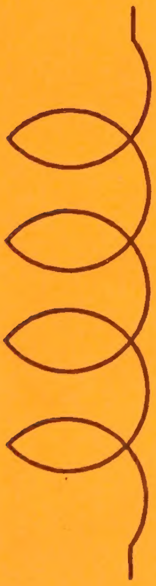
THE INDUCTOR HANDBOOK



A COMPREHENSIVE GUIDE FOR CORRECT COMPONENT SELECTION IN ALL CIRCUIT APPLICATIONS.
KNOW WHAT TO USE WHEN AND WHERE.

Cletus J. Kaiser

THE
INDUCTOR HANDBOOK

First Edition

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Published
by

CJ Publishing
2851 W. 127th Street
Olathe, KS 66061

FIRST PRINTING -- 1996
SECOND PRINTING -- 1997

ISBN: 0-9628525-4-6
Library of Congress Catalog Card Number: 96-096325

Printed in the United States of America.

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Acknowledgements

The author is thankful to The Lord.

The author is deeply indebted to his family for their support.

Appreciation is also expressed to my friends.

Preface

With the completion of "The Capacitor Handbook" and "The Resistor Handbook," the only passive component left to cover is the inductor. Hence, "The Inductor Handbook" has been published. This book provides guidance for the use of inductors and transformers in electronic and electrical circuits.

As with all of my books, the chapters are arranged with theory and construction discussed first, followed by application information. With all chapters arranged in this manner, reading and using this book for reference will be easier.

The first chapter covers the fundamentals of all inductors. The other chapters discuss the different types of inductors and transformers with their practical applications. A glossary of terms, symbols, equations and a comprehensive index is included.

Chapter 1

Fundamentals For All Inductors

Of all passive components, inductors are probably the most difficult to specify because of the many parameters which must be considered. The purpose of this chapter is to examine some of these parameters, their measurement, and their application in practical circuits.

These symbols are used in the following equations:

- C = capacitance in Farads
- di/dt = rate of current change in amps per second
- e = induced electromotive force in volts
- f = frequency in Hertz
- f_o = center or resonant frequency
- L = inductance in Henries
    - 1 microhenry (1 µH) = 10^{-6} Henry
    - 1 millihenry (1 mH) = 10^{-3} Henry
- R_e = effective resistance in Ohms
- X_C = capacitive reactance in Ohms
- X_L = inductive reactance in Ohms
- Z = impedance in Ohms

INDUCTANCE

The basic principles of electromagnetism are well known: a current passed through a coil of wire will produce a magnetic field; a changing magnetic field induces an emf which opposes the field-producing current. Thus the coil tends to impede changes of current, a property called inductance.

Inductance is usually defined in terms of the voltage produced by a variation in current flow through an inductor, as expressed in equation 1:

Equation 1. $e = L\,di/dt$

For practical applications, it is more useful to define inductance in terms of its relationship to impedance, frequency, or energy, as follows:

The impedance of an ideal inductor (i.e., one without losses) to a sinusoidal AC signal is equal to its inductive reactance:

Equation 2. $Z = X_L = j2\pi fL$

The "j" indicates that the voltage and current are 90° out of phase, with the current LAGGING the voltage. The impedance of an ideal capacitor is given by the equation:

Equation 3. $Z = X_C = \dfrac{-j}{2\pi fC}$

the "-j" indicates that the current LEADS the voltage by 90°.

If a capacitor and inductor are connected in series, there will be some frequency at which the magnitudes of the two impedances are equal. This is the resonant frequency, which, from (2) and (3) is given by:

Equation 4. $f = \dfrac{1}{2\pi\sqrt{LC}}$

At this resonant frequency, the voltages across the two components will be equal in magnitude and 180° out of phase, cancelling each other. Thus, for ideal components, the series combination will have an impedance of zero.

Similarly, it can be shown that for a parallel L-C combination at resonance, the currents will cancel each other and the combination will have infinite impedance.

In practical circuits there will always be some losses so that the theoretical values of zero and infinity cannot be obtained, but for most cases the following rules apply at the resonant frequency given by equation 4.

- For SERIES resonant L-C circuits the impedance will be very low (inversely proportional to Q) and will appear to be purely resistive.
- For PARALLEL resonant L-C circuits the impedance will be very high (directly proportional to Q) and will appear to be purely resistive.

The capacitor losses will usually be negligible compared to the inductor losses, therefore the circuit impedance will be mainly determined by the Q of the inductor.

The energy that is stored in an inductor carrying a current (I), is given by equation 5:

Equation 5. Energy (joules or watt-seconds) = $1/2\ LI^2$

Q (Quality factor)

Any practical inductor will have some direct current resistance (DCR), resulting in I^2R power losses. When alternating current is passed through an inductor, there will be additional losses from skin effect resistance, hysteresis, and eddy currents. At any given frequency, the sum of these losses can be represented by the power dissipated in a single equivalent resistance, called the effective resistance.

Q, or quality factor, is a measure of the relative losses in an inductor, and is defined in equation 6 as the ratio of inductive reactance to effective resistance:

Equation 6. $Q = \frac{X_L}{R_e} = \frac{2\pi fL}{R_e}$

Note that both X_L and R_e are functions of frequency, therefore the test frequency must be given when specifying Q. At low frequencies, X_L will usually increase with frequency at a faster rate than R_e, and at high frequencies the reverse is true, resulting in a bell-shaped Q vs. f curve. The Q curve will sometimes have a double peak, because of the dominance of skin-effect at certain frequencies.

For resonant L-C circuits, the Q will determine the selectivity, as expressed in equation 7.

Equation 7. $$\text{Bandwith (3db)} = \frac{f_o}{Q}$$

DISTRIBUTED CAPACITANCE AND SRF

In any practical inductor, each turn of wire acts as a capacitor plate, and the combined effect of turn-to-turn and (in the case of iron-core inductors) winding-to-core capacitance can be represented as a single capacitance, called the distributed capacitance (C_d), in parallel with the inductor. This parallel combination will resonate at some frequency, given by equation 4, which is called the self-resonant frequency (SRF). At this frequency the inductor will act as a pure resistor with high impedance, and at frequencies above the SRF the capacitive reactance of the parallel combination will become dominant.

SATURATION AND CURRENT RATING

The core materials used for inductors are of two basic types: "air-core" - meaning any nonmagnetic material including air, plastic, or ceramic; and "iron-core" - meaning any magnetic material including ferrite, powdered iron, and nickel or steel alloys.

As the current through an iron-core inductor is increased, the flux density within the core will increase until saturation is reached, at which point the inductance will begin to decrease. This is a nonlinear function, and once a core begins to saturate, the inductance will drop off rapidly with further increases in current.

In some cases, such as the measurement of shielded ferrite chokes, the voltage applied by the test instrument may be sufficient to saturate the core. For this reason, the method of measurement and test voltage are often specified, in order to ensure correlation of measured values.

A core which has been driven to saturation will not immediately recover, and will show a temporary change in permeability. The effect is usually negligible, but when an inductor is measured soon after being subjected to magnetic shock such as dc continuity testing, the inductance may appear to be out of tolerance.

There are three factors to consider when determining the current rating of an inductor:

- **Direct Current Rating** (Rated IDC) is simply calculated from the DC resistance and power dissipation for a given temperature rise.
- **Alternating Current Rating** is also based on power dissipation, but the effective resistance ($R_e = X_L/Q$), a function of frequency, must be used in calculating I^2R losses.
- **Incremental Current Rating** (INCR I) is the current at which the inductance in decreased by 5% because of saturation.

It is standard practice to specify both Rated IDC and INCR I for iron-core inductors. Depending on the core and winding configuration, the Incremental Current Rating may be considerably higher or lower than the Direct Current Rating.

An inductor with a high Incremental Current Rating would be useful in a circuit with high peak currents where the inductance must be maintained, while one with a higher Direct Current Rating might find application where the DC voltage drop is a primary consideration.

MEASUREMENT

THE Q-METER

The standard instrument for measurement of small RF inductors is the Q-meter; - which consists of a variable frequency signal generator, a calibrated variable capacitor, and a high impedance RF voltmeter.

The inductor to be measured is connected in series with the capacitor, and the capacitor and/or signal generator are adjusted for resonance. The inductance can then be calculated using equation (4). For convenience, the capacitor dial is calibrated for direct inductance readings at certain frequencies as follows:

0.1 to 1.0 µH — 25 MHz
1.0 to 10 µH — 7.9MHz
10 to 100 µH — 2.5 MHz
0.1 to 1.0 mH — 790 KHz
1.0 to 10 mH — 250 KHz
10 to 100 mH — 79 KHz

The Q is proportional to the voltage across the capacitor, which is measured by the RF voltmeter.

SOURCES OF ERROR

Residual Inductance - The internal inductance of the Q-meter, as well as the inductance of any test clips or fixtures can introduce a significant error when measuring small inductances. This residual inductance can be measured by substituting a shorting bar of known inductance for the inductor being measured. The residual inductance thus determined is subtracted from the measured inductance to give the EFFECTIVE INDUCTANCE.

The inductance of the coil leads can also affect the accuracy of the reading. To ensure repeatability and correlation of measurements, the effective lead length from coil body to point of connection should be specified for inductance under approximately 10µH.

Distributed Capacitance - Because the Q-meter method of inductance measurement depends on accurate measurement of resonating capacitance, the distributed capacitance (C_d) of an inductor is another possible source of error, especially when a low value of resonating capacitance is used to measure large inductors.

In addition, the proximity of any large metal object, such as the instrument chassis, causes capacitance to ground which may result in inaccurate measurements.

CORRECTIONS

The following terms apply to Q-meter measurement of inductance:

- MEASURED INDUCTANCE is the inductance as read directly or calculated from the capacitance dial, without correction.
- EFFECTIVE INDUCTANCE is a measured inductance which has been corrected for residual inductance.
- TRUE INDUCTANCE is an effective inductance which has been corrected for distributed capacitance effects.

It is standard practice to specify the effective inductance of RF inductors, thus the true inductance can be considerably lower than the indicated value. For this reason, inductors with relatively high values of C_d are often measured on an impedance bridge at 1KHz where the effects of C_d become negligible.

Testing Inductors at Application Frequencies

The accurate measurement of an inductor has always been more difficult than the measurement of other passive components. The primary difficulty with coil measurements lies in the fact that coil inductance and Q are frequency dependent; similarly, coil parasitics will vary dramatically with frequency. The measurement of coils at application frequencies, so-called "use frequency testing," is more accurate than testing at typical traditional frequencies.

Often, the value of a measurement frequency is specified for measurement convenience alone. If the measurement frequency is not the circuit (or "use") frequency, the result of testing will generally not yield the same inductance value seen by the intended circuit. Given that recent developments of equipment and methods now allow more flexibility in test frequency selection, particularly if tight tolerances are required, inductors should be tested at the actual frequency of use.

Inductor Parameters

The primary electrical parameters of a coil are inductance, Q, self resonant frequency (SRF), and the direct current resistance (DCR). All primary electrical parameters are design controlled, although not independently. The inductance and Q are highly dependent on both the frequency and instrumentation of a test.

Inductance is expressed in Henries (and sometimes as its capacitance corollary in Farads) with a tolerance, at a specific frequency. The Q parameter is unitless as a figure of merit which is specified as a minimum. The SRF is stated in Hertz as a minimum and the DCR is expressed in Ohms as a maximum.

Other parameters such as current carrying capability and impedance are sometimes specified but are not necessary; they are reflected in the primary parameters. For example, the wire gauge determines both the DCR and current carrying ability. The DCR is easier to measure than the current density (or the temperature rise) and both parameters are indicated in the Q.

Test Equipment

The appropriate selection of a method of test largely determines the accuracy of a measurement. Instrumentation and test methods vary for individual electrical parameters, and every instrument has further limitations in terms of range, frequency, and error. Parasitics and their effects associated with test fixturing is a significant consideration in a measurement. In all cases, the instrument, fixturing, frequency, and current (if applicable) must be specified in order to have a repeatable and reliable test.

A Q-meter or impedance analyzer is generally used for inductance and the Q measurements. There have also been some recent efforts in measuring Q on a network analyzer. The SRF can be measured on a grid dip meter or a network analyzer. The DCR is usually measured on a low Ohm meter or a Wheatstone bridge.

The choice of equipment may establish the measurement units for inductance. For example, when coils are measured on a Q-meter, the affected units for inductance are picoFarads (pF). A Q-meter resonates the coil under test with a variable capacitor; the meter indicates this value of capacitance. In most cases, the measurements are not converted to Henries, but left as picoFarads.

The selection of instrumentation also influences measured values. The influence an instrument has is a result of the various measurement methods and frequencies employed by each piece of equipment. In the case of inductance, the following table shows the typical variations that can be expected, for the same coil, when measured on different instruments, at different frequencies.

INSTRUMENT	FREQUENCY	INDUCTANCE
HP4342A	25 MHz	682.3 nH (59.4 pF)
MEGURO MQ-171	25 MHz	580.6 nH (69.8 pF)
MEGURO MQ-171	50 MHz	603.1 nH (16.8 pF)
HP4192A	0.130 MHz	607.0 nH
HP4192A	10 MHz	592.7 nH
BOONTON 62AD	1 MHz	594.0 nH
TEKTRONIX LC130	0.130 MHz	1300.0 nH
HP4191A	100 MHz	1065.0 nH

As we can see from the previous table, an inductance value can vary considerably, depending upon how the inductor has been measured. Part of the discrepancy can be attributed to the different instruments, but the majority is due to the different frequencies. In general, the proper instrument to specify is one that is accurate and repeatable at the frequency required.

