_4. The voltage e_0 corresponds to T_0 in the preceding case and is determined from the formula

$$e_0 = \sqrt{e_1 e_2} \tag{48}$$

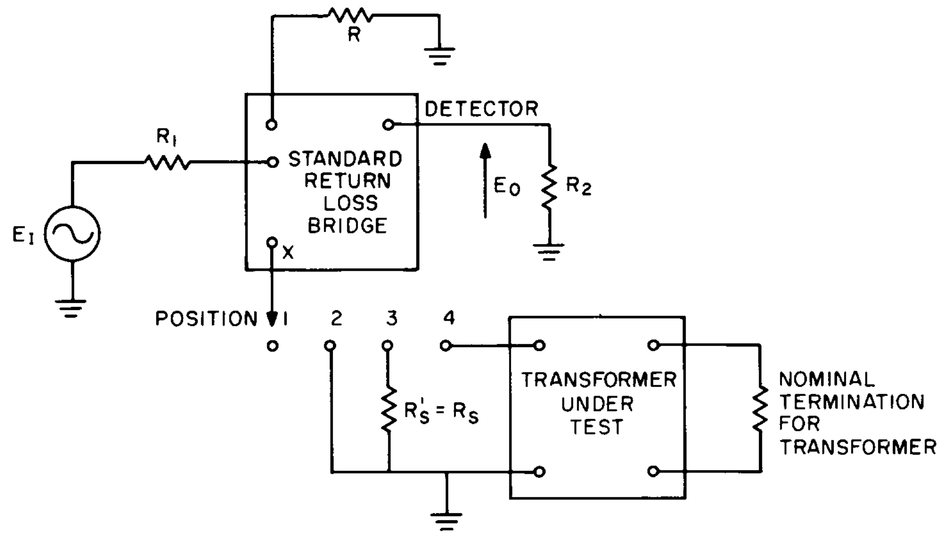


Figure 34—Return-loss measurement

e_3 provides a test of the bridge and should always be no greater than $\frac{e_4}{2}$. The return loss RL is given by

$$RL = 20 \log \frac{e_0}{e_4} \quad [dB] \tag{49}$$

NOTE—Errors will always be in the direction to show optimistic values of return loss.

The error will be on the order of $|(T_2 - T_1)/2|$ due to mismatch and will vary from approximately 1 dB if the balance of the bridge is 20 dB better than that measured from the transformer, to 6 dB if the bridge balance is equal to the balance of the transformer.

14. Temperature rise tests

14.1 Test methods

The maximum temperature rise of transformers or inductors can be measured by imbedding a thermocouple in the center of the coil and measuring the temperature with a thermocouple meter.

If it is not possible to insert a thermocouple in the coil, the temperature rise can be determined by measuring the resistance change of the inside winding coil.

Since the resistance of conductor materials changes with temperature, the resistance of the coil winding will change with temperature.

When thermal stability is achieved, disconnect the power source and load, and measure final (hot) dc resistance of the winding.

NOTE—Thermal stability can be defined to occur when two consecutive measurements of winding temperature, taken 30 min apart, are the same, within 1 °C.

To determine the mean temperature rise of the coil winding, use the equation

$$T_r = \frac{R_h - R_c}{R_c}(T_s + K) - (T_f - T_s) \tag{50}$$

where

T_r	is winding temperature rise, in °C
T_s	is ambient temperature at start of test, in °C
T_f	is ambient temperature at end of test, in °C
R_c	is “cold” dc resistance at T_s
R_h	is “hot” dc resistance at T_f
K	is thermal coefficient of resistivity: 234.5 for copper; 226 for aluminum

Since it is difficult to measure the resistance of a winding while power is still applied to that winding, several resistance measurements shall be made after power interruption at predetermined time intervals to determine the resistance at zero time, R_h, as, for example, shown in figure 35. For further information, see IEEE Std 119-1974.

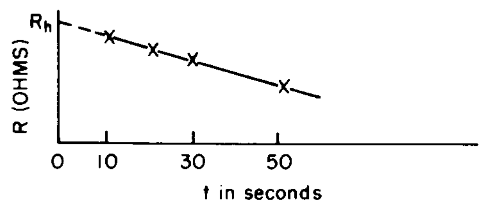


Figure 35—Determination of resistance at zero time

Temperature rise can be measured without interruption of power by installing a bifilar winding (usually on a single layer), inside the coil. The bifilar winding is connected in series opposition, so that induced voltages cancel.

This permits continuous monitoring of the “hot” dc resistance while the transformer is under load.

NOTE—In small high-frequency transformers, two problems can arise, resulting from noise pickup, or from a change in the apparent resistance of the sense winding. Errors in the temperature calculation can therefore be avoided by momentarily turning off power while taking readings of the “hot” value of resistance.

14.2 Notes on the technique of measurement

14.2.1 Ambient temperature

In the case of natural convection, the transformer should be tested in an area free from drafts. Where practical, ambient temperature should be measured in close proximity to the surface of the transformer (typically 15–30 cm from its surface). In general, the rise in ambient, $T_f - T_s$, should not exceed 5 °C, which can help increase the accuracy of the final calculation of temperature rise. The test document should record the exact location of where the ambient temperature was measured.

14.2.2 Physical environment

In many instances, an accurate measurement of temperature rise will depend on a close simulation of the physical environment of the inductive component. This can be especially important in a high-frequency power supply that operates at high levels of power density (rated watts per unit volume).

In the case of forced convection, tests should be performed under conditions that simulate the layout of components and with the specified rate of air flow.

In the case of conductive heat transfer, it is desirable that the ambient temperature be measured in a region of the heat sink (e.g., base-plate of chassis) that is in close proximity to the mounting area of the transformer.

14.2.3 Thermocouples

Great care should be taken in the method of attaching a thermocouple to a region inside a coil or to a surface. To ensure repeatability of measurements, strive to minimize thermal resistance by obtaining intimate contact at the interface. The test documents should state the exact location of each thermocouple, as well as the technique used in attaching the thermocouple to the surface.

CAUTION

The best thermocouple attachment is often with minimal insulation; therefore, a large voltage difference may exist between multiple thermocouples or to ground, creating a possible shock hazard.

NOTE—The magnetic field of the transformer can induce eddy currents in a thermocouple, and thereby add undesirable parasitic losses. Precautions should therefore be taken to determine that the size and material of the thermocouple do not introduce an unacceptable error in the temperature reading.

14.2.4 Lead resistance

Substantial errors in temperature rise measurements can be introduced by the attachment of leads to a winding that has very low resistance (in the order of milliohms). Avoid the use of alligator clips, which are often responsible for a variation in contact resistance, and can make it difficult to repeat measurements of the temperature rise. Keep exterior leads short. Be sure to subtract the resistance of lead wire, so as to establish the actual resistance of a winding.

15. Self-resonance

15.1 General

Self-resonance in a transformer or inductor results from the interactions of the various distributed self-inductances and capacitances normally present due to its physical construction. An examination of the midband and high-frequency equivalent circuits will show that there are several possible self-resonance conditions, either series or parallel resonance. Of particular concern are those within the frequency band covered by the device's application, although in some cases out-of-band self-resonances can cause parasitic damped oscillations or other effects. Self-resonance may manifest itself in a wideband transformer as an irregularity in frequency response or reflected impedance, or in a wave-filter inductor as an undesired response characteristic.