Traditional Method of Specifying and Testing Inductors

Q-meter measurements have been the conventional method of specifying L and Q parameter values. However, a Q-meter requires a test frequency that is both within the range of its oscillator and within the limits of its tuning capacitor. Also, a Q-meter measurement yields an inductance that is in terms of capacitance (pF). Q-meters also have specific frequencies (called blue lines) that are easier to use than intermediate frequencies because an alternate scale on the tuning capacitor allows direct inductance readings.

Commercially available Q-meters have accuracies of no better than 3% and typically much worse than this. Q-meter inaccuracy requires the use of setup "standards" (correlation pieces). Correlation pieces are deemed to be the standard for a particular component part and are used to set up each instrument every time a test is performed. Correlation is still the traditional method and results in very little error, excellent repeatability, and is applicable at any frequency. However, the method of correlation has significant logistical disadvantages: the establishment and accountability of specific correlation pieces between the manufacturer and the customer, as well as the task of adjusting every instrument for each test.

The other electrical parameters, SRF and DCR are generally specified along with L and Q but there is rarely any reference to the test method used. The lack of a specific method of testing SRF and DCR reflects the fact that the inductance is the dominant parameter and one which requires the most diligent control.

Application Frequency Testing

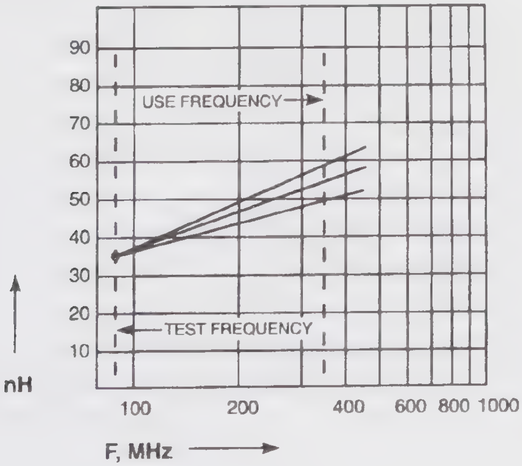
The basic complication with the traditional testing method is that the coils are tested at one frequency and used at another frequency. Graphs shown in figures 1 and 2 show the fundamental problem with the traditional testing method.

Inductance vs. Frequency

Figure 1 shows a log sweep of inductance vs. frequency for three different inductors. When these parts are tested at a typical Q-meter frequency, they appear to be identical in terms of inductance. At the actual circuit frequency, these coils are quite different. These three coils can represent three different designs or three different coils of the same design. If the use frequency is the point where all three converge, then these coils are effectively the same. If the circuit frequency is significantly different from the test frequency, then the inductance at the use frequency cannot generally be implied from the test frequency.

Even if these coils had not deviated from each other, there is still a change in inductance. The consequence of an inductance shift with frequency is generally observed when a design requires a specific inductance and the particular coil does not perform as expected. The shift is generally attributed to fixturing differences and circuit parasitics and the specified value is compensated accordingly.

FIGURE 1. Inductance vs. Frequency.

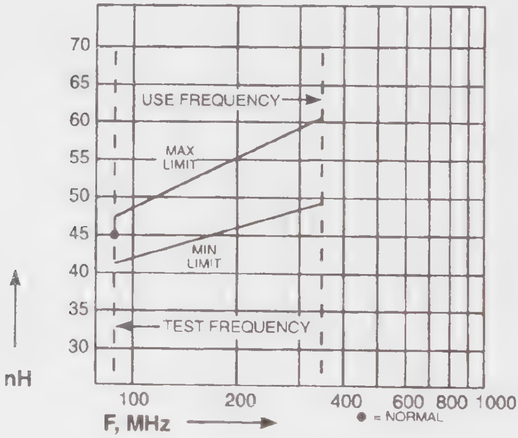


Inductance vs. Frequency

Figure 2 shows another consequence of testing at a frequency other than the application frequency. It displays an inductance vs. frequency sweep for a particular coil. The normal inductance is indicated at both frequencies along with their tolerances. If this coil had been specified as a 5% inductor at the test frequency, it would have resulted in a 10% tolerance at the application frequency. These limits could have compressed over frequency or spread even farther, depending on the design. The fact that the limits are not constant with respect to each other results in a loss of tolerance control over frequency.

In all cases where the inductance changes with frequency, testing at the application frequency results in better specification control. Although correlation is still the most expedient method of reducing testing error, it is almost always applied at some frequency other than the application frequency. Testing at the application frequency, without correlation, can be an extremely effective and superior method of determining the application suitability of a coil.

FIGURE 2. Inductance vs. Frequency.



Specifying Inductor Tests at Application Frequencies

The procedure to specify and the electrical test for an inductor at use frequencies is as follows:

1) Inductance

1.1) Specify the nominal inductance.

1.2) Specify the test instrument, fixture and frequency.

1.3) Specify the inductance tolerances.

- 1.3.1) Determine the allowable tolerance as a % using 6 sigma or other appropriate methods. The customer should test to this tolerance without correlation.
- 1.3.2) Evaluate the instrument error for the nominal impedance at the test frequency. Subtract this error % from the allowable tolerance found in step 1.3.1.
- 1.3.3) Establish the instrument and fixture repeatability. Subtract this error % from the result of step 1.3.2. This result is the tolerance to specify. The manufacturer should test to the specified tolerance without correlation as all errors have been accounted for.

2) Q

2.1) Specify the absolute minimum Q (allowable minimum).

- 2.1.1) Determine the allowable minimum using 6 sigma or other appropriate methods. The customer should test to this tolerance.

2.2) Specify the test instrument, fixture and frequency.

2.3) Specify the Q minimum for manufacture.

- 2.3.1) Evaluate the instrument error for the nominal impedance at the test frequency.
  Adjust the allowable minimum found in step 2.1.1 (i.e., increase the minimum Q by an amount equal to the instrument error).
- 2.3.2) Establish the instrument and fixture repeatability.
  Adjust the new allowable minimum found in step 2.3.1 (i.e., increase the minimum Q by an amount equal to the test repeatability). This result is the tolerance to specify. The manufacturer should test to the specified tolerance without correlation because all errors have been incorporated into the final adjustment of the Q specification.

3) DCR

3.1) Specify the absolute maximum DCR (allowable maximum).

- 3.1.1) Determine the allowable maximum using 6 sigma or other appropriate methods. The customer should test to this tolerance without correlation.

3.2) Specify the test instrument and fixture.

3.3) Specify the DCR maximum for manufacture.

- 3.3.1) Evaluate the instrument error for the nominal resistance. Adjust the allowable maximum found in step 3.1.1 (i.e., decrease the maximum DCR by an amount equal to the instrument error).
- 3.3.2) Establish the instrument and fixture repeatability. Adjust the new allowable maximum found in step 3.3.1 (i.e., decrease the maximum DCR by an amount equal to the test repeatability). This result is the tolerance to specify. The manufacturer should test to the specified tolerance without correlation because all errors have been incorporated into the final adjustment of the DCR specification.

4) SRF

4.1) Specify the absolute minimum SRF (allowable minimum).

- 4.1.1) See SPECIFICATION EXAMPLE below.

Component prints should indicate that inductance is the critical parameter. Assuming the need exists for a sorted population, only inductance needs to be 100% tested. The other parameters will track inductance and should only be qualified.

Specification Example

As an example, we will specify the aforementioned parameters for a chip inductor. The coil in question was measured at 300 MHz, the use frequency of the circuit for which the component was designed.

INDUCTANCE

The coil was measured to have an average inductance of 53.8 nH with a standard deviation of 1.1 nH. We will use the average as the nominal.

A six sigma specification for the indicated coil would be:
53.8 nH ± (6 x 1.1 nH), or ± 12.3% of a nominal 53.8 nH. We have then satisfied steps 1.1 and 1.3.1, the specification of the nominal inductance and its tolerance.

We will choose the tester and a test fixture. We will specify the component to be tested at the 300 MHz use frequency. We have now satisfied step 1.2, the identification of test instrumentation and frequency.

The basic measurement error associated with the tester, testing at 300 MHz for a 53.8 nH inductance is approximately 0.5%. We subtract the measurement error from 6 sigma tolerance of 12.3% and obtain nominal ±11.8% tolerance to compensate for instrumentation accuracy. We have then satisfied step 1.3.2.

A testing setup may have an overall session to session repeatability of 1.0%. We adjust the resultant tolerance of 11.8% given above by an additional repeatability error of 1.0%. We have the final manufacturing tolerance (print specification) of 10.8%, thus satisfying step 1.3.3.

Q

The Q parameter tracks the inductance parameter very closely, that is given the same designed coil the Q will not deviate without a noticeable change in L. For example, a shorted turn will affect both the Q and L parameters simultaneously. Because of the dependency of Q on L, reliable Q measurements can be made on Q-meters which cannot support high application frequencies and which are most accurate at lower frequencies, although application frequency Q measurements are ideal whenever possible.

The average Q of the coil was measured to be 42.2 on a Q-meter at 150 MHz with a standard deviation of 0.8. Using the average again as the nominal we obtain a six sigma minimum of 42.2 - (6 x 0.8), or an allowable minimum Q of 37.4 (nominal - 11.4%). Step 2.1.1 is completed.

We will specify a Q-meter, with a test fixture, and a 150 MHz test frequency (Step 2.2). The instrument error of the tester at 150 MHz is 7.5%. We adjust the minimum Q specification to compensate for instrument error and obtain a new minimum Q of 40.6 (nominal - 3.9%). Step 2.3.1 is completed.

The session to session repeatability of the test fixture is 2.0%. We make our final adjustment to the Q minimum specification by subtracting the repeatability error from the last adjustment. The adjustment is 3.9% - 2.0% = 1.9%, which brings the manufacturing Q minimum (print specification) to 41.4 (Step 2.3.2).

DCR

The average DCR of the coil was measured to be 1.97 Ohms with a standard deviation of 0.05 Ohms. Using the average as the nominal we obtain a six sigma maximum of 1.97 + (6 x 0.05), or an allowable maximum DCR of 2.27 Ohms (nominal + 15.2%). Step 3.1.1 is completed.

We will specify the meter model and the chip fixture to perform the DCR tests. (Step 3.2).

The instrument error of the meter is 0.25% for a 2.0 Ohm measurement. We adjust the maximum DCR specification to compensate for instrument error and obtain a new maximum DCR of 2.26 (nominal + 14.95%). Step 3.3.1 is completed.

The session to session repeatability of the fixture is 0.5%. We make our final adjustment to the DCR maximum specification by subtracting the repeatability error from the last adjustment.
The adjustment is 14.95% - 0.5% = 14.45%, which brings the maximum specification (print specification) to 2.25 Ohms. Step 3.3.2 is completed.

SRF

The SRF is not an independent parameter, given that the other parameters of L, Q. and DCR are specified. The SRF should generally be specified as at least twice the application frequency. The manufacturer can indicate the expected SRF of particular coils and coil constructions.

Specifications With Correlation

Each of the primary coil parameters in the above example are specified without the use of correlation. In this technique, each of the individual measurement errors are reflected in the specifications. The use of correlation in conjunction with use frequency testing can eliminate these errors and allow the specifications to be tightened. Correlation can be used with any of the inductor parameters but is most often applied to inductance and Q.

Conclusion

Coil inductance and Q are frequency dependent and testing methods have further influence on these parameters. Specifying and testing at the factual circuit frequency is an appropriate method of controlling inductor parameters. Use frequency testing assures components consistent with their intended application.

The primary inductor parameters are interrelated design functions of the coil and the print should reflect that the inductance is the dominant test parameter. The inductor specification should also take into account the component variations and instrumentation errors.

Application Information

Chip Inductors have been designed for the need of today's high frequency designer. Their ceramic construction delivers high SRF's as well as excellent Q values. A nonmagnetic coilform also assures the utmost in thermal stability, predictability and batch consistency. Molded construction provides superior strength and moisture resistance.

Air-core Inductors provide high Q over a wide range of frequencies. They feature tight inductance tolerance and thermal stability which can often eliminate the need for circuit tuning.

Fixed RF Inductors are molded inductors. This precision construction technique assures constant winding pitch, long term stability, and tight tolerance on inductance.

Variable Inductors (Tuneable RF Coils) are molded in a plastic housing for the compactness and the low drift reliability of an insert molded coil. The windings are precision molded into a single piece for mechanical and electrical stability. Optional plated brass shield cans with solderable tabs provide integral shielding and additional mounting stability. Tuning is done by means of a threaded core with an easy adjustment.

Power Inductors are effective as DC-DC converter boost or buck inductors and as output ripple filter chokes in all types of downsized switching power supplies. The large inductance makes these parts ideal for all sorts of energy storage, smoothing, and EMI reduction applications.

Surface Mount Power Inductors are designed for the higher current requirements of portable, handheld devices.

DC-DC Converter Inductors are ideal for use in step-up or step-down converters, flash memory programmers, etc.

Conformal Coated Axial Lead Chokes are totally encapsulated in a durable epoxy coating. Their low cost compared to molded type chokes makes them attractive to high volume users. Availability is based on the usage of standard values.

Molded Axial Lead Chokes have an epoxy molded construction which provides superior moisture protection and sturdy construction.

Axial Lead Power Chokes are for use as output chokes in low power, downsized switching supplies, DC/DC converters and inverters, EMI/RFI noise suppression, and smoothing choke applications.

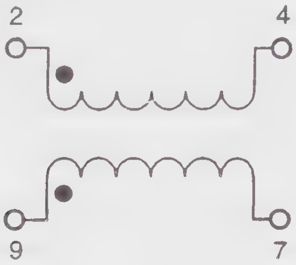
Shielded Chokes are intended for uses in low-pass filtering applications where high stability is not required. These chokes incorporate a flanged ferrite bobbin with a close-fitting ferrite sleeve to provide magnetic shielding, resulting in high inductance with low DC resistance. Because the gap between the sleeve and the bobbin is small, core saturation occurs at relatively low current levels, thus the incremental current is much lower than the current rating. Also, because the gap can vary with temperature changes and aging, these parts are subject to inductance drift over time and may exceed tolerance limits by up to 5% after leaving the factory. Available tolerances are ±10% and ±25%, subject to the above conditions, and GMV (inductance guaranteed to be not less than the nominal value).

High Current Filter Inductors are used in noise filtering for switching regulators, power amplifiers, power supplies and SCR or Triac control circuits.

Low Pass LC Filter Modules applications include analog signals, video and communication signals, 10BaseT and other LANs, FCC and VDE EMI filtering. Space saving SIP construction of these filters offer all the benefits of an LC design, low insertion loss, high power handling and controlled group delay.

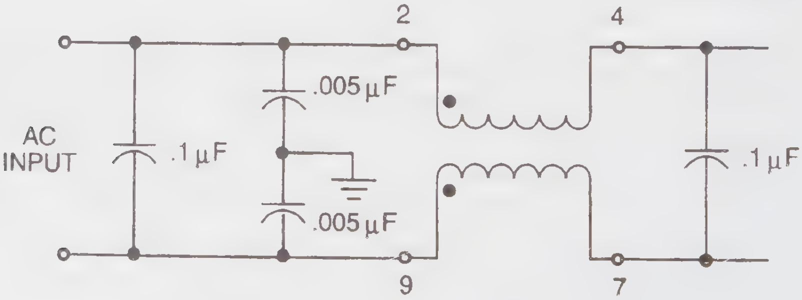
Common Mode Line Chokes are used greatly to reduce AC line conducted interference produced by switching power supplies. This configuration produces opposing magnetic fluxes in the core that serve to cancel inphase noise signals appearing across the AC line. This mode allows much more filtering capability in a given core size that would be possible using differential filtering alone. It is important to note that only a common mode signal will be attenuated. Differential mode power such as 50/60 Hz power line currents or data signals will pass through unimpeded by the common mode impedance. Use of common mode chokes therefore reduces component count as well as inductor size.

FIGURE 3. Common Mode Choke Schematic.



Combination Line Filter Choke Assembly is constructed by combining both common and differential mode filtering in a single component.. These chokes are intended for use in AC line filters for switching power supplies. By using these parts, designers can eliminate two extra filter chokes compared to a standard common mode choke. The unique combination of windings and materials provides much better filtering performance than either a common mode choke or a single winding inductor alone. The differential filtering frequency response has been designed to provide filtering at higher frequencies while still allowing the AC line power to pass through without loss. This gives the power supply designer the added flexibility require to adjust the filtering to meet FCC, VDE, and other requirements as well as optimize the circuit for the particular noise levels and frequencies produced by each power supply design.

FIGURE 4. Common Mode Choke Typical Application.



Distributed Capacitance

Every winding has distributed capacitance resulting from the differential voltage between adjacent turns, the type of dielectric insulation on the core and wire, and the moisture present in the dielectric. This distributed capacitance is equivalent to a single capacitor shunted across the coil inductance. This parallel circuit resonates at the coil self-resonant frequency, f_o.

$$f_o = \frac{1}{2\pi\sqrt{LC_d}}$$

f_o = frequency (Hertz)
L = coil inductance (Henries)
C_d = distributed capacitance (Farads).

Minimizing this distributed capacitance is an important core winding goal. An operating frequency too close to the self-resonant frequency of the coil, which can occur with high inductances and/or high operating frequencies, reduces the effective circuit Q and increases the apparent inductance. If the operating frequency is 10% of f_o, the actual Q is 99% of the calculated Q. If the operating frequency is 50% of f_o, the actual Q is only 75%, of the predicted value.

Distributed capacitance may be minimized by proper choice of winding technique. Whichever technique is chosen, the first and last winding turns must not be adjacent. As the lowest capacitance windings are also the highest cost windings, a cost/quality decision must be made.

Random wound cores utilizing 360°of the core yield the highest capacitances; 50 pF to 75 pF are typical values at 1000 turns. Random wound cores utilizing 330°, sector-wound cores, and bank-wound cores all reduce the distributed capacitance.

Progressive winding (modified bank) with a 1/8" gap between start and finish offers economy with relatively low distributed capacitance. On medium size cores, progressive winding typically yields less than 10 pF at 100 turns, 12 to 18 pF at 500 turns, and only 25 to 35 pF at 1000 turns.

At higher frequencies, litz wire may be desirable and should be considered; however, stranded wire significantly reduces the winding factor and thus increases the DC winding resistance. Potting compounds, either waxes or epoxies, can also increase the distributed capacitance 30% to 70%. Moisture content must be kept at an absolute minimum if low distributed capacitance is essential.

Chapter 2

Ferrites

The electronics industry as we know it today, could not exist without the widespread use of ferrites. The term "ferrite" is derived from the Latin word "ferrum," meaning iron. Ferrites are homogeneous ceramic materials composed of various oxides containing iron oxide as their main constituent. Being ceramic, ferrites are hard, inert, and free of organic substances.

What makes ferrites especially useful in the electronics industry is a combination of two key characteristics:

- high magnetic permeability which concentrates and reinforces the magnetic field and
- high electrical resistivity which limits the amount of electric current flow in the ferrite.

With these two characteristics, ferrites exhibit low energy losses, are highly efficient, and function at high frequencies (1 KHz to 1,000 MHz). These qualities make ferrites the ideal building blocks in the manufacture of miniaturized high frequency electronic components.

Ferrites are manufactured in several different shapes; (toroids, beads, multi-hole cores, etc.) and in different sizes. They can also be finished, wound, assembled, and packaged according to specifications.

History

The history of magnetism began with the discovery of the properties of a mineral called magnetite (Fe_3O_4). The most plentiful deposits were found in the district of Magnesia in Asia Minor (hence the mineral's name) where it was observed, centuries before the birth of Christ, that these naturally occurring stones would attract iron. Later on it found application in the lodestone of the early navigators. Then in 1600 William Gilbert published *De Magnete*, the first scientific study on magnetism. In 1819 Hans Christian Oersted observed that an electric current in a wire affected a magnetic compass needle, thus with later contributions by Faraday, Maxwell, Hertz and others, the new science of electromagnetism came into being.

Even though the existence of naturally occurring magnetite, a weak type of hard ferrite, had been known since antiquity, producing an analogous soft magnetic material in the laboratory proved elusive. Research on magnetic oxides was going on concurrently during the 1930's, primarily in Japan and the Netherlands. However, it was not until 1945 that J. L. Snoek of the Philips' Research Laboratories in the Netherlands succeeded in producing a soft ferrite material for commercial applications.

Not far behind was the manufacture and sale of soft ferrites for use in the electronics industry. The ensuing years have seen a rather crude product, which was available in only a few shapes and materials, develop into a major line of ferrite components for inductive devices, produced in many core configurations with a wide selection of materials. The application of ferrites in EMI suppression as shield beads and broadband chokes, where an effective resistive impedance is produced at high frequencies, has grown so fast in the last decade, that their use as EMI suppressors is limited only by the imagination of the end user.

Soft Ferrites

The single most important characteristic of soft ferrites, as compared to other magnetic materials, is the high volume resistivity exhibited in the monolithic form. Since eddy current losses are inversely proportional to resistivity and these losses increase with the square of the frequency, high resistivity becomes an essential factor in magnetic materials intended for high frequency operation. The magnetic properties of ferrite components are isotropic, and by employing various pressing, extruding, and/or grinding techniques, a wide range of complex shapes can be formed. There is no other class of magnetic material that can match soft ferrites in performance, cost and volumetric efficiency, over the range from audio frequencies to above 200 MHz.

During the last 25 years the basic constituents of ferrites have changed little, but purity of raw materials and process control have improved dramatically. Ferrites are ceramic materials with the general chemical formula MO Fe_2O_3, where MO is a divalent metal oxide blended with 48 to 60 mole percent of iron oxide.

Three broad groups of soft ferrite materials are:

- **Manganese zinc** - Manganese zinc ferrites are completely vitrified and have very low porosity. They have the highest permeabilities and can exhibit volume resistivities ranging from one hundred to several thousand ohm-centimeter. Manganese zinc ferrite components are used in tuned circuits and magnetic power designs from the low kilohertz range into the broadcast spectrum. These ferrites have a linear expansion coefficient of approximately 10 ppm/°C.
- **Nickel zinc** - The nickel zinc ferrites will vary in porosity, and frequently contain the oxides of other metals, such as those of magnesium, manganese, copper or cobalt. Volume resistivities range from several kilohm-centimeter to tens of megohm-centimeter. In general, they are used at higher frequencies (above 1 MHz), and are suitable for low flux density applications. These ferrites have a linear expansion coefficient of approximately 8 ppm/°C.
- **Manganese** - The manganese ferrite is a dense, temperature stable material displaying a high degree of squareness in its hysteresis loop. This makes this material uniquely suited for such applications as multiple output control in switched-mode power supplies and high frequency magnetic amplifiers.

Ferrites are dense, homogeneous ceramic structures made by mixing iron oxide (Fe_2O_3) with oxides or carbonates of one or more metals such as manganese, zinc, nickel, or magnesium. They are pressed, then fired in a kiln at 2000°F, and machined as needed to meet various operational requirements.

Basic Differences- Composition and Structure

The difference in properties and performance of ferrites as compared with most other magnetic materials is due to the fact that the ferrites are oxide materials rather than metals. Ferromagnetism is derived from the unpaired electron spins in only a few metal atoms, these being iron, cobalt, nickel, manganese, and some rare earth elements. It is not surprising that the highest magnetic moments and therefore the highest saturation magnetizations are to be found in the metals themselves or in alloys of these metals. The oxides, on the other hand, suffer from a dilution effect of the large oxygen ions in the crystal lattice. In addition, the net moment resulting from ferromagnetic alignment of the atomic spins is reduced because a different, less efficient type of exchange mechanism is operative. The oxygen ions do serve a useful purpose, however, since they insulate the metal ions and, therefore, greatly increase the resistivity. This property makes the ferrite especially useful at higher frequencies. The purpose of this chapter is to list the various considerations which enter into the choice of a material for a specific application and to contrast pertinent ferrite properties with those of bulk metal or powdered metal materials.

Material Considerations - Magnetic and Mechanical Properties

Saturation Magnetization

As mentioned previously, the highest saturation values are found in the metals and alloys. Thus, if high flux densities are required in high power applications, the bulk metals, iron, silicon-iron and cobalt-iron are unexcelled. Since the flux in Maxwells $\varnothing = BA$, where B = flux density in Gausses and A = cross-sectional area in cm^2, obtaining high total flux in materials such as ferrites or permalloy powder cores can be accomplished only by increasing the cross-sectional area. Powdered iron has a fairly high saturation value, but exhibits low permeabilities.

Curie Temperatures

All magnetic materials lose their ferromagnetism at the Curie temperature. One overriding consideration for a magnetic material is that the Curie point of the material be well above the proposed operating temperatures. Table 1 lists the Curie Temperatures of the various materials. The Curie point depends only on composition and not on geometry. Even though some of the magnetic materials shown can be used at higher operating temperatures than others, very often the temperature limitations of the accessory items (wire insulation, potting or damping compound) can be more limiting; in this case, no practical advantage may be gained by the higher curie point materials.

Table 1 - Properties of Soft Magnetic Materials

Material	Initial Perm. μ_o	B max Kilo Gausses	Loss Coefficient $e \times 10^6$	Loss Coefficient $a \times 10^3$	Loss Coefficient $c \times 10^3$	Curie Temp. °C	Resistivity (Ohm-cm)	$\mu_o Q$ at 100 kHz	Operating Frequencies
Fe	250	22	-	-	-	770	10×10^{-6}	-	60 - 1000 Hz
Si-Fe (unoriented)	400	20	870	120	75	740	50×10^{-6}	-	60 - 1000 Hz
Si-Fe (oriented)	1500	20	-	-	-	740	50×10^{-6}	-	60 - 1000 Hz
50-50 Ni Fe (grain-oriented)	2000	16	-	-	-	360	40×10^{-6}	-	60 - 1000 Hz
79 Permalloy	12,000 to 100,000	8 to 11	173	-	-	450	55×10^{-6}	8,000 to 12,000	1 kHz-75kHz
Permalloy powder	14 to 550	3	0.01 to 0.04	0.002	0.05 to 0.1	450	1.0	10,000	10kHz-1MHz
High Flux powder	14 to 160	15	-	-	-	360	-	-	10kHz-1MHz
Iron powder	5 to 80	10	0.002 to 0.04	0.002 to 0.4	0.2 to 1.4	770	10^4	2,000 to 30,000	100 kHz-100 MHz
Ferrite-MnZn	750 to 15,000	3 to 5	0.001	0.002	0.01	100 to 300	10 to 100	100,000 to 500,000	10 kHz-2 MHz
Ferrite-NiZn	10 to 1500	3 to 5	-	-	-	150 to 450	10^6	30,000	200 kHz-100 MHz
Co-Fe 50%	800	24	-	-	-	980	70×10^{-6}	-	-

Magnetic Losses

The magnetic losses in an AC application can be represented by the Legg equation:

$$Rm = \mu fL(ef+aB_m+c)$$

where:
Rm= total core loss in Ohms
e= eddy current coefficient
a= hysteresis coefficient
c= residual loss coefficient
μ= magnetic permeability
f= frequency in Hertz
L= inductance in Henries
B_m = maximum flux density in Gausses.