15.2 Measurement

Self-resonance in an inductive device is indicated by a peak in impedance for a parallel resonance and by a dip in impedance for a series resonance. While the usual means of measuring impedance described in clause 13 may be used, they are cumbersome to use in determining frequency of resonance. The technique described in this clause lends itself to a quick determination of self-resonant frequencies.

Commercial equipment is available giving a direct reading of impedance as well as impedance angle using techniques similar to that described here. In most cases, however, only the type of self-resonance and the frequency at which it occurs is needed to determine the probable cause or causes. The technique to be described should be adequate for most situations.

The technique shown in figure 36 consists of monitoring the voltage across the device in an essentially constant-current circuit for maxima and minima in voltage as frequency is varied. While the technique does not necessarily apply the normal working voltage across the device, this should cause little problem provided that the device is operating in a nonsaturated state.

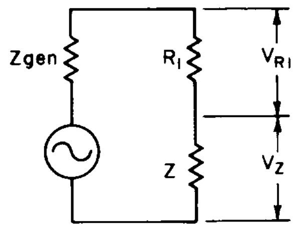


Figure 36—Measurement of self-resonance

The circuit in figure 36 utilizes an oscillator capable of covering the frequency range of interest and having an output voltage sufficient to provide readable voltages on the voltmeter at the minimum frequency of interest. The total of the source resistance of the oscillator in series with external resistor R_1 should be about 50 times the highest expected value of the open-circuit impedance of the device. The frequency is varied, so there is a need to watch for changes in direction of meter deflection. A maximum in voltage indicates a parallel resonance; a minimum indicates a series resonance. The types of resonances and the frequencies at which they occur should be noted.

The approximate impedance can be calculated using the following formula (the error will increase as Z approaches R_1):

$$Z \approx \left(\frac{V_z}{V_R^1}\right) \cdot R_1 \text{ for } \mathrm{R}_1 \geq 50 Z_{\mathrm{gen}} \tag{51}$$

16. Voltage-time product rating

16.1 General

The voltage-time product rating of a magnetic component is intended as an aid in pulse power conversion applications.

The voltage-time product is often determined by observing the pulse-exciting current on an oscilloscope or recorder. In that case, the voltage-time product of a winding is specified as the maximum voltage-time integral of a rectangular voltage pulse that can be applied to the winding before core-saturation effects cause the resultant exciting-current pulse waveform to deviate from a linear ramp by a given percentage (see 16.2).

Generally, the applied pulse-voltage amplitude shall be that used in 11.2. It shall not exceed the test voltage used in the induced voltage electric strength test (see 5.3), and the pulse-repetition frequency shall be kept sufficiently low to prevent transformer heating that could affect the magnetic characteristics or damage the transformer. The recommended pulse duration for the determination of this rating is that for which the transformer was designed. The voltage-time product does not remain constant when the pulse duration is very short. If the transformer application involves a direct current in any of the windings, the measurement shall be made with rated direct current ampere turns and polarity applied through a suitable impedance to prevent pulse-loading effects. Input pulse source impedance and pulse parameters shall be specified for this test. The preferred test method is described in 11.2.

16.2 Recommended voltage-time product test methods

16.2.1 Rectangular voltage pulse applied from low-impedance source, and pulse-exciting current response observed

NOTE—This test method is nearly the same as the recommended test method for open-circuit parameters (magnetizing pulse inductance and exciting current) normally used for pulse transformers.

The voltage-time product of a winding shall be measured as the product of the pulse-voltage amplitude of a rectangular voltage pulse applied to the winding and the pulse duration measured at a time t_2 where the resultant pulse-exciting current has increased to a certain value that is a specified percentage (e.g., 50%) above the linearly extrapolated value of the linear ramp waveform. The departure of the exciting-current waveform from a linear ramp is due to a rapid decrease in magnetizing pulse inductance as the core flux reaches the saturation region (figure 37).

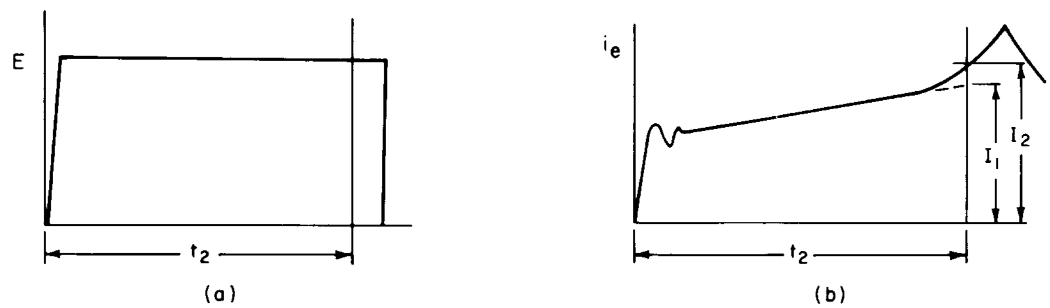
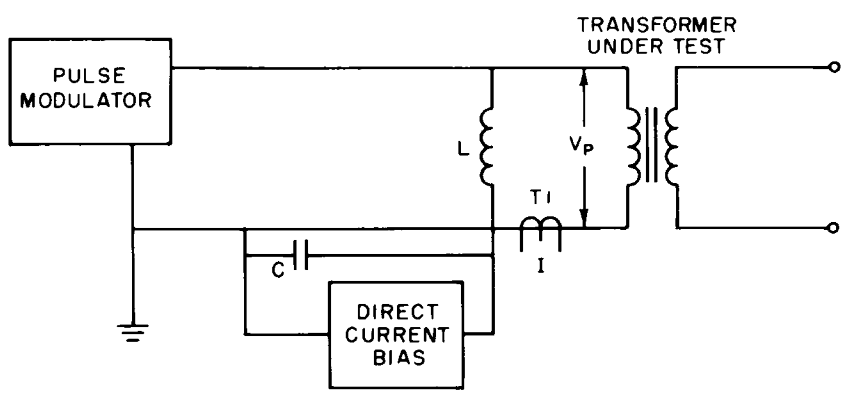


Figure 37—Waveforms for voltage-time product test method
(a) Applied voltage, (b) Resulting magnetizing current

The droop of the applied voltage pulse shall be less than 2%.

16.2.2 Rectangular voltage pulse applied from a pulse-modulator source

An example of a recommended test circuit for rectangular pulse excitation is shown in figure 38. In general, a pulse source simulating the circuit in which the transformer will be used shall be employed in the test setup.



L is isolation inductor
C is bypass capacitor
T is current probe
V_p is adjusted to required transformer primary voltage

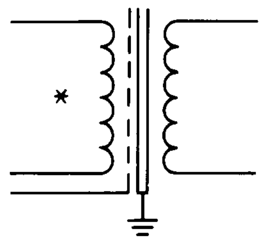
Figure 38—Test circuit for rectangular pulse excitation

17. Shielding

17.1 Electrostatic shielding

17.1.1 Symbols

The basic electrostatic symbol is shown in figure 39. Other possible variations of shield configurations are shown in figures 40 through 45.



*Shielded winding

Figure 39—Basic electrostatic symbol

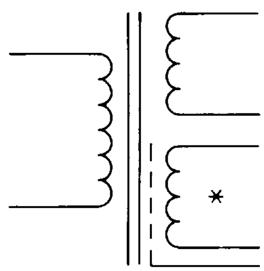


Figure 40—Shielded single winding, core floating

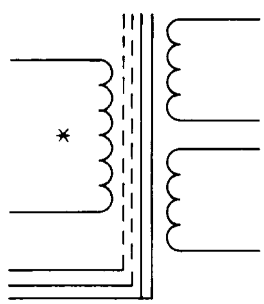


Figure 41—Multiple shielded single winding, core terminal (lead) provided

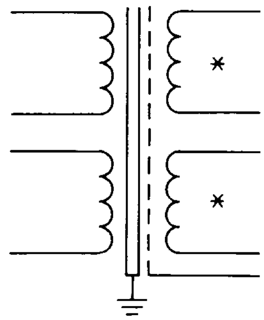


Figure 42—Shielded two-winding, secondary core grounded

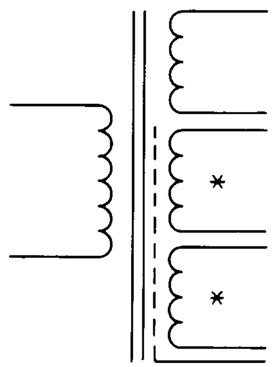


Figure 43—Shielded group of windings, core floating

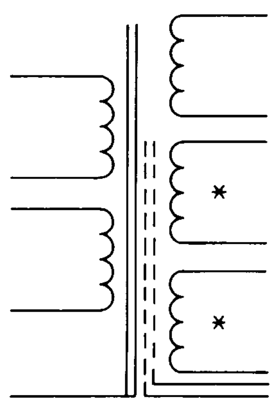


Figure 44—Multiple shielded group of windings, core terminal (lead) provided

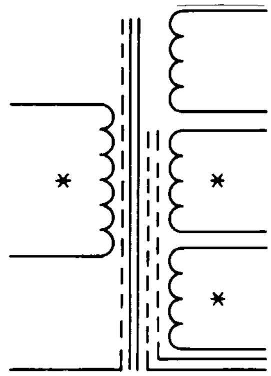


Figure 45—Combination of shielding conditions

17.1.2 Theoretical discussion

A typical two-winding shielded transformer may be represented as shown in figure 46, where the capacitance has been "lumped" at the center of each winding, and the core has been omitted for clarity. C_1 and C_2 are the capacitances between the shield and the primary and the secondary, respectively, and C_d is the direct interwinding capacitance. The representation may be further simplified as shown in figure 47.

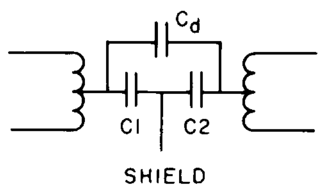


Figure 46—Typical two-winding shielded transformer

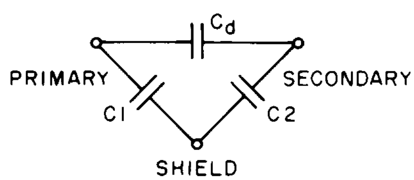


Figure 47—Simplified representation of figure 46

The purpose of electrostatic shielding of transformers is to minimize the effect of C_d, the direct interwinding capacitance, since it is via this capacitance that unwanted signals are coupled between windings.

The degree of shielding obtained depends on how the shields are incorporated in the transformer and how they are connected in the circuit.

Multiple shielding may be used to isolate specific capacitances to a particular point or points in the circuit in addition to limiting common-mode coupling. This will provide some "control" of these stray capacitances that may be effective in particular applications.

The effectiveness of shielding is generally described in terms of direct capacitance between windings, or as attenuation in decibels of unwanted signals. When measuring transformers for common-mode rejection, it is advisable to simulate the circuit in which the transformer is intended to be used.

17.1.3 Measurement methods

17.1.3.1 Indirect method

The recommended method is the indirect method shown in figure 48.

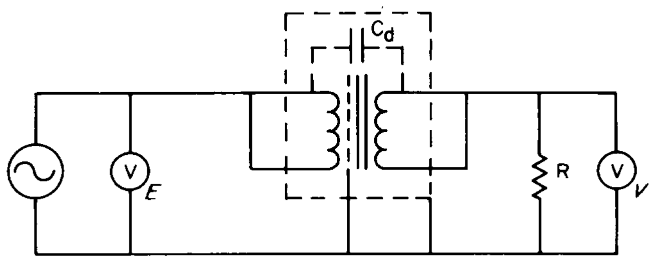


Figure 48—Indirect measuring method for electrostatic shielding

The direct capacitance C_d is

$$C_d = \frac{V}{\omega R E} \tag{52}$$

where

V	is coupled voltage
ω	is frequency of generator, in radians per second
R	is load resistor (including voltmeter loading)
E	is generator voltage

If, for example, $\omega = 10^5$ (f = 15.9 kHz), E = 100 V, and R = 10 kΩ, then a V of 1 mV corresponds to a C_d of 0.01 pF.

In practice there is a shunt capacitance C_s across R, and a shunt resistance R_i across C_d. Given the values stated above, C_s may be as high as 100 pF. If $R_i = 10^{10}$ Ω minimum, the system will measure values of C_d between 0.01 pF and 100 pF.

17.1.3.2 Direct method

An alternate method, which may be used for measuring values of C_d above 1 pF, is direct reading with a three-terminal capacitance meter. The shield is connected to ground and the windings in question to high and low.

It has the following limitations:

- — Stray capacitances and sensitivity limit applicability to direct (coupling) capacitance above 1 pF
- — Grounded winding with floating core may be awkward to implement mechanically

NOTE—Either of the test methods may be used to obtain the ratio of capacitance with and without shield by adding a switch in the shield lead:

$$RSE = \frac{C \text{ shield floating}}{C \text{ shield connected}}$$

17.2 Magnetic shielding

Magnetic shielding is specified as required either to reduce a magnetic flux field emanating from a transformer (or inductor) or to reduce a magnetic flux field entering a transformer (or inductor). Magnetic shielding is utilized, when needed, for magnetic flux with a frequency in the range of zero to approximately 10 kHz.

When an emanating flux is being reduced, the specification should define the maximum flux-density in millitesla (or in gauss) adjacent to any surface of the device at a specified distance. The flux-density level can be measured with a high-impedance detector (greater than or equal to 10 MΩ input impedance) and a magnetic search coil type of pickup probe (see annex C for description) or with a Hall-effect type gaussmeter. Because the flux emanating from the transformer (or inductor) may not be perpendicular to its surface, the flux measuring device (or the transformer or inductor) should be rotated to obtain the point of maximum emanating flux. If the magnetic shield is removable, the above technique may be utilized to determine the magnetic shield effectiveness by comparing measurements with and without the magnetic shield in place. If the transformer or inductor under test normally is operated with a direct current flowing in a winding, the test should be performed with that current flowing.