Eddy current losses will increase quite rapidly with frequency. In bulk metals, these high frequency losses can be reduced by reducing the thickness of the material perpendicular to the flux flow. This is accomplished by using thin gauge tapes or laminations or by powdering and insulating the particles. In ferrites, the same result is obtained by increasing the resistivity by many orders of magnitude. Thus, at the highest operating frequencies where further gauge or particle reduction is impractical, ferrites are the only available materials.

The hysteresis losses are proportional to the flux density and can be depicted as the area inside the hysteresis loop. High hysteresis losses are accompanied by the presence of unwanted harmonics. The nickel-iron (permalloy) alloys have low hysteresis losses and a great asset to the permalloy powder core is that these low losses are maintained with the accompanying reduction in eddy current losses.

The residual losses are not too well understood and perhaps represent an expression of our ignorance of the system. They apparently are tied in partially to absorption of energy from the system by gyromagnetic resonance. A listing of the various losses in the materials is given in Table 1.

Permeability

Permeability is a function of composition and processing. The highest initial permeabilities (those measured at very low flux levels) are found in the nickel-iron alloys, particularly in supermalloy where the value is about 100,000. Powdered iron cores have low permeabilities (10-100) while permalloy powder cores are somewhat higher (15-550). Ferrites can be made over a wide range of permeabilities. The linear filter type permeabilities vary from 100-2000, while those used in power applications range from 3000-15,000. As the operating frequency increases, ferrites with lower permeabilities are used because these have distinctly lower losses in these regions. The permeabilities for a variety of materials are listed in Table 1.

Figure of Merit

A useful figure of merit for linear core materials is the μQ product. Values of this factor are tabulated in Table 1. At frequencies of 100 KHz and above, the value for ferrites is considerably above all other materials.

Squareness

The squareness ratio is defined as the ratio of B_r to B_m and is especially important in memory and switching core applications. Magnesium-manganese ferrites can be produced with extremely high squareness ratios. While some metal tape and bobbin cores possess similar high ratios, their higher cost and difficulty in miniaturization made the ferrites the material of choice in large scale memory applications in early computer models. Thin magnetic film memories, which may be considered bulk metals, have become increasingly important, and along with semiconductor, disk, tape and bubble memories, have replaced the old core memories. The importance of this phase is emphasized by the fact that the market value for computer magnetics is now greater in dollars to that of the power materials market. It is interesting to note that disk media and bubble materials are ferrite type oxides.

Brittleness

One drawback to the ferrite core as compared with metal cores is its brittleness. Being ceramic in nature, care must be exercised in the handling of these cores. Powder cores are also somewhat brittle and similar precautions are required. Although metal tape cores are not brittle, (except amorphous metal cores), they nevertheless are sensitive to strain and mechanical shock, especially in the high permeability materials. Consequently, tape wound cores are often embedded in a damping compound which prevents the transfer of strain or shock to the cores.

Hardness

Ferrites are very hard materials as compared with the other materials under consideration. This property is especially useful in applications in which wear is a factor. Consequently, ferrite material is being used extensively in magnetic recorder head applications.

Geometry Considerations

Formability

The three types of materials - bulk metal, powdered metal and ferrite - are produced by widely varying techniques and consequently the available geometries also vary.

- **Bulk metals** -These are produced mostly by standard metallurgical process involving melting followed by hot and cold rolling. The sheet material produced is either slit and wound into tape or bobbin cores or punched into laminations. Photoetching, a new method of forming small complex parts, avoids costly tooling, and produces stress-free parts.
- **Powdered Iron and permalloy** - These materials are always die-pressed into toroids or slugs, molypermalloy usually in toroids and powdered iron into slugs.
- **Ferrites** - Because ferrites are produced by a ceramic technique, they can be made in a large number of shapes. Unlike the bulk metals, they can be molded directly, and unlike the powdered permalloy, they can be machined and ground to close tolerances after firing. Various forming processes for ferrites include die pressing, extrusion, hydrostatic pressing and hot pressing. The available shapes include toroids, E-I cores, U-I cores, pot cores, rods, tubes, beads and blocks.

Tunability

An exact inductance is required in certain L-C circuits. If the shape of the inductor is toroidal, the inductance can be trimmed only by the addition or removal of turns, a time consuming and costly procedure. If a ferrite pot core is used, the tuning can be accomplished by means of a screw-type trimmer core which changes the effective air gap of the core. Threaded rods of powdered iron or ferrite materials are used extensively as tuning elements in slug tuned inductors.

Winding Considerations

Winding turns on a toroid involves specialized equipment and the process involves winding each core separately in a relatively time consuming operation. The bobbins used in ferrite pot cores can be wound many at a time on a rather simple machine. This ease of winding constitutes an important advantage for ferrite pot cores.

Magnetic Shielding

If magnetic components are relatively close in a circuit, the fields produced by one component may affect the performance of other cores. One solution is to increase the space between components. This increases the overall size of the system. Another is to use a magnetic shield which increases weight and size. A ferrite pot core is inherently self-shielding by nature of the enclosed magnetic circuit.

Inductance Stability Considerations

Temperature Stability

In telecommunications circuitry (tuned L-C filters), the maintenance of a near-constant inductance as a function of temperature and time is most critical. One method of achieving this stability is by the insertion of an air gap. The gap may be distributed as in powder cores or localized as in gapped ferrite pot cores. Gapping also results in a reduction in the effective permeability but often this is not a serious limitation. In gapped ferrite cores, the temperature coefficient (TC) can be linear to match a capacitor with an equal but opposite TC (polystyrene) or relatively flat if a flat TC capacitor (silver-mica) is used.

$$\mathrm{T}C = \frac{\Delta L}{L \Delta T}$$

where ΔL and ΔT are corresponding changes in inductance and temperature and L is inductance at a standard temperature.

As pointed out, the use of an air gap greatly increases the temperature stability. The powder core toroid and ferrite pot core are thus used to good advantage. In the powder core, the TC is built into the toroid, whereas in the pot core, the TC can be varied by changing the gap. However, in the latter, the effective permeability and therefore the inductance of the core is changed. By choice of the proper size core with the proper gap, the optimum inductance and TC can be obtained.

Permeability vs. AC Flux Density

It is often desirable to have a minimum change in permeability with AC excitation. Here again, the airgap in either permalloy powder cores or ferrites can be used to this advantage.

Permeability vs. DC Bias

Often an AC circuit has a superimposed DC bias condition. Minimum variation of permeability with DC is desirable. Powder cores are especially resistant to these changes. Gapped ferrite pot cores have a similar effect.

Permeability vs. Time

In most magnetic materials there is a slight decrease in permeability with time after the material is demagnetized or after it is first produced. This effect is known as disaccommodation. In nonlinear applications this effect is not too important. However, in low flux level circuits where a constant inductance is required, the effect must be considered. The effect is pronounced in low permeability materials and is negligible for high permeability materials. However, the effect can be minimized greatly by reduction of the effective permeability by insertion of an air gap. Thus in powder cores, the change of permeability due to this effect is less than 0.1%. In ferrite pot cores, the localized gap reduces the effect in proportion to the effective permeability compared with the toroidal permeability. Since the effect is logarithmic, most of the decrease occurs in the first few days after firing. If some aging of the cores occurs before usage, the change of inductance due to time will be negligible.

Application Information

Power Design

Ferrite is an ideal core material for transformers, inverters and inductors in the frequency range 5 kHz to 100 kHz, due to the combination of low core cost and low core losses. Tape wound cores do offer higher flux densities and better temperature stability, advantages which may off-set their higher cost.

Ferrite is an excellent material for high frequency (5 kHz to 500 kHz) inverter power supplies. Ferrites may be used in the saturating mode for low power, low frequency operation (<50 watts and 10 kHz). For high power operation a two transformer design, using a tape wound core as the saturating core and a ferrite core as the output transformer, offers maximum performance. The two transformer design offers high efficiency, excellent frequency stability, and low switching losses.

Ferrite cores may also be used in fly-back transformer designs, which offer low core cost, low circuit cost and high voltage capability. High frequency power supplies, both inverters and converters, offer lower cost, and lower weight and volume than conventional 60 Hertz and 400 Hertz power sources, at power levels up to 1 Kw.

Cores are gapped to avoid saturation under dc bias conditions.

Bobbins for many cores are available. VDE requirements have been taken into account in bobbin designs for EC, PQ, and metric E Cores. Many bobbins are also available commercially.

Core materials offering the lowest core losses and highest saturation flux density are most suitable for high power/ high temperature operation. Some material core losses decrease with temperature up to 70˚C, other material losses decrease up to 100˚C. Certain ferrite material is recommended for frequencies over 200 kHz.

CORE GEOMETRIES

Pot Cores

Pot cores, when assembled, nearly surround the wound bobbin. This aids in shielding the coil from pickup of EMI from outside sources. The core configuration provides a high degree of self-shielding and facilitates gapping to enhance its utility for a variety of magnetic designs. The pot core dimensions all follow IEC standards so that there is interchangeability between manufacturers. Both plain and printed circuit bobbins are available, as are mounting and assembly hardware. Because of its design, the pot core is a more expensive core than other shapes of a comparable size. Pot cores for high power applications are not readily available.

Double Slab Cores

Slab-sided solid center post cores resemble pot cores, but have a section cut off on either side of the skirt. Large openings allow large size wires to be accommodated and assist in removing heat from the assembly. Printed circuit or plain bobbins are available. Simple one piece clamps allow simple assembly. Low profile is possible. The solid center post generates less core loss and this minimizes heat buildup.

E Cores

E cores are less expensive than pot cores, and have the advantages of simple bobbin winding plus easy assembly. Gang winding is possible for the bobbins used with these cores. E cores do not, however, offer self-shielding. Lamination size E shapes are available to fit commercially available bobbins previously designed to fit the strip stampings of standard lamination sizes. Metric and DIN sizes are also available. E cores can be pressed to different thickness, providing a selection of cross-sectional areas. Bobbins for these different cross sectional areas are often available commercially.

E cores can be mounted in different directions, and if desired provide a low-profile. Printed circuit bobbins are available for low-profile mounting. E cores are popular shapes due to their lower cost, ease of assembly and winding, and the ready availability of a variety of hardware.

The E core geometry offers an economical design approach for a wide range of applications. In a power ferrite, E cores are used in a variety of power designs. In a high permeability material they are utilized for matching and broadband transformers.

EC and ETD Cores

EC cores are a cross between E cores and pot cores. Like E cores, they provide a wide opening on each side. This gives adequate space for the large size wires required for low output voltage switched-mode power supplies. It also allows for a flow of air which keeps the assembly cooler. The center post is round, like that of the pot core. One of the advantages of the round center post is that the winding has a shorter path length around it (11% shorter) than the wire around a square center post with an equal area. This reduces the losses of the windings by 11% and enables the core to handle a higher output power. The round center post also eliminates the sharp bend in the wire that occurs with winding on a square center post.

ETD cores have been designed to make optimum use of a given volume of ferrite material for maximum throughput power, specifically for forward converter transformers. Their structure, which includes a round center post, approaches a nearly uniform cross-sectional area throughout the core and provides a winding area that minimizes winding losses.

ETD cores are used mainly in switched-mode power supplies and permit off-line designs where IEC and VDE isolation requirements must be met.

PQ Cores

PQ cores are designed especially for use in power applications, mainly in switched-mode power supplies. These cores are employed in filter and transformer designs in switched-mode power supplies. The design provides an optimized ratio of volume to winding area and surface area. As a result, both maximum inductance and winding area are possible with a minimum core size. The cores thus provide maximum power output with a minimum assembled transformer weight and volume, in addition to taking up a minimum amount of area on the printed circuit board. Assembly with printed circuit bobbins and one piece clamps is simplified. This efficient design provides a more uniform cross-sectional area; thus cores tend to operate with fewer hot spots than with other designs. The large core surface area for the volume of the core aids in heat dissipation.

EP Cores

EP Cores are round center-post cubical shapes which enclose the coil completely except for the printed circuit board terminals. The particular shape minimizes the effect of air gaps formed at mating surfaces in the magnetic path and provides a larger volume ratio to total space used.

Shielding is excellent. The EP core design reduces the effect of residual air gap upon the effective permeability of the core, hence it minimizes coil volume for a given inductance. Also, the core geometry provides a high degree of isolation from adjacent components. EP cores are advantageously used in low power devices, matching and broadband transformers.

Toroids

The ring configuration provides the ultimate in the utilization of the ferrite material properties. Power input filters, ground-fault interrupters and a variety of pulse and matching transformers are only a few of the applications for the core type. Toroids are economical to manufacture; hence, they are least costly of all comparable core shapes. Since no bobbin is required, accessory and assembly costs are nil. Winding is done on toroidal winding machines. Toroids can be gapped to provide an air gap for dc bias applications. Shielding is relatively good.

Composite Toroids

Composite toroidal cores combines a high permeability toroid with a high saturation iron powder toroid and are supplied with a uniform coating of insulating enamel. Composite cores are particularly useful for choke applications that have large swings in the direct current as well as in some EMI suppressor designs with bias currents.