If an external magnetic flux entering the transformer (or inductor) is to be reduced by magnetic shielding, the shielding specification relative to the specific component may define either item a) or b):

a) Specify the reduction in decibels that the magnetic shielding gives at a specified flux density at a specified frequency or frequency range. Tests of the shield effectiveness may be made on a sample of the transformer made with its magnetic shielding and an identical sample made without its magnetic shielding. A high-impedance voltmeter is connected to the highest voltage or highest impedance winding of the transformer (or inductor) while that transformer (or inductor) is slowly rotated through the specified magnetic field until a maximum is observed. The ratio of the maximum voltage reading without shielding to the maximum voltage reading with shielding is a measure of the magnetic shielding effectiveness. The reduction may be stated in decibels by the formula

$$\mathrm{dB} = 20 \log \frac{E_{\mathrm{ns}}}{E_{\mathrm{ws}}} \tag{53}$$

where

E_{ns} is voltage reading without shielding
E_{ws} is voltage reading with shielding

b) Specify the maximum voltage reading when a high-impedance voltmeter is connected to the highest voltage or highest impedance winding of the transformer under test while the device is slowly rotated in the specified magnetic field.

18. Measurement of quality factor *Q*

18.1 Definition

The quality factor Q equals 2π times the ratio of the maximum stored energy to the energy dissipated per cycle at a given frequency. This can be shown to be equal to $\omega L_s / R_s$ for a coil represented as an inductance L_s in series with a resistor R_s, or $R_p / \omega L_p$ for a coil represented as an inductance L_p in parallel with a resistance R_p. The quality factor values vary from less than 1 to more than 1000.

18.2 Methods

- — Bridge measurements
- — *Q*-meter measurements
- — Insertion-loss measurements
- — Transmission method
- — Damped-oscillation method
- — Vector voltmeter measurements

18.3 Bridge measurements

Measurements made on "bridges" are impedance measurements where the real and imaginary parts of the impedance are given or calculable from the values of the bridge arms when the bridge is balanced. As noted in the definition, the ratio of the real and imaginary portions of the impedance or vice versa, depending upon whether the impedance is in series or parallel form, gives the value of Q, so that the bridge in its various configurations is ideally suited for making measurements of $Q \leq 50$. In general, many commercial bridges with configurations as outlined in clause 10 allow Q measurements to be made with up to a few volts across the inductor or transformer. In some cases this may restrict the flux-density level in the core to some low value (< 100 gauss). The various types of ac bridges differ only in the types of impedance standards used and in the relative positions in the bridge arms of the standards and the unknown impedance. Some usual bridge arrangements are shown in figure 27, which also gives the balance conditions in each case. The resonant bridge of figure 27(f) differs from the others in that the balance depends critically on the testing frequency, which has to be controlled to a greater degree of accuracy than is necessary for the nonresonant types. The bridge circuit of figure 27(f) is useful since it enables both the resistive and the inductive portions of the unknown impedance to be measured in terms of differences in the settings of the variable capacitors and while all the resistances are fixed. This is an advantage at high frequencies where an air dielectric capacitor forms a more satisfactory impedance standard than a variable resistor. This is a substitution bridge in which an initial balance is made with the "unknown" terminals short-circuited and the bridge balanced again with the unknown impedance inserted. The choice of the best bridge to be used for a given purpose depends upon the frequency range and the range of inductance and Q values that are to be measured.

NOTE—The balance of the bridge arms should be obtained at a specified voltage across, or at a specific current through, the device under test.

18.4 *Q*-Meter measurements

This instrument gives a direct reading of the Q of a coil or transformer. The winding under test is connected in series with a variable capacitor that forms the impedance standard. A known alternating voltage at the test frequency is applied across the tuned circuit (made up of the test winding and the variable capacitor), and the circuit is adjusted to resonance as indicated by a maximum reading on a voltmeter that is connected across the capacitor. The impedance of the capacitor will be almost a pure reactance that, at resonance, is numerically equal to that of the test coil. The ratio of the modulus of the voltage across the capacitor to that of the applied EMF to the circuit is thus $\omega L / R + R_s$ where R_s is the equivalent series-loss resistance of the capacitor. If R_s is negligibly small, then the ratio of voltages gives the Q of the winding under test. Since the method relies on the peaking of the voltage across the capacitor at resonance, two basic points are worth noting: (1) Q of less than 10 are barely detected and (2) Q of 1000 can produce up to 10 V across the test winding; this may result in relatively high flux densities in small cores. Commercial instruments are available with the generator, variable capacitor, and detector in a single box, and the arrangement of dials and indicators makes them quick and easy to use. The frequency range is 50 kHz to 50 MHz for one model and 20 MHz to 200 MHz for another. A low-frequency adapter device is available to extend Q measurements down to 1 kHz. A circuit diagram is shown in figure 49.

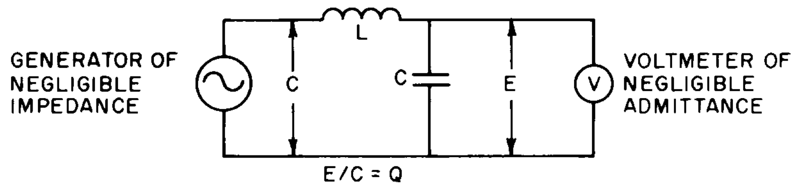


Figure 49—Circuit magnification for basic method

In the instruments mentioned, the voltage applied to the circuit is held constant at 10 mV (times 1 range), 20 mV (times 2 range), etc. For a Q of 100, the voltage across the coil is then 1 V or 2 V, depending on the range of Q selected. The output reading of the voltmeter is calibrated directly in values of Q.

18.5 Transmission method

The inductor to be measured is combined with a high-quality capacitor to form an LC circuit resonating at the measuring frequency. The effective resistance at resonance is measured by determining the attenuation of a transmission path containing the resonant circuit. The quality factor is calculated from this resistance and the reactance of the inductor measured under identical conditions, taking into account the losses due to circuit components.

18.5.1 Circuits (figure 50)

The generator G is variable in frequency and voltage. The resistance of each of the two resistors R_1 and R'_1 is equal to the characteristic resistance of the attenuator. Resistance R_2 is approximately twice the effective series resistance R_x of the resonant circuit (which must be estimated in advance), or 5 Ω, whichever is greater. The attenuator shall be able to resolve 0.1 dB with sufficient accuracy. The accuracy of this method depends mainly on the accuracy of the attenuator.

Voltmeter V_1 is a high-impedance ac voltmeter, and voltmeter V_2 is a tuned detector having an input impedance very much greater than R_1 and R_2.

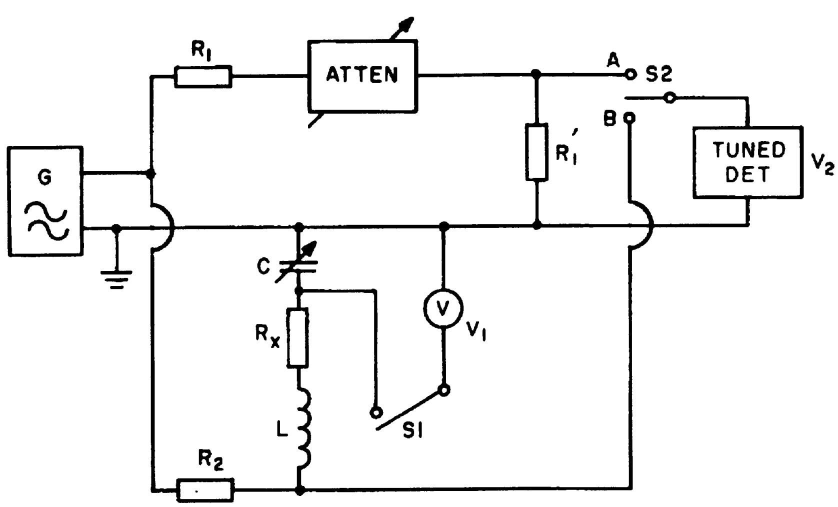


Figure 50—Transmission method

18.5.2 Measuring procedure

With switch S1 closed, the frequency of the generator is adjusted to the resonant frequency of the LC circuit as indicated by maximum deflection of V1.

If the frequency is then not sufficiently close to the specified measuring frequency, the capacitance of capacitor C shall be changed until the circuit resonates at the required frequency. The voltage of the generator shall be so adjusted to a reasonable value. The frequency is then readjusted to resonance if necessary and switch S1 is opened. With switch S2 in position B, the tuning of detector V2 is adjusted for maximum deflection. The capacitor C is then returned for minimum deflection of V2 (to compensate the opening of S1). The attenuator is adjusted until the detector indication is the same for either position of switch S2. The attenuator setting α (in decibels) is noted.

18.5.3 Calculation

$$Q_{LC} = \frac{\omega L}{R_X} = \frac{\omega L}{R_2}\left(2 \cdot 10^{\alpha/20} - 1\right) \tag{54}$$

where

Q_{LC} is quality factor of LC
L is circuit series inductance of inductor under test
R_X is equivalent series resistance of LC circuit
R_2 is ~2 (R_X) or 5 Ω, whichever is greater

Corrections for capacitor loss and for self-capacitance of the inductor are

$$Q_L = \frac{Q_{LC} Q_C}{Q_C - Q_{LC}}\left(1 + \frac{2C_s}{C}\right) \tag{55}$$

where

Q_L is corrected quality factor of inductor under test
Q_C is quality factor of capacitor
C_s is self-capacitance of inductor under test
C is capacitance of resonating capacitor

18.6 Damped oscillation method

The inductor to be measured is combined with a high-quality capacitor to form an LC circuit resonating at the measuring frequency. A pulse train is applied to this circuit, and the voltage across the inductor between the pulses is displayed on a calibrated oscilloscope. The quality factor of the inductor is calculated from the voltage ratio occurring at a certain time interval as measured on the oscilloscope, taking into account the losses due to circuit components (figure 51).

18.6.1 Accuracy

An accuracy of ±4% can be obtained provided the corrections for probe resistance and capacitor loss are relatively small.

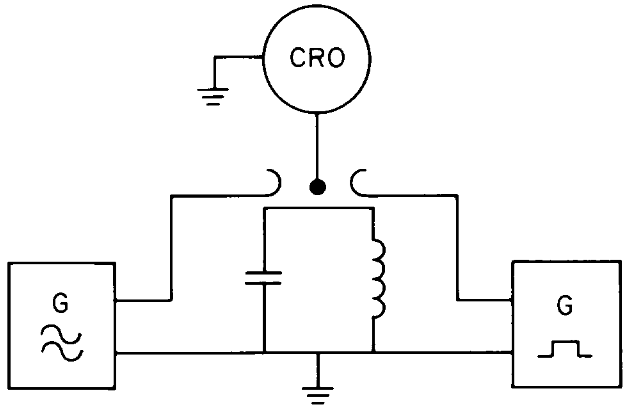


Figure 51—Damped oscillation method

18.6.2 Generators

18.6.2.1 Signal generators

The frequency range shall coincide with the frequencies at which measurement shall be made. The open-circuit output voltage shall be adjustable between 0 V and 10 V.

18.6.2.2 Pulse generators

The pulse width shall be between 10 µs and 100 µs, and it shall be possible to set the repetition rate at a value between 500 and 300 pulses per second. The open-circuit output voltage may, for example, be 100 V.

18.6.2.3 Antenna

The generators are coupled with the LC circuit by means of a wire placed approximately 20 mm from the connection between inductor and capacitor. To prevent damping of this circuit, the aerial coupling should be kept as small as possible.

18.6.3 Capacitor

The capacitance value shall be such that the resonance frequency of the LC circuit is sufficiently close to the specified measuring frequency. The quality factor should be as high as possible and be known as accurately as possible to allow for correction. It is recommended that an air dielectric capacitor be used.

18.6.4 Measuring circuit

18.6.4.1 Oscilloscope

The sensitivity shall not be less than 0.1 mV/mm. The relative inaccuracy of voltage measurement shall not exceed 1.5%.

18.6.4.2 Probe

The input impedance should be as high as is consistent with practical attenuation ratios. The following probe is recommended: attenuation ratio 10:1; input impedance 10 MΩ, 7 pF.

18.6.5 Measuring procedure

With the pulse generator switched off, the frequency of the oscillator is varied until maximum deflection is obtained on the oscilloscope. The frequency is noted and shall be sufficiently close to the specified measuring frequency. When this is not the case, the capacitance value shall be adjusted. The signal generator is then switched off and the pulse generator is switched on. Its antenna is so adjusted that a reasonable deflection appears on the oscilloscope. The sweep of the oscilloscope is adjusted to obtain a picture in accordance with figure 52 (a) (for measurement of high-quality factors) or with figure 52 (b) (for measurement of lower-quality factors). The voltage values at two points of time are read from the oscilloscope, together with either the time interval or the number of cycles between these two values.

Note—The Q-factor for excitation level sensitive inductors and transformers may depend on the position and magnitude of t (or n) for which the voltages U_1 and U_2 are measured. The higher the Q-factor and the higher the excitation level applied to the LC circuit, the stronger may be the dependence of the Q-factor on the position and magnitude of t (or n). If accurate Q-factor measurements are required, then the position and magnitude of t (or n) should be specified, or the voltage U_1 and U_2 should be measured during a later period of the dampened oscillations.