U Cores

The U core offers an economical core design with a nearly uniform cross-sectional area. In a low loss power ferrite material they are frequently used in output chokes, power input filters and transformers for switched-mode power supplies and HF fluorescent ballasts.

Summary

Ferrite geometries offer a wide selection in shapes and sizes. When choosing a core for power applications, parameters shown in Table 2 should be evaluated.

TABLE 2.
FERRITE CORE COMPARATIVE GEOMETRYCONSIDERATIONS.

	Pot Core	Double Slab Core	E Core	EC, ETD, Cores	PQ Core	EP Core	Toroid
Core Cost	high	high	low	med.	high	med.	very low
Bobbin Cost	low	low	low	med.	high	high	none
Winding Cost	low	low	low	low	low	low	high
Winding Flexibility	good	good	excellent	excellent	good	good	fair
Assembly	simple	simple	simple	med.	simple	simple	none
Mounting Flexibility*	good	good	good	fair	fair	good	poor
Heat Dissipation	poor	good	excellent	good	good	poor	good
Shielding	excellent	good	poor	poor	fair	excellent	good

*Hardware is required for clamping core halves together and mounting assembled core on a circuit board or chassis.

TEMPERATURE CONSIDERATIONS

The power handling capability of ferrite cores is limited by the temperature rise of the transformer. Major factors affecting this temperature rise include wire loss, core loss, and core geometry.

Wire Loss

For optimum efficiency, wire loss in the transformer should equal the core loss. For ferrite cores the ratio of wire loss to core loss can range from 4 to 0.25; the equivalent efficiency range is 80% to 90%, the 90% value occurring at a ratio of 1.0.

Wire loss with dc current increases with temperature (°C) as shown:
$R_t = R_{25} [1 + 0.004(T-25)]$

Where:
R_t = wire resistance at operational temperature
R_{25} = resistance at room temperature (25°C)
T = operating temperature (°C).

Skin effect losses and proximity losses become a significant factor at high frequencies. At 20 kHz (square wave), wire sizes larger than #24 will increase wire losses significantly. For high current / high frequency operations, stranded wire, litz wire, strip winding, or other special windings are recommended.

Core Loss

Core loss versus flux density and frequency is shown graphically for each power material. A core loss of 100 mW/cm^3 will result in an approximate temperature rise of 40°C. If the core loss due to the chosen flux density and/or frequency exceeds 100 mW/cm^3, special cooling precautions (heat sinks and/or fans) may be necessary.

For high power operation, the power handling capability is proportional to the temperature rise (e.g., a unit with a temperature rise of 80°C yields 1.5 to 2 times the power output of the same unit with a temperature rise of only 40°C); thus ΔT should be as high as possible. Transformer hot spot temperatures, however, should not exceed 120°C.

Cores driven with square waves exhibit slightly lower core losses than when driven with the sine waves; however, wire losses are much greater with square waves than with sine waves.

Core Geometry

Transformer temperature rise is directly related to the ratio of transformer surface area to transformer volume; larger cores are thus less efficient in radiating heat losses. For small cores (WaAc < 0.5) the graphical P_o will generate temperature rises $<40^\circ C$; for large cores (WaAc > 1.0) the graphical P_o will generate temperature rises $> 60^\circ C$.

E-U-I cores offer optimum cooling geometry; toroids are satisfactory; pot cores are least effective as the coil is shielded by the core.

The winding area/core area ratio (Wa/Ac) is only 0.4 to 0.5 for pot cores, and 0.6 to 1.0 for E cores, and ranges from 1 to 10 for toroids. Thus, for a given WaAc, toroids will accept more wire, but will also saturate more readily than E cores or pot cores.

Typical output power versus temperature rise behavior (for any geometry) as a function of square wave frequency and flux density is shown by graphs.

INDUCTOR CORE SIZE SELECTION (using Hanna curves)

When dc passes through a coil wound on an ungapped ferrite core, the flux in the core often reaches saturation, thus drastically reducing the permeability of the core. To minimize this effect, an air gap can be inserted in the core's magnetic path. This gap reduces the effective inductance of the core, but the inductance with dc in the coil is greater than that of a saturated ungapped core.

The amount of air gap needed to keep a ferrite core from saturating depends on the maximum dc current, the core geometry, and the minimum inductance required. A useful method of selecting the optimum core size and required air gap for a particular inductance and dc bias requirement is by the use of a Hanna curve.

To use the Hanna curve, first calculate your required LI^2, where L is in Henries and I in Amperes. Then choose a suitable value of LI^2 / VOLUME from the middle of the vertical scale; with these numbers, a core of suitable volume can be selected. The curve also determines gap size and number of turns. The first choice may be a core too small for the number of turns required, or it may be too large. However, the Hanna curve provides a simple cut and try method of selecting the proper core.

With practice in using the Hanna curve, one can become proficient in determining the proper core size and air gap required for dc bias applications.

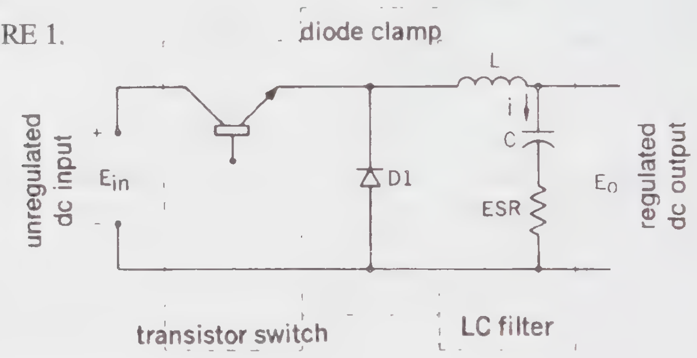
INDUCTOR CORE SIZE SELECTION (using core selector charts)

Description

A typical regulator circuit consists of three parts: transistor switch, diode clamp, and an LC filter. An unregulated dc voltage is applied to the transistor switch which usually operates at a frequency of 1 to 50 kilohertz. When the switch is ON, the input voltage, E_{in}, is applied to the LC filter, thus causing current through the inductor to increase; excess energy is stored in the inductor and capacitor to maintain output power during the OFF time of the switch. Regulation is obtained by adjusting the ON time, t_{on}, of the transistor switch, using a feedback system from the output (figure 1). The result is a regulated dc output, expressed as:

$$E_{out} = E_{in} \, t_{on} \, f$$

FIGURE 1.



Component Selection

The switching system consists of a transistor and a feedback from the output of the regulator. Transistor selection involves two factors -- (1) voltage ratings should be greater than the maximum input voltage, and (2) the frequency cut-off characteristics must be high compared to the actual switching frequency to insure efficient operation. The feedback circuits usually include operational amplifiers and comparators. Requirements for the diode clamp are identical to those of the transistor. The design of the LC filter stage is easily achieved. Given:

- maximum and minimum input voltage,
- required output,
- maximum allowable ripple voltage,
- maximum and minimum load currents, and
- the desired switching frequency.

The values for the inductance and capacitance can be obtained. First, off-time (t_{off}) of the transistor is calculated:

$$t_{off} = (1 - E_{out}/E_{in} \max) / f$$

When E_{in} decreases to its minimum value:

$$f_{min} = (1 - E_{out}/E_{in} \min) / t_{off}$$

With these values, the required L and C can be calculated.

Allowing the peak to peak ripple current (Δi) through the inductor to be given by:

$$\Delta i = 2 I_o \min$$

the inductance is calculated using:

$$L = E_{out} t_{off} / \Delta i$$

The value calculated for Δi is somewhat arbitrary and can be adjusted to obtain a practical value for the inductance. The minimum capacitance is given by:

$$C = \Delta i / 8 f_{min} \Delta e_o$$

Finally, the maximum ESR of the capacitor is:

$$ESR \max = \Delta e_o / \Delta i$$

Inductor Design

Ferrite E cores and pot cores offer the advantages of decreased cost and low core losses at high frequencies. For switching regulators, materials that are recommended because of their temperature and dc bias characteristics. By adding air gaps to these ferrite shapes, the cores can be used efficiently while avoiding saturation.

These core selection procedures will simplify the design of inductors for switching regulator applications. One can determine the smallest core size, assuming a winding factor of 50% and wire current carrying capacity of 500 circular mils per Ampere.

Only two parameters of the design application must be known:
(a) Inductance required with dc bias and (b) dc current.

1. Compute the product of LI^2 where:
L = inductance required with dc bias (milliHenries)
I = maximum dc output current = I_o max + Δi

2. Locate the LI^2 value on the Ferrite Core Selector charts. Follow this coordinate to the intersection with the first core size curve. Read the maximum nominal inductance, A_L, on the Y-axis. This represents the smallest core size and maximum A_L at which saturation will be avoided.

3. Any core size line that intersects the LI^2 coordinate represents a workable core for the inductor if the core's A_L value is less than the maximum value obtained on the chart. If possible, it is advisable to use standard gapped cores because of their availability.

4. Required inductance L, core size, and core nominal inductance (A_L) are known. Calculate the number of turns using:

$$N = 10^3 \sqrt{\frac{L}{A_L}}$$

where L is in milliHenries.

5. Choose the wire size from a wire table using 500 circular mils per Amp.

Low Level Applications

Pot Cores

The information contained in this section is primarily concerned with the design of linear inductors for high frequency LC tuned circuits using ferrite pot cores. The data presented in this section are compiled mainly for selecting cores for high Q resonant LC circuits. However, much of this information can also be used to design pot cores into many other applications, including high frequency transformers, chokes, and other magnetic elements.

A ferrite pot core assembly includes the following items:

- Two matched pot core halves
- Bobbin on which the coils are wound
- Tuning assembly
- A clamp for holding the core halves together.

The pot core shape provides a convenient means of adjusting the ferrite structure to meet the specific requirements of the inductor. Both high circuit Q and good temperature stability of inductance can be obtained with these cores. The self-shielded pot core isolates the winding from stray magnetic fields or effects from other surrounding circuit elements.

The effective permeability (μ_e) is adjusted by grinding a small air gap in the center post of the pot core. For transformers and some inductors, no ground air gap is introduced, and the effective permeability is maximized. The effective permeability of the pot core will always be less than the material initial permeability (μ_o) because of the small air gap at the mating surfaces of the pot core halves. For other inductors where stability of inductance, Q, and temperature coefficient must be closely specified, a controlled air gap is carefully ground into the center post of one or both of the pot core halves. When fitted together, the total air gap then will determine the effective permeability and control the magnetic characteristics of the pot core. Finer adjustment of the effective permeability (gapped pot core inductance) can be accomplished by moving a ferrite cylinder or rod into the air gap through a hole in the center post.

Ferrites are available in various initial permeabilities (μ_o) which for filter applications cover frequency ranges into the megaHertz region. Manufacturers produces a wide variety of pot core sizes. Each pot core half is tested and matched with another half to produce a core with a specified inductance tolerance.

Advantages of Pot Core Assemblies

- SELF-SHIELDING
  Because the wound coil is enclosed within the ferrite core, self-shielding prevents stray magnetic fields from entering or leaving the structure.

- COMPACTNESS
  Self-shielding permits more compact arrangement of circuit components, especially on printed circuit boards.

- MECHANICAL CONVENIENCE
  Ferrite pot cores are easy to assemble, mount, and wire to the circuit.

- LOW COST
  As compared to other core materials, ferrites are easier to make in unusual configurations (such as pot cores), resulting in a lower cost component. In addition, winding a pot core is usually quick and inexpensive because coils can be pre-wound on bobbins. When other costs of assembly, mounting, wiring, and adjustment are added, the total cost is often less than with other core materials or shapes.

- ADJUSTABILITY
  Final adjustment is accomplished by moving a threaded core in and out of the centerpost, and adjustment in the field is relatively easy as compared to any other type of construction.

- IMPROVED TEMPERATURE STABILITY AND Q
  Air gaps inserted between the mating surfaces of the center-posts provide good temperature stability and high Q.

- WIDE CORE SELECTION
  Many combinations of materials, physical sizes, and inductances offer the design engineer a large number of choices in core selection.

- LOW LOSSES AND LOW DISTORTION
  Since ferrites have high resistivities, eddy current losses are extremely low over the applicable frequency range and can be neglected. Hysteresis losses can be kept low with proper selection of material, core size, and excitation level.

- CHOICE OF LINEAR OR FLAT TEMPERATURE CHARACTERISTICS
  Provides a close match to corresponding capacitors.

- CONSISTENCY AND UNIFORMITY
  Modern equipment with closely controlled manufacturing processes produce ferrite pot cores that are magnetically uniform, not only within one lot but from lot to lot.

- ONE PIECE CLAMP
  Some manufacturers offer a clamp that provides simple assembly of the two core halves. Easy bending action allows insertion of the core assembly into the clamp, and spring tension holds the assembly rigidly and permanently in place. Rivet, screw, or circuit board tab mounting is available.

Pot Core Design Notes Important Considerations

The selection of a pot core for use in LC resonant circuits and high frequency inductors requires a careful analysis of the design, including the following:

- Operating frequency
- Inductance of the wound pot core assembly
- Temperature coefficient of the inductor
- Q of the inductor over the frequency range
- Dimensional limitations of the coil assembly
- Maximum current flowing through the coil
- Long term stability.