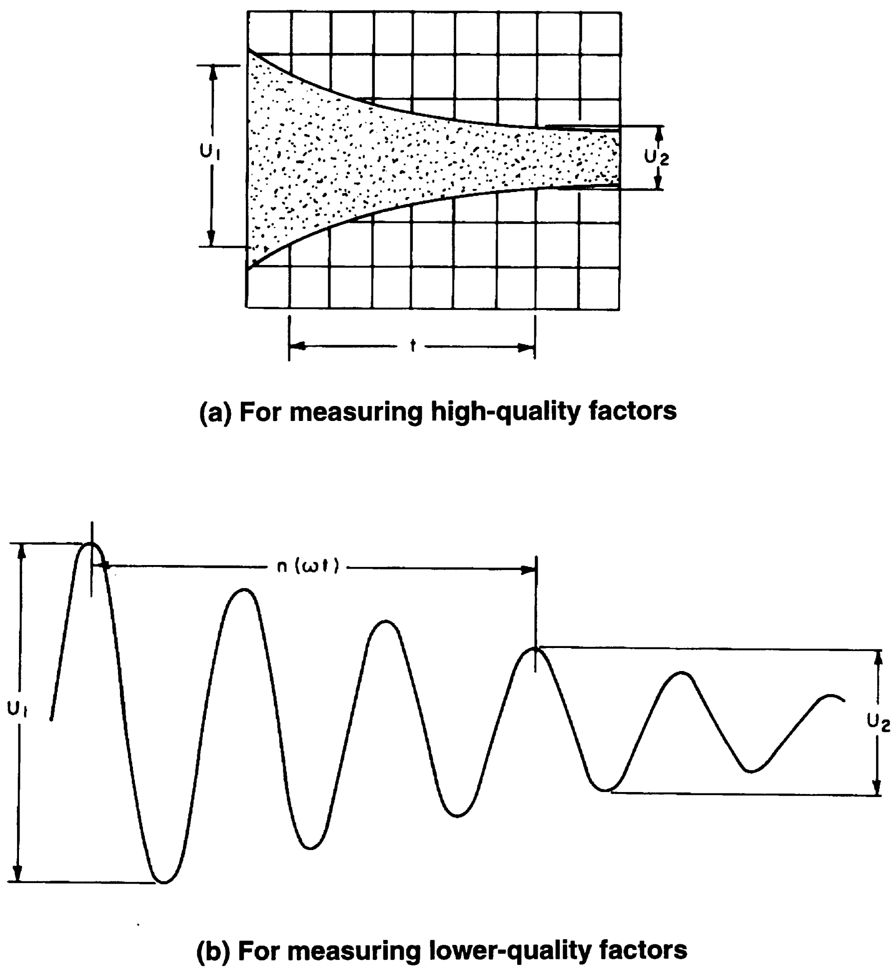


Figure 52—Oscilloscope sweep for damped oscillation method

18.6.6 Calculation

The high-quality factor is

$$Q_{\mathrm{i}} = \frac{\omega t}{2 \log \dfrac{U_1}{U_2}} \tag{56}$$

and the lower-quality factor is

$$Q_{\mathrm{i}} = \frac{\pi n}{\log \dfrac{U_1}{U_2}} \tag{57}$$

where

Q_{i} is uncorrected quality factor for LC circuit
ω is $2\pi f$, the measuring frequency

t, n, U_1, and U_2 are given in figure 52.

The correction for probe resistance is

$$Q_{\mathrm{LC}} = \frac{R_0 \cdot Q_{\mathrm{i}}}{R_0 - Q_{\mathrm{i}} \omega L} \tag{58}$$

where

Q_{LC} is corrected quality factor of LC circuit
R_0 is input resistance of probe
L is parallel inductance for inductor under test

The correction for capacitor loss is

$$Q_{\mathrm{L}} = \frac{Q_{\mathrm{LC}} \cdot Q_{\mathrm{C}}}{Q_{\mathrm{C}} - Q_{\mathrm{LC}}} \tag{59}$$

where

Q_{L} is quality factor of inductor under test
Q_{C} is quality factor of capacitor

19. Common-mode rejection test

The concept of the common-mode rejection test is illustrated in figure 53. The input terminals of the transformer are connected to the high terminal of a voltage source through equal impedances R1. The low terminal of the source is connected to ground reference. The voltage induced across the secondary winding (terminated with a specified impedance R2) is measured. The common-mode voltage transfer ratio is $r = e_2/e_1$.

The rejection of the common-mode (CMR) voltage is

$$\text{CMR} = -20 \log(e_2/e_1) \quad [\text{dB}]$$

If the primary or secondary windings of the transformer, or both, have center taps, these shall be either grounded or in a defined state of termination. The same applies to other windings of the transformer. The driving impedances R1 can be omitted, and the terminating impedance R2 can be equal to the impedance of the measuring instrument for e_2, if so specified.

NOTE—The following information must be stated: transformer connections, including earth connections; test frequency; input impedance (2R1); and load resistor R2.

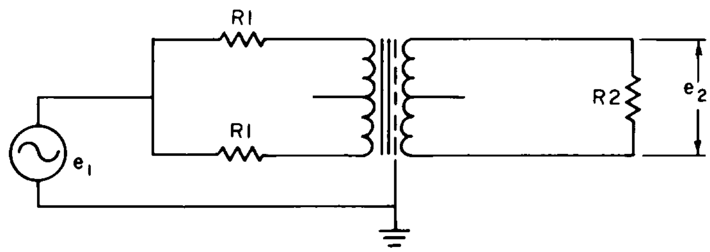


Figure 53—Common-mode rejection test

20. Inrush-current evaluation and measurement

20.1 Measurement

For small transformers, the inrush-current due to closure of the switch connecting the transformer primary to the supply lines can be measured by oscillographic or storage oscilloscope records of primary no-load exciting currents for a number of circuit closures. For single-phase transformers, 20 closures will probably be sufficient to approximate the maximum inrush current. For three-phase transformers at least 60 closures will be required to approximate the maximum value of inrush current. If a synchronous switch (with controlled but variable closure time with respect to power supply) is available, then the worst case (highest value) of inrush current can be obtained with many fewer closures if the closure time is adjusted to occur at the time of zero voltage crossing for single-phase operation and the number of conducting half-cycles is made an odd number. For three-phase operation, both the closure time and the conducting time of the other phases should be varied until a worst-case condition is obtained. Information relative to or control of the inrush current, or both, is of concern in determining the required fuse rating for overcurrent protection of all transformers and for determining the mechanical bracing of the coils of large transformers for rectifier power supplies. Since transformer manufacturers may not have ac power sources of sufficiently low impedance to determine inrush-current characteristics by test, means have been devised to determine these characteristics by calculation.

20.2 Calculation

Reference [B2] discusses the inrush-current phenomenon and suggests a method for calculating the peak value of the first current peak. The equation uses the magnetic characteristics of the core material, the operating flux density of the design, and certain dimensions of the core and of the primary coils, and assumes zero source impedance and zero winding resistance. Peak inrush current (in MKS units) can be found as follows:

$$i_{pk} = \frac{V_f - V_a}{\text{air core inductance of coil}} = \frac{2 \cdot B_m(NA_c) - (B_s - B_r)(NA_c)}{\mu_0 N^2 A_s / b} \tag{60}$$

where

V_f is volt-time product of first cycle applied
V_a is volt-time product absorbed by core
b is coil length (winding traverse of conductor), in meters
A_s is total area of space enclosed by the mean turn of excited winding, in square meters
A_c is net area of magnetic core, in square meters
N is number of turns of excited coil winding
B_r is residual flux density at time of voltage application, in tesla
B_s is saturation flux density, in tesla
B_m is design peak steady-state operating flux density, in tesla
μ_0 is magnetic constant of free space in henry per meter

The value to use for B_r is a function of the design value B_m and of the type of core fabrication used in the transformer. The following ratios are commonly accepted maximums for oriented silicon steel cores:

Core type	Ratio
Tape wound but uncut	$B_r = 0.90\ B_m$
Tape wound and cut	$B_r = 0.60\ B_m$
Laid-up strip with lap joints	$B_r = 0.80\ B_m$

Equation (59) applies to a single-phase transformer and to the coil current of delta-connected primary windings of a three-phase transformer or three single-phase transformers. For a three-phase wye-delta-connected bank of single-phase transformers and a three-legged core-type three-phase unit in wye connection, the inrush currents are approximately two thirds of the value for a single-phase unit.

NOTE—For single-phase core-type units with a coil on each leg, if coils are series-connected, N is total turns and b is $2 \cdot b$ of each leg; if coils are paralleled, i_{pk} is two times that for a single coil on one leg.

20.3 Other considerations

More extensive and detailed information regarding the inrush-current characteristics of transformers is presented in [B1]. That reference includes information regarding several factors affecting the magnitude of inrush currents such as source impedance, system voltage level, switching angle, etc. It also includes the characteristics that affect the duration of the inrush current, which is of importance with respect to the ratings of protective fuses. It reports that as a result of an extensive study of the inrush currents of representative samples from a complete line of transformers, under conditions designed to produce the greatest inrush current and the slowest rate of delay, it was found that for each transformer tested, and allowing a reasonable margin in excess of the actual measured I^2t, the fusing should be such as to allow 12 times the transformer full-load primary current to flow for 0.1 s without fuse blowout. For a more sophisticated determination of the I^2t values, rms values, and average values of inrush current, consult [B9].

21. Current transformer test

21.1 General

The term *current transformer* is used to designate a transformer connected in series with a source or load. Current transformers are used primarily as instrumentation devices for measuring or sensing current. A current transformer is a coupled circuit and all the principles and tests described in the preceding clauses can be applied. However, due to differences in construction, in voltage and current ratings, and in applications as compared to conventional voltage transformers, there are certain terms, tests, and specifications that are unique to current transformers.

It is often convenient to think of the current transformer as the inverse of the voltage transformer: the no-load condition of a current transformer is with the secondary short-circuited on itself; an open-circuited secondary can result in dangerous voltage levels; current transformers are almost invariably current step-down transformers, i.e., the primary has a very low number of turns, often only one turn; and being a series-connected device, a current transformer is designed for minimum voltage between primary terminals. There are certain other differences between current and voltage transformers. Accuracy of the ratio of primary to secondary current is usually the most important specification of a current transformer and is specified as both a magnitude and a phase-angle error or accuracy. The *load* on the secondary of a current transformer is an unwanted quantity (since it causes the ratio and phase-angle errors), and is referred to as the *burden* (see clause 3). The primary of many instrument transformers is often not an integral part of the transformer itself, but merely a lead of the circuit whose current is being measured. In electronics applications, current transformers are frequently applied to systems with pulsed excitation. Therefore, the terminology and definitions of clause 8, 11.2, and clause 16 of this recommended practice and those of IEEE Std 393-1991 should be reviewed.

21.2 Recommended test procedure for current-transformation ratio and phase angle

The complex ratio of transformation may be expressed as

$$\frac{I_1}{I_2} = N(1+a)e^{-j\beta} \tag{61}$$

where

I_1, I_2 are primary and secondary current phasors
N is their nominal ratio
a is correction to nominal ratio (small quantity)
β is phase angle between currents, in radians

In Cartesian form, equation (61) becomes, approximately, for β very small,

$$\frac{I_1}{I_2} = N(1+a-j\beta) \tag{62}$$

where

$(1+a)$ is ratio correction factor

A recommended test circuit is shown in figure 54. The primary of the current transformer is connected in series with a noninductive resistor R_1. The secondary is connected in series with a noninductive potentiometer R_2 and the primary winding of a suitable, variable, mutual inductor *M*. The resistors are connected in series with a suitable detector and the secondary of the mutual inductor. Note that the polarity of the windings of the current transformer shall be as shown so that the voltages across R_1 and R_2 are in series opposition with respect to the detector. The polarity of the windings of *M* should also be as shown.

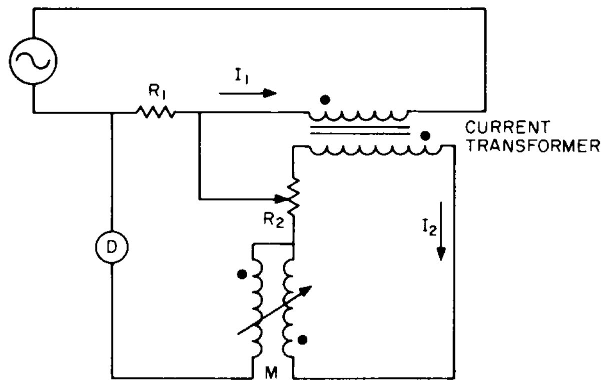


Figure 54—Test circuit for current transformers

Balance is obtained by adjusting R_2 and *M* for a null on the detector. At balance, the equations for the current transformation ratio and the phase angle are, to a first approximation,

$$\text{current transformation ratio} = \frac{R_2}{R_1}$$

$$\text{phase angle } \beta = \frac{\omega M}{R_2} + \theta_1 - \theta_2 \text{ [radian]}$$

where R_1 and R_2 are the values of the resistors (note that R_2 is the portion of that resistor in the detector loop), and θ_1 and θ_2 are the phase angles of I_1 and I_2, respectively.

22. Bibliography

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Annex A

(informative)

Instrumentation for voltage and current measurements on inductors and transformers

A.1 General

The unique nature of inductive devices having ferromagnetic cores requires a clarification of two terms associated with measuring alternating currents and voltages, that is, the meaning of *true rms* and *flux voltage* measurements.

A.2 True rms measurements

The magnitude of alternating current or voltage is traditionally defined by its root-mean-square (rms) value. The rms value is a measure of the energy content of the voltage or current. It is defined mathematically as

$$V_{\mathrm{rms}} = \sqrt{\frac{1}{T}\int_0^T (v(t))^2 \mathrm{d}t} \tag{A.1}$$

where

$v(t)$ is the instantaneous value of voltage (or current)
T is the period of its alternations

Unless otherwise identified (as average, peak, or peak-to-peak reading) ac voltmeters by tradition indicate the rms value of a measured sinusoidal voltage. Because of variations in their operating principles, not all ac instruments measure the true rms value when the measured quantity has frequency components other than $1/T$. To measure true rms value, the instrument must indicate in proportion with the power absorbed by it from the voltage or current source. Of the common analog indicating instruments, the following are true rms indicating: the thermocouple type and the dynamometer type. In electronic voltmeters, including digital instruments, true rms measurements are accomplished either by thermal conversion of a current to an equivalent dc value or by analog evaluation or approximation of the root-mean-square or mean-square values of the input waveform. True rms measuring instruments are normally identified as such by their manufacturers. Because in electronic instruments the true rms detector is normally preceded by linear analog circuits, care must be taken not to exceed the maximum peak input and rate of change they can handle without distortion. The manufacturer's rating for input frequency and maximum crest factor shall be observed.

A.3 Flux voltage measurements

The instantaneous voltage induced in a winding is a direct function of the rate of change of the magnetic flux linking it (Faraday's law). The average value of this voltage over a period of time is a direct function of the net change of flux over the same period. Thus in the measurement of voltages induced across a winding, it is often desirable to measure the average value rather than the true rms value. By convention a flux volt is defined as 1.111 times the average of the absolute value of an alternating, undistorted, sinusoidal voltage of 1 V rms. The average of the absolute value is defined mathematically as

$$V_{avg} = \frac{1}{T} = \int_0^T |v(t)|\,dt \qquad \text{(A.2)}$$

Therefore if an instrument detects the average value of the measured quantity but is calibrated so that it displays 1.111 times that, it will indicate the rms value of the undistorted sinusoidal function that would have the same average of absolute value. These instruments are called “average responding, rms calibrated.” Most low-cost electronic-type instruments fall into this category since they detect ac signals by simple rectification and averaging (filtering). Of the analog meters the D’Arsonval type movement with rectified input is a common instrument of this type.