The important characteristics which strongly influence the above requirements are:

1. Relative loss factor - $1/\mu iQ$. This factor reflects the relative losses in the core and varies with different ferrite materials and changes in operating frequency. When selecting the proper material, it is best to choose the one giving the lowest $1/\mu iQ$ over the range of operating frequencies. In this way, the highest circuit Q can be expected. In a situation where the $1/\mu iQ$ curves may cross over or coincide at various frequencies, each ferrite material should be considered in view of all circuit parameters of importance, including size, temperature coefficient, and disaccommodation, as well as Q. With this analysis, little doubt is left concerning the optimum selection of a proper core material.

2. Inductance factor (A_L). The selection of this parameter is based on a logarithmic progressive series of values obtained by dividing a logarithmic decade into 5 equal parts (International Standardization Organization R5 series of preferred numbers). Since the A_L values for the various core sizes are standard, they may be graphed or charted for ease of determining the required turns (N) to give the value of inductance needed. Pot cores with various A_L values are obtained by grinding closely-controlled air gaps in the centerposts of the cores. Small gaps are processed by gapping one core half. For larger gaps, both halves are gapped.

3. Temperature Coefficient (TCe). The temperature coefficient of the pot core is important in LC tuned circuits and filters when attempting to stabilize the resonant frequency over a wide range of temperatures. This temperature coefficient (TCe) is determined by the properties of the ferrite material and the amount of air gap introduced. Ferrite materials have been designed to produce gapped pot core temperature coefficients that balance the opposite temperature characteristics of polystyrene capacitors, or match similar flat temperature coefficients of silvered mica capacitors. Therefore, careful selection of both capacitors and pot cores with regard to temperature coefficient will insure the optimum temperature stability.

4. Quality Factor (Q). The quality factor is a measure of the effects of the various losses on circuit performance. From the designer's point of view, these losses should include core losses, copper losses, and winding capacitive losses. Therefore, Q will be affected greatly by the number and placement of the turns on the bobbin, and the type and size of wire used. At higher frequencies, litz wire would reduce the eddy current losses in the windings and produce a higher Q than solid wire. Q data include the effects of winding and capacitive losses, which, if removed, would produce significantly higher calculated Q values. Consequently, the Q curves represent more realistically the actual Q values that would be obtained from circuit designs.

5. Dimensional Limitations. Many circuit designs contain dimensional and weight limitations which restrict the size of the inductor and the mounting techniques used. Sometimes, minimum weight or volume is sacrificed to obtain better circuit performance.

6. Current Carrying Capacity. Inductive circuits containing ferrite pot cores are normally operated at extremely low levels of AC excitation to insure the best possible performance. However, the current flowing in the coil may be much higher than anticipated due to superimposed DC currents, or unexpected surges of AC. Therefore, the selection of the wire size used in an inductor design is influenced by both of these factors. Wire data is presented as a guide in considering these operating conditions.

7. Long Term Stability (DFe). In critical inductive designs, especially resonant circuits, the designer must be concerned with long term drift in resonant frequency. This stability drift (or decrease in inductance), known as disaccommodation, can be calculated for each pot core size and inductance factor (A_L). It occurs at a logarithmic rate, and the long term change of inductance may be calculated from the formula:

$$\frac{\Delta L}{L} = DF_e \times \log \frac{t_2}{t_1}$$

where:

ΔL is the decrease in inductance between the times t_1 and t_2.
DF_e is the Effective Disaccommodation Coefficient of the core selected.
t_1 and t_2 is the elapsed time between manufacture of the core (stamped on shipping container) and its assembly into the circuit, while t_2 is the time from manufacture of the core to the end of the expected life of the device.

Disaccommodation starts immediately after the core is manufactured as it cools through its Curie Temperature. At any later time as the core is demagnetized or thermally or mechanically shocked, the inductance may increase to its original value and disaccommodation begins again. Therefore, consideration must be given to increases in inductance due to magnetic, thermal or physical shock, as well as decreases in inductance due to time. If no extreme conditioning is expected during the equipment life, changes in inductance will be small, because most of the change occurs during the first few months after manufacture of the core.

Limits on Excitation

Inductors designed using pot cores are usually identified as linear magnetic components because they are operated within the range of negligible change of effective permeability with excitation To calculate suggested maximum AC excitation levels, use the following formula:

$$B = \frac{E_{rms} \times 10^8}{4.44\ A_e\ N\ f}$$

4.44 for sine wave
4.0 for square wave

where:
B = 200 gausses, the suggested conservative limit
N = turns on pot core
f = operating frequency in Hertz
A_e = equivalent area of the pot core in cm^2.

Because superimposed DC current also affects linearity of inductance in pot cores, consideration for DC currents must also be given. The equation shown above must be modified to include effect of DC bias. The combined equation now becomes:

$$B_{(combined)} = \frac{E_{rms} \times 10^8}{4.44\ A_e\ N\ f} + \frac{N I_{dc}\ A_L}{10 A_e}$$

where:
B = 200 Gausses, the suggested conservative limit
I_{dc} = bias current in Amperes.

Material Selection

In most designs the ferrite material type can be obtained by knowing circuit operating frequencies. The material type that gives the lowest 1/µiQ over the desired frequency range is normally selected.

Pot Core Selection

Various methods are used by engineers and designers to select the best pot core size for an application. These vary from a random trial approach to the more exact techniques involving the use of Q curves or IS0-Q contour curves. Manufacturer's catalogs include many graphs, charts, and tables that simplify selection techniques.

Q curves are developed from standard Q versus frequency measurements using coils with different numbers of turns and wire sizes. From the data obtained, standard Q curves are plotted and are available upon request. For convenience in selecting the optimum core, the TCe and bobbin area are also listed with the Q graphs.

The only other remaining factors that must be determined are the exact wire size, dimensions, and the effects of disaccommodation. This information can be obtained from graphs, charts, or formulas.

Depending on the requirements of the application, certain factors become more critical than others and will influence the design procedure used. For example, in an LC tuned circuit, if highest Q is required, larger cores must be used, and the C of the capacitor is usually obtained after determining the optimum L of the inductor. However, if size is critical, and maximum inductance is required to fit an available space, highest Q and optimum TC must be sacrificed. If all design factors are critical (as is now frequently demanded), each must be carefully evaluated with respect to its effect on the others to arrive at the best design.

The following is a review of several common types of application problems found in inductor design. Included is a suggested procedure to be followed to obtain an optimum solution:

Application 1. Frequency and Inductance are specified. Smallest size is important. Q or temperature stability is not critical.

Solution. For the frequency range specified, select the best suited core material, using the 1/μiQ curves.

Find the smallest core size with the highest A_L available in the material selected, by referring to the data listed for each size. For this A_L, determine the turns required to meet the inductance specified, by using the manufacturer's graph.

Knowing the core size, select the maximum wire size from a graph. As an alternate procedure, divide the turns by the Bobbin area to obtain the turns per square inch and refer to the Wire Tables for maximum wire size. If this wire size doesn't exceed the one that can economically be used, the design is complete. If it does, select the next larger core size and the highest A_L, and recheck until a satisfactory wire size is obtained.

Application 2. Frequency is specified. Q must be maximum.
The L consistent with this maximum Q will be used to determine the C in a resonant circuit. Size or temperature stability is not critical.

Solution. Select the core material best suited for the frequency range using the $1/\mu iQ$ curves. Scan the Q curves for the material chosen to find the core size and A_L that give the highest Q over the frequency range specified. Record the core size, A_L, and inductance L shown for that Q.

For the A_L chosen, obtain the turns required to produce the inductance L. Knowing the core size, select the wire size. The use of litz wire will usually increase Q.

Further increases in Q may be accomplished with the use of double or triple section bobbins or by spacing the winding away from the air gap.
Determine the C required for resonance.

Application 3. Frequency, inductance, minimum Q, temperature coefficient, and maximum size are specified.

Solution. Using the $1/\mu iQ$ curves, select the core material best suited for the frequency range. Scan the Q curves for the material chosen and select the core sizes and A_L's that give the desired TCe. List these cores.

Using the Q curves, eliminate those sizes that do not meet the minimum Q at the frequency and inductance value specified. From the remaining cores, choose the smallest size. Determine the number of turns required to meet the inductance. As an alternate procedure, divide the turns by the bobbin area to obtain the turns per square inch and refer to wire tables for the maximum wire size. The use of litz wire will usually increase Q.

Check the design dimensions against the core size. If the maximum size is exceeded, Q and inductances must be re-evaluated, and a new core size selected.

Note: While not mentioned in the above applications, critical LC tuned circuits must remain stable over long periods of time. Disaccommodation, the decrease of inductance over time, can be calculated from the formula, $\frac{\Delta L}{L} = D\,F_e \log\frac{t_2}{t_1}$, as previously noted.

From the calculated decrease of inductance, the shift in the resonant frequency can be predicted. Disaccommodation can be ignored if the shift is within limits.

Core Selection Charts

These tables summarize the design procedures discussed in the preceding section.

Application 1.
Frequency and Inductance specified. Size must be minimum.

Steps	Operation
1.	Select core material.
2.	List smallest core size with highest A_L.
3.	Find turns required.
4.	Select wire size.
5.	If too small to wind economically, repeat steps 2, 3, and 4 for next larger core size.

Application 2.
Frequency specified with Q to be maximum.
C to be determined from L when Q is maximum.

Steps	Operation
1.	Select core material.
2.	Find core with highest Q over frequency range.
3.	Record core size A_L, and inductance L.
4.	Obtain turns required.
5.	Select wire size.
6.	Determine C for resonance using formula $C = 1/(2\pi f)^2 L$.

Application 3.
All requirements equally important.

Steps	Operation
1.	Select core material.
2.	List all cores with acceptable TCe values.
3.	From cores in step 2 select those with acceptable Q at the inductance and frequency specified.
4.	Using the smallest core selected in 3, find the turns N.
5.	Select wire size.
6.	Check size of core against design dimensions.
7.	If size is exceeded, re-evaluate Q and and repeat steps 2 to 6 to select a smaller core.
8.	Calculate decrease of inductance for life of circuit.
9.	Evaluate AC and DC bias currents to insure wire size is adequate.

Pot Core Assembly Notes

Ferrite pot cores can be assembled with or without clamping hardware or tuning devices. These clamps normally eliminate the need to cement the pot cores halves together. The mating surfaces of the pot core must be cleaned to remove moisture, grease, dust, or other foreign particles, before clamping or cementing.

If the cementing method is chosen, a small amount of cement is placed on the mating surface of the pot core skirt, being careful to keep the centerpost free of all cement. The pot core halves are brought together and rotated together under slight pressure to distribute the cement evenly around the skirt. The halves are separated and the wound bobbin is set in place. A small amount of cement is now placed on the exposed flange of the bobbin to bond it in the pot core assembly and thus insure no movement. The other core half is replaced, the centerpost holes and wire apertures aligned, and the unit clamped together in a pressure jig. Permanent bonding is accomplished by curing the cement at elevated temperatures according to the manufacturer's recommendations. After curing, storage for a minimum of 24 hours, and heat cycling between room temperature and 70° C are suggested before final testing or tuning is completed.

The tuning adjusters can be inserted into the pot core immediately after the cemented core halves have been cured and the assembly can then be heat cycled. Some adjusters require insertion of the base into the centerpost hole prior to assembly of the pot core into the clamp when a clamp is used for mounting. The adjuster is usually made in two parts-the plastic base with a threaded hole, and a ferrite cylinder imbedded in a plastic screw. The base is pressed into the centerpost of the pot core, and the plastic screw is turned into the base until the ferrite cylinder enters the air gap. Tuning is completed when the inductance of the pot core assembly reaches the proper value. If this initial adjustment is expected to be the final one, cementing is recommended to prevent accidental detuning. If precise inductance values are expected, final tuning should not be completed earlier than 24 hours after the pot core assembly has been cured or clamped.

All "TB-P" bases, which are polypropylene, must be etched in order to roughen the adhering surface, if they are to be cemented to the core.
Plastic screw drivers are available for use in final tuning.

PRINTED CIRCUIT BOBBINS AND MOUNTING HARDWARE

Many sizes in the standard pot cores can be supplied with printed circuit board bobbins. The grid pattern illustrates the location of pin type bobbins. The pin length must be sufficient for the board thickness. Terminal pin details should be illustrated. The board holes should be specified by diameter (# drill). The bobbin should be cemented to the lower pot core half.

For some core types, printed circuit board mounting clamps are also available. When clamps are not available, the pot core halves must be cemented together.

PRINTED CIRCUIT BOBBINS SOLDERING INSTRUCTIONS

1. A solder pot should be used to solder the leads to the terminals. Preferred solder is 63/37 tin/lead eutectic. The solder temperature should be between 525-575°F. Lower or higher temperatures will both damage the bobbin. Modern soldering techniques commonly use temperatures in excess of the softening points of all thermoplastic bobbin materials. Extreme care is required to prevent loosening of the terminals during soldering.

2. Insulation should be removed from the ends of the wire before soldering. This is especially important when litz wire is used. The preferred method is by burning. Evaluate AC and DC bias currents to insure wire size is adequate.

3. Dip wound terminals into liquid soldering flux. A rosin based flux in alcohol solution should be used. Allow flux to air dry.

4. The bobbin should be immersed only far enough to cover the terminals.

5. The part should be immersed in the solder for 2-4 seconds, depending on the size of the wire used.

TOROIDS

Ferrite toroids offer high magnetic efficiency as there is no air gap, and the cross sectional area is uniform. They are available in many sizes (O.D.'s from 0.100" to 3.375"), and materials (permeabilities from 750 to 10,000) are typical examples.