A.4 Applications

The need to use “flux” voltmeters arises from the need to know accurately the magnitude of flux changes in the magnetic cores. In an equation such as $E = 4.44\ NfAB_m$, the equation relating induced voltage to frequency, number of turns, and net flux change for periodic excitation, B_m, can be accurately computed if E has frequency components other than f, even though the form factor is applicable to sine wave only. The true rms values shall be measured wherever power, or the energy content of the measured quantity, is needed. The values of winding currents shall always be measured true rms, also the winding voltages for loaded tests. As an example, when the no-load exciting current of a transformer is to be measured, at a peak flux in the core, the voltage applied to the winding should be measured by a flux voltmeter and the current by a true rms instrument.

Annex B

(informative)

AC High-potential dielectric testing

B.1 General

AC high-potential dielectric testing (also known as hi-pot testing) is the application of an abnormally high alternating voltage between two (or more) isolated elements of an electrical device to test the integrity of the electrical insulation system. If there is no breakdown of the insulation system after applying a specified rms voltage level for a specified time, then the insulation system is considered to have satisfactory integrity. In order to define what is and what is not a breakdown (failure), the nature of leakage current, corona, and breakdown, in that order, must be discussed.

B.2 Leakage current

The total leakage current I_T at any specified rms voltage and frequency is a composite current made up of insulation resistance current I_R and capacitive current I_C, which is in quadrature. This is related by the equation

$$I_T = \sqrt{(I_R)^2 + (I_C)^2}$$

Almost always the largest component of leakage current by far is the capacitive current I_C. The amount of I_C can vary considerably from one type of device to another. Some larger transformers with one winding surrounded by other windings or electrostatic shields can easily have a few milliamperes of I_C at 2500 V rms, 60 Hz (a commonly specified voltage and frequency). Some test equipment meters the leakage current I_T while others “sense” the current through a resistor using the resultant voltage drop on the resistor to “trip” a relay causing a failure light or a horn to sound. This type of equipment can indicate a “failure” when, in fact, there is no breakdown but a moderately high leakage current due to I_C. Generally speaking, leakage current I_T is, within some reasonable limits, not a cause for rejection of a device. When ac high-potential testing is done immediately after a “burn-out” test (operating a transformer at rated input with secondary windings short-circuited for 6 h or until it burns out opening the primary), the insulation system will be very hot. At these temperatures the dielectric constant usually increases substantially, which increases I_C proportionately. Also at these temperatures the insulation resistance usually drops a few orders of magnitude, causing I_R to increase proportionately. Again, this type of leakage current is acceptable and the device would pass the test providing that there is no breakdown of the insulation system.

B.3 Corona

Corona is a discharge due to ionization of air in voids, microvoids, or space around terminations. Corona itself is bad because it causes an accelerated aging of the insulation material leading to failure and breakdown. The aging effects of corona are accumulative. Corona inception occurs at certain voltage levels in all insulation materials. While that voltage level is different for different materials and for different thicknesses of the same material, it can generally be said that most insulation materials have corona inception at approximately 100 V/mil (mil = 0.001 in) or approximately 3900 V/mm. When two or more different materials are used in an insulation system between two electrodes, the voltage dividing effect of the different dielectric constants must be considered when determining the volts per mil (or per millimeter) on each insulation. The material with the lowest dielectric constant will inherently have the highest volt per mil (or per millimeter) stress. Care must be taken not to allow corona at normal or maximum operating voltage levels. However, under high-potential testing some corona may be allowed and generally exists. If it occurs in high levels at the high-potential voltage, it may cause an arc-through in the insulation and result in an insulation breakdown. If no breakdown occurs after 1 min of high-potential test, then it is acceptable.

B.4 Breakdown

Breakdown of the insulation system may take different forms. It may occur during the high-potential test, or it may have occurred prior to the test due to damaged insulation material or an assembly error that caused electrical connections to touch or come in close proximity with each other allowing the high voltage to arc over on application. When breakdown occurs during the high-potential test, it can be detected by sporadic and intermittent increases in leakage current I_T at first, followed by a great increase to some constant high level, which on many high-potential test sets is accompanied by a collapse of the voltage to some lower level or completely. This type of breakdown may be caused by partial breakdown in ionized voids and weakened insulation, which then avalanches to a complete breakdown in the form of an arc-over or an arc-through.

B.5 Test equipment requirement

The range of types of transformers and other electrical devices that require high-potential testing up to approximately 2500 V rms, 50 or 60 Hz, are such that the capacitive current I_C will generally not be over 5 mA rms. A high-potential test set with approximately 15 mA rms current limiting at 2500 V should be quite adequate. However, limiting the current to 5 mA rms or less may be restrictive in the test of electrical devices with high interelectrode capacitance. Measurement of interelectrode capacitance will allow one to determine the value of I_C expected during any given high-potential test.

Annex C

(informative)

An ac magnetic field pickup probe

A small random-wound multilayer solenoid coil on a standard bobbin, such as used in commercial 120 V solenoid structures of medium size, can be used for making up the coil with as many turns as can be applied, using a #36 AWG enamel wire. Typically one has been wound with about 4100 turns. The leads from this coil are connected to a twisted pair of copper shielded cable. The coil is positioned within a brass or copper can with flanged cover, large enough to accommodate the bobbin. The wound coil is centrally located within the can with the axis of the bobbin being coincident with that of the can. To eliminate the shunt effect of the can, it should be slit on one side and halfway radially across the closed end. The same thing should be done to the cover, which, of course, should also have a clearance hole in its center to permit egress of the cable. The copper-braided outer sheath of the cable should then be soldered to the side wall of the can or the cover. To permanently position the bobbin and secure the cable, the can should be filled with a potting compound, preferably the epoxy type. During the operation, the slotted area of the can must be sealed, using a pressure-sensitive tape sufficiently strong to retain the epoxy until it sets up. After the casting resin has properly cured, the cover can be positioned on the can, being certain to align its slot with that of the can. The cover should then be soldered into position. The other end of the cable should then be attached to a shield plug. This plug is made up with two banana-type connectors on 3/4 in centers, which will fit on most voltmeter input terminals. The twisted wires should be connected to the terminals as well as a shunt resistance of about 100 000 Ω. The outer copper braid is terminated on the ground side of the plug. It should be noted that by having the braided shield connected only to the shield can and to the ground plug at the voltmeter input, the possibility of a loop pickup effect in the cable is minimized. Prior to final assembly of the plug, the aluminum housing can be lined with a single layer of magnetic shielding foil. This will ensure maximum electromagnetic shielding of the exposed areas of the plug. This completes the pickup coil as far as the physical assembly is concerned. One more operation is necessary to complete the probe calibration. The effective geometric center of the coil shall be determined and marked on the outer shield case. The only remaining problem is that of calibration, that is, determining millivolts of output per gauss. To accomplish this, a known ac field source shall be generated.