Ferrite toroids provide an often convenient and very effective shape for many wide band, pulse and power transformers and inductors. The continuous magnetic path yields the highest effective permeability and lowest flux leakage of any shape.

BASIC CONSIDERATIONS

The inductance may be calculated from equation 1.

Equation 1. $L = \frac{0.4\pi \mu N^2 A_e}{\ell_e} \times 10^{-8}$ Henries

Here ℓ_e and A_e are the effective magnetic path length and cross sectional area of the core, μ is the effective permeability of the material, and N is the number of turns. This formula may be used for any shape under all conditions provided the correct value of μ is used and stray reactances are given proper consideration. In a toroidal core, this may be expressed as shown in equation 2:

Equation 2. $L = 2 \mu N^2 \, 2.54 \, H \, \mathrm{Ln} \frac{OD}{ID} \times 10^{-9}$ Henries

where OD, ID, and H are the dimensions in inches. For low level conditions at comparatively low frequencies, the formula may be simplified by using the Inductance Index, AL, used by many manufacturers in their catalogs. Then, equation 3:

Equation 3. $L = N^2 A_L$ nanoHenries

The other value most frequently needed is peak flux density, which may be calculated from equation 4:

Equation 4. $$B = \frac{E}{4.44\,A_e\,N\,f} \times 10^8 \text{ Gauss}$$

Here E is the RMS voltage, 4.44 is a constant depending on the wave shape (use 4 when E is symmetrical square wave and 1 where E is a unipolar pulse), and f is the frequency in Hertz.

LOW LEVEL INDUCTORS

This section considers those applications where nonlinearity and losses due to hysteresis are negligible. Generally this means flux densities below a few hundred Gauss. The first material choice is the one having both the highest permeability and lowest loss factor, $\tan\delta / \mu$ at the operating frequency. Considering the space available, select a core from the table and, using its inductance index, A_L, calculate the number of turns required to give the desired inductance. Now select the largest practical wire size that will fit on the core. This is somewhat difficult for a toroid, but generally the total wire cross section in the winding can be 30-60% of the window opening. If there are Q or loss requirements calculate the resistance of the winding, taking into consideration the skin effect if the frequency is high, and add it to the equivalent series resistance contributed by the core losses. Equation 5 shows the relationship between loss factor, Q and resistance.

Equation 5. $$\tan\delta / \mu = \frac{1}{\mu\,Q} = \frac{R}{\mu\,2\,\pi\,f\,L}$$

If the calculated Q is inadequate, you must reduce the total series resistance by selecting a larger core that will allow fewer turns of larger wire. Select a less lossy material, or use Litz wire at high frequencies to minimize the skin effect.

If losses are critical, it is important to remember that hysteresis losses have been assumed to be negligible. Above a few 10's of Gauss these losses are measurable and increase as approximately the 2.5 power of flux density. Also, remember that ferrites like other magnetic materials show variation in inductance from part to part, with temperature and with magnetizing force. Unlike powdered metals which have air gaps between the particles a ferrite toroid is a continuous magnetic material with variability effects undiluted by air gaps. This means that tight tolerances such as required for wave filters are not attainable in a toroid, but will generally require a gapped structure such as an E core, pot core, or slug.

POWER INDUCTORS

In this section, we consider inductors where the design is limited by saturation or heating due to core or winding losses. Although there is no systematic connection between permeability and losses, below about 1 MHz the relatively high permeability manganese-zinc ferrites have the most desirable combination of high saturation flux density and low hysteresis losses. The first step is to select one of those materials having the desired properties and select a core based on space limitations. Then select a suitable operating flux density. As a general rule, at room temperature materials may be operated to the knee of the BH loop when the frequency is 20 kHz or less. At higher frequencies hysteresis losses produce enough heat to require that the flux density be decreased. As a first approximation, the product of flux density and frequency can be held constant above 20 kHz. Knowing the voltage, frequency, flux density and area of the chosen core the minimum number of turns may be calculated from equation 4. The inductance can then be estimated from equation 3 or calculated more exactly from equations 1 or 2 by using the appropriate value of permeability under these operating conditions. If this inductance is less than the desired value, the number of turns can be adjusted upward provided there is sufficient space for the winding. If the inductance is too great it will be necessary to choose a larger core whose cross sectional area is greater but whose ratio of A_e/ℓ_e, is less, or a material with lower permeability.

For inductors operating above 1 MHz the material choice becomes more difficult since other requirements such as return loss may be more important. The material choice and design procedure will depend on which factors predominate in your particular design.

Inductors having dc current superimposed on the ac excitation must be given special treatment. The magnetizing force may be calculated by using equation 6:

$$\text{Equation 6. } H = \frac{0.4\,\pi\,N\,I}{\ell_e} \text{ Oersteds}$$

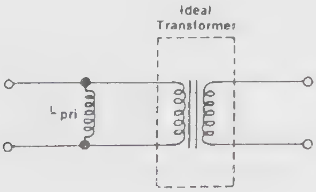
With this information it is possible to estimate from the BH curves how significant will be the effect of the dc current. Generally, dc magnetizing forces less the coercive force will have only a small effect on permeability, moderate values will depress the permeability, and magnetizing forces approaching the knee of the BH loop will considerably reduce the permeability and severely limit the peak flux density available for ac excitation. In these cases, unless a higher inductance can be used it will be necessary to go to a core with a considerably longer magnetic path length or to provide an air gap such as by slotting the core.

In many power applications, thermal considerations control the design. One rule of thumb that may be useful for first approximations is that core losses of 100 to 600 mW/cm^3 produce an approximate 40°C temperature rise. The exact value depends on inductor geometry and thermodynamic considerations beyond the scope of this chapter. You must also consider the power dissipated in the winding and its contribution to inductor heating. Heat sinking or coolants may be used to remove this heat, but the thermal conductivity of ferrite is relatively low, so the interior core temperature will be higher. Should a large temperature gradient develop, the core may crack from thermal stresses. Also, where considerable temperature excursions occur due either to self heating or ambient temperatures, the effect of these changes must also be considered with respect to changes in saturation flux density and inductance.

LOW LEVEL TRANSFORMERS

The design procedure here is essentially the same as for low level inductors except, of course, that the winding space must be shared between the primary and secondary windings. Usually half the space is allotted to each. In selecting the inductance required it is easiest to envision the equivalent circuit as an ideal transformer (figure 2) with a primary self inductance shunting the transformer primary.

FIGURE 2.

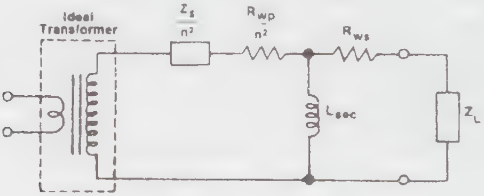


When the impedance represented by this inductance is high compared to the primary and transformed secondary impedances it may be neglected and an ideal transformer results. Ordinarily this impedance is selected to be between 3 and 10 times the source impedance. At very high frequencies losses or winding capacitance and leakage inductance may predominate. These situations are considered in the later paragraphs.

CURRENT TRANSFORMERS

This special class of low level (and sometimes power) transformers includes ground fault interrupters (GFI) sensors. In this case it is simpler to design around the secondary. Because there are often few primary turns (usually one) and many secondary turns, the transformed source impedance (Z_s) and primary winding resistance (R_{wp}) can not always be neglected. These impedances are increased by $1 / n^2$ (where n is the primary to secondary turns ratio). As n is decreased to raise the secondary voltage all four internal impedances shown in figure 3 increase. This limits the available load voltage, so a compromise must be made for optimum performance. Since the core losses of high permeability ferrites are small at audio frequencies, they may often be neglected. For this reason, ferrite toroids are usually selected for grounded neutral transformers in GFI's--particularly when frequencies above 60 Hz are used for this test. Special manufacturing and test techniques can be used to enhance the properties of ferrite toroids for GFI differential fault transformers, as well.

FIGURE 3.



POWER TRANSFORMERS

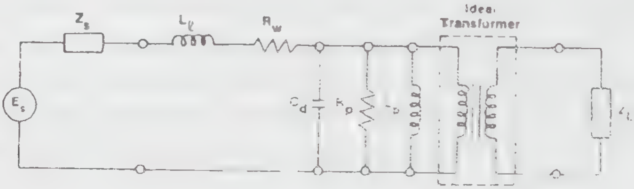
Here we are considering the same kinds of situations we covered under power inductors, that is, those cases where the design is limited by saturation flux density or self-heating due to core and winding losses. At low frequencies, say below 1 MHz, the design procedure is the same as that for power inductors except, of course, that winding space must be allowed for both windings. Ordinarily allot half each to the primary and secondary, or with a push-pull primary, slightly less than one third to each primary half. In most cases the voltage and frequency are known (use the lowest operating frequency for design purposes). Select a material and flux density in the same manner as for power inductors. Then using equation 4, calculate the product of A_e and N required. It is then a simple matter to go down the list of suitable core sizes substituting for A_e, calculating the minimum number of turns required and checking the fit of the winding in that core. Calculating the primary inductance from equations 1, 2, or 3, you will ordinarily find that the inductance will be large enough that the magnetizing current may be neglected under full load. (This is the current drawn by the primary inductance which shunts the ideal transformer). The rest of the transformer design is fairly straight-forward and is covered in other publications. Most devices of this type are limited by either saturation or heat dissipation, temperature rise and efficiency. Often winding losses are greater than core losses below 50 kHz. In some cases other considerations such as regulation may take precedence, but the considerations described above must still be met.

At higher frequencies--in the MHz range--other factors such as eddy currents influence the design. For this reason higher resistivity nickel zinc ferrites are ordinarily used. Furthermore, winding design can be of major importance because of the critical nature of winding losses (including skin effects), leakage inductance and self capacitance. Again, cooling is often a major problem and increasing core size is limited by its effect on winding characteristics. It is sometimes helpful to assemble the core as a stack or two stacks of a number of smaller toroids since this facilitates cooling, and results in a compact winding. Occasionally oil cooling or heat sinking are used to improve heat transfer. Material selection is difficult because of the influence of several factors which do not lend themselves to analytical prediction. Lacking previous experience with a similar design, some guesses will have to be made. A good starting point is that material having the lowest loss factor at the minimum operating frequency. A trial design can be worked up using the same core selection criteria as at lower frequencies. Usually the flux density will have to be limited to a few hundred Gauss or less. Care should be taken to select a core which will allow a compact winding so that leakage inductance and winding self capacitance will be small. Winding design requires careful consideration also because skin effects will make the winding resistance (and, hence, loss) much greater than at low frequencies. A technique popular when one winding is a single turn is to use tubing. The wall thickness should be chosen to be slightly more than the current penetration depth, and the secondary winding can go within the tubing. Litz wire can also be used to reduce the effective resistance. A trial design and a few iterations are usually required to optimize RF power transformer designs.

WIDE BAND TRANSFORMERS

The best starting point is with the equivalent circuit shown in figure 4.

FIGURE 4.



Here L_p and R_p are the parallel inductance and resistance (loss) of the wound core, R_w is the winding resistance, C_d is the distributed self capacitance of the winding, L_ℓ is the leakage inductance (representing flux that does not link the core), and Z_s and Z_L are the source and load impedances. At low frequencies the contribution of L_ℓ and C_d are so small they may be neglected. The low frequency cut-off, where insertion loss, VSWR or source loading become unacceptable, is then determined by L_p, R_p, and R_w. Since the reactance of L_p ($X = 2\pi f L_p$) is proportional to frequency, it is usually the determining factor. The objective is then to choose a core, material and winding that will have the highest L_p and R_p at the lower frequency while keeping R_w small. To do this, select a material having high permeability and low loss at that frequency. Choosing a core with a high A_L, it must be wound so that L_p and R_p are high enough and R_w low enough to meet the insertion loss, VSWR, return loss or loading requirements. At the high frequency cut-off, L_p can usually be neglected while L_ℓ and C_d assume critical importance. These elements depend almost entirely on the winding and very little on the core. They can not be readily calculated, but are minimized by keeping the winding length and number of turns low. The optimum core is difficult to select since it must balance these considerations with winding space, ease of winding, integer turns, space limitations and core manufacturing constraints. Generally, it is best to choose a core with a large OD/ID ratio and the greatest practical height. For this reason high frequency wide band transformers are often wound on cores found in the BEAD and MULTI-HOLE sections.

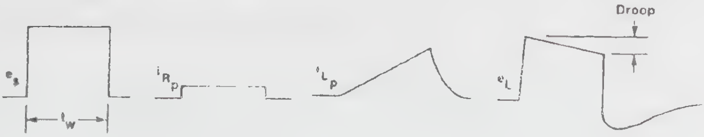
There are also techniques covered in this chapter on winding transformers with transmission lines such that at low frequencies the device operates conventionally as above. At higher frequencies coupling is via the transmission line enabling extension of the upper operating limit.

You will find curves of X_p, R, and Z versus frequency for certain cores. This data simplifies material, core and winding selection. With the exception of the highest frequencies, these curves may be shifted upward or downward to fit a given application by the ratio of N^2 of the new winding to N^2 indicated on the graph.

PULSE TRANSFORMERS

In many ways pulse transformers are a special case of wide band transformer because the pulse train can be represented by a number of sine waves of different frequencies. The turns ratio, though, is usually determined by voltage or current ratios rather than impedance matching, so the design approach is governed by pulse fidelity requirements rather than insertion or return loss. The equivalent circuit of figure 5 can help illustrate the elements influencing fidelity. Looking first at the flat top (low frequency) portion of a rectangular pulse (e_s), figure 5 shows some of the voltage and current wave shapes.

FIGURE 5.



Neglecting the rise and fall portions (high frequencies), current through R_p is constant during the pulse and current through L_p flows according to equation 7:

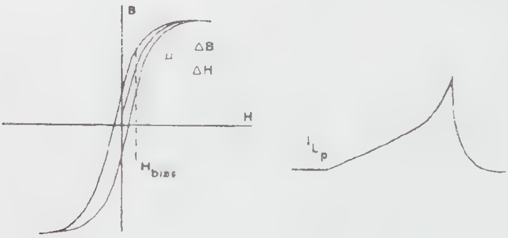
Equation 7. $i_{L_p} = \int \frac{e}{L_p}\, d\,t \approx \frac{e\,t}{L_p}$ Amperes

If the voltage and inductance are constant the current will rise linearly with time. This produces a drop across Z_s accounting for droop of the load voltage pulse (e_L). In order to minimize droop, L must be made large. This can be accomplished by choosing the highest permeability material and largest core practical. To determine the A_L value under pulse conditions, multiply the sine A_L by 1.1. L_p can then be calculated from equation 3. Also, flux density must be considered. Equation 4 may be rewritten as equation 8:

Equation 8. $B = \frac{e\,t}{A_e\,N} \times 10^8$ Gauss

It can be seen that flux density rises linearly with time. As this approaches the knee of the hysteresis (B-H) loop, permeability and inductance start to fall and the current begins to rise rapidly (figure 6).

FIGURE 6.



From equation 3 and 8, you can see that increasing N will both raise L and diminish B. However, rise and fall time are limited by leakage inductance (L_ℓ) and distributed self capacitance (C_d) in the same way as high frequency response in a wide band transformer. Therefore, the number of turns must be balanced between these conflicting requirements. The tools available are higher permeability and flux density material, and a larger core.

High pulse repetition rate can have two effects. The dc level represented by averaging the pulses produces magnetizing force (H) to bias the starting point of each pulse to the right on the B-H loop (figure 6). This can significantly reduce the available flux density. One possible solution is described under Slotted Toroids. Second, each pulse traverses a minor hysteresis loop producing an energy loss. This can cause core heating that will affect saturation flux density and permeability.

SLOTTED TOROIDS

In a number of applications described above the design is limited either by dc current, excessive inductance, or variability effects of the ferrite. A slot cut through the cross section can sometimes be used to this advantage. The effect of the gap is magnified by the material permeability according to:

Equation 9. $\ell_e = \ell_m = \mu \ell_g$

Where ℓ_m and ℓ_g are the path length in the magnetic material and the gap respectively and μ is the material permeability. This can be used to reduce the effect of dc bias when the ℓ_e calculated above is substituted into equation 6. For example: 1 Adc flowing through 10 turns on a core with a path length of 2 cm produces a magnetizing force (H) of 6.28 Oe. This is enough to saturate most high permeability materials. If a 0.010" (0.0254 cm) slot is cut and the material permeability is 5000, the effective path length (from equation 9) is 129 cm. The magnetizing force from the dc is reduced to 0.097 Oe and the effect of the dc bias is very small. In similar fashion a gap can be used to reduce inductance to the required value when the minimum turns are dictated by flux density considerations. The effective permeability of a gapped core can be calculated from equation 10:

Equation 10. $\mu_e = \dfrac{\mu \ell_m}{\ell_m + \mu \ell_g}$

This value of μ can be used with equations 1 or 2 to calculate inductance.

It is also apparent from equation 10 that as ℓ_g is increased, relative ℓ_m changes in μ will have a smaller effect on μ_e. This can be used to reduce changes in inductance caused by permeability variations due to temperature, flux density, bias, stress, time, etc. For example, with 5000 permeability material and $\ell_g / \ell_m = 0.01$, a 20% change in μ will result in only a 0.2% change in μ_e.

Equations 9 and 10 are exact only when there is no flux fringing in the gap. This is a good assumption when $A >> \ell_g$, but as the gap increases the actual μ_e will be greater than the calculated value and actual ℓ_e will be less. More elaborate equations can extend the range of accuracy somewhat, but with larger gaps some experimentation is necessary. A wide range of slot widths are available.

OTHER APPLICATIONS

Most other uses for toroids are variations on the above classes. Toroids used for noise or RFI suppression are covered in the BEADS section.

COATINGS:

Ferrites are hard, abrasive ceramic materials which can abrade wire insulation films during winding. Toroids are ordinarily tumbled so that sharp edges are rounded. However, if a higher level of insulation protection is desired, a smooth as well as an insulating coating can be provided. This coating should be soft to prevent stressing the core upon curing or during temperature cycling, have a low coefficient of friction, withstand normal environments (including cleaning solvents) and provide some additional insulation. For smaller cores, a coating is vapor deposited - a process well suited to bulk coating - and produces an exceptionally uniform coating normally about 0.0006 inches thick. Epoxy is used on larger cores. It is sprayed producing a variable thickness of about 0.001-0.005 inches, and has better physical and chemical properties than other choices. Standard minimum voltage breakdown for both coated cores is 500V. If a higher level of protection is required, please consult with a manufacturer.

Multi-Hole Wide Band Cores

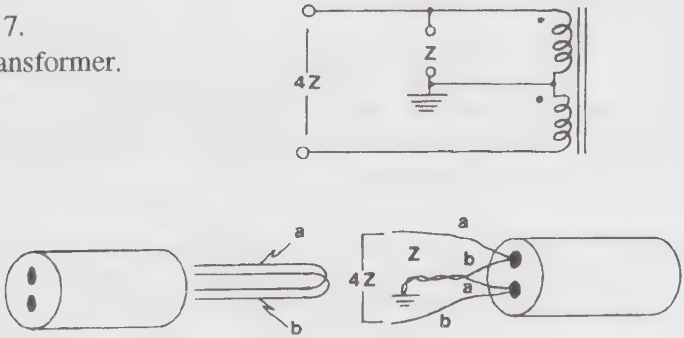
Multi-hole cores provide specialized shapes that are sometimes more useful than single hole devices. One example is wide band transformers where good coupling between short windings is needed over a wide frequency range. Since most of these cores are used in applications of this kind, the standard tests are performed at critical frequencies within their practical operating range. This assures that the cores will operate properly in your application.

You will find curves of X_p, R_p and Z versus frequency for many of these cores. This data was taken with a single winding passing through both holes, the number of turns being selected for convenience in testing. Except for the highest frequencies, where results are controlled by stray reactances rather than the core, characteristics for other numbers of turns may be determined by multiplying by the square of the turns ratio $(N_{new}/ N_{curve})^2$. The effect is to shift the curves up or down by this ratio.

BALUNS

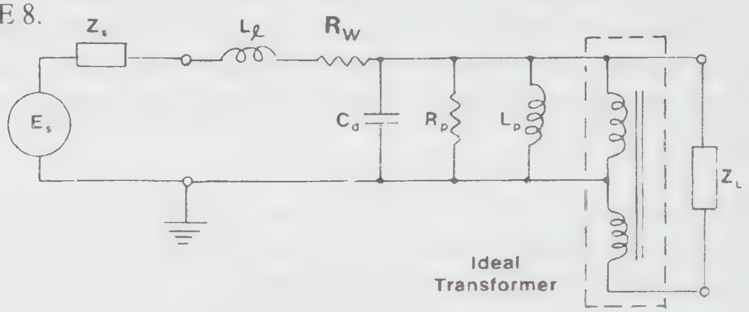
These are transformers used for impedance matching, usually with one side balanced to ground and the other unbalanced (hence the name). They may be wound on any core shape, but often a two hole core is used. Both cylindrical types and "binocular" or "shotgun" types (commonly called "balun cores") are available. Many possible winding arrangements are used, but one simple type is shown in figure 7 . Two U-shaped wires (a, b) are inserted and connected as shown. Then one winding, consisting of a single (two hole) turn, forms the low impedance connection, while two turns in series form the high impedance winding. Since impedance transformation is proportional to the turns ratio squared, it is 4.1. The center-tap may be grounded or left floating.

FIGURE 7.
Balun Transformer.



As with other wide band transformers (see TOROID section) the lower frequency limit is determined by the shunting effect of the reactance produced by the winding inductance as shown in figure 8.

FIGURE 8.



The upper frequency limit is determined by the leakage inductance (L_ℓ) and distributed self capacitance (C_d). Insertion loss is determined by core losses(R_p), winding losses (R_W) and L_p.

TWO-TRANSFORMER DEVICES

There are a number of devices requiring two transformers in conjunction such as wide band cable (CATV, MATV) directional taps, splitters and other hybrids. Although two toroids or beads are sometimes used, a two hole core is often more convenient. Each device is wound through one hole and around the outside. The leads are then in convenient and consistent locations for interconnection. There will be a small amount of coupling between the core halves which should be experimentally examined in your application for its influence on performance.

COMMON MODE CHOKES

Simple noise suppression devices for power lines can be made by passing each side of a wire pair through one hole of a two hole core. At low current levels each half of the core acts as a choke on its own conductor. But at higher currents, when individual beads would saturate, only the web between holes saturates (figure 9).

FIGURE 9.

The power frequency currents in the outer portion cancel so that the outer ring may function as a choke to common mode noise signals. For further information see the BEAD section.

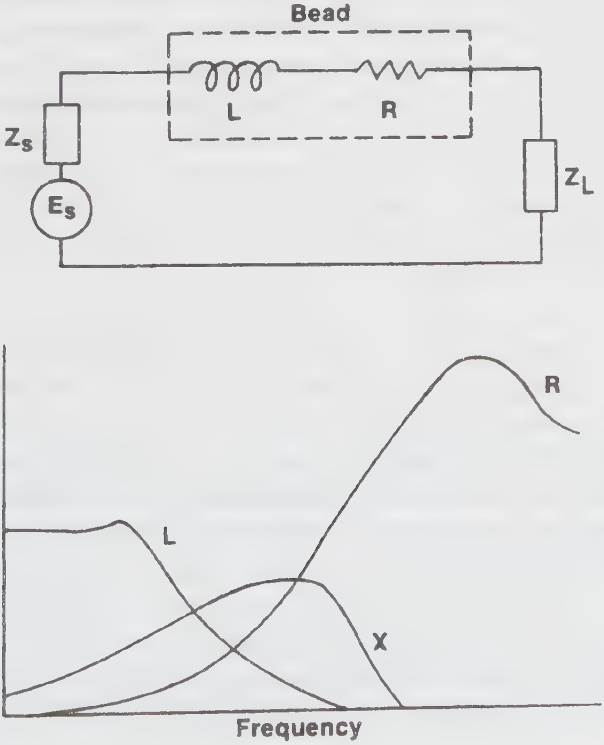
Beads

Ferrite beads provide a simple, economical method for attenuating unwanted high frequency noise or oscillations. By slipping a bead over a wire an RF choke or suppressor is produced which possesses low impedance at low frequencies and relatively high impedance over a wide high frequency band. The effectiveness of this impedance in reducing EMI or RFI depends on the relative magnitudes of the source, suppressor and load impedances. Beads are also available fixed on a wire, taped and reeled for automatic insertion (beads on leads).

HOW THEY WORK

At high frequencies the permeability and losses of ferrites vary with frequency. The permeability declines while the losses rise to a broad peak. The equivalent circuit and curves in figure 10 show how this property can be used as a broadband filter.

FIGURE 10.



Ordinarily beads of ferrite are slipped over a wire producing a one-turn device. To low frequencies the component presents a small inductance whose reactance can often be neglected, while to high frequencies the device presents a higher series resistance with near zero reactance. Since this resistance is a result of material losses it is a true dissipative element. Furthermore, since the reactance is low, there is little chance for resonance with stray capacitance which would spoil the suppression.

DETERMINING IMPEDANCE

In manufacturer's catalogs, curves are presented for some standard parts. These show inductance, resistance and impedance versus frequency for a single straight- through conductor (1 turn). Similar values for other sizes in the same materials can be calculated by the ratio of $Ae/\ell e$ (equation 11) for the two cores.

Equation 11. $$\frac{Ae}{\ell_e} = \frac{2.54\,H \ln OD/ID}{2\pi}$$

Here OD, ID and H are the dimensions in inches of a cylindrical bead. Also, ℓe and Ae (in cm and cm^2) are listed in manufacturer's catalogs for all standards parts. For standard beads manufacturers may also list an impedance for each core. This consists of a measurement near the peak impedance frequency using a single turn of short #20 AWG wire. This makes an excellent incoming QC test, as well as a means for comparing the effectiveness of various core choices.

CHOOSING A BEAD

The best material is the one that gives high impedance or resistance at the noise frequencies and low at the desired signal frequencies. Since the frequency range for high resistance is quite wide-- about two decades--this choice is simple and noncritical. It also is necessary that the impedance presented by the bead at noise frequencies be large enough compared to other circuit impedances to provide the desired attenuation. Frequently the source and load impedance are unknown, but if they are known, insertion loss may be calculated from:

Equation 12. $$IL = 20 \log \frac{Z_S + Z_L}{Z_S + Z_L + Z_{core}} \text{ db}$$

INCREASING SUPPRESSION

Bead impedance is directly proportional to the total height dimension and may be increased either by using longer beads or by stringing more than one. The effect of height on beads may be shown in the curves supplied by the manufacturer. Either method giving the same total height is equivalent. Since the magnetic field is totally contained, it does not matter whether the beads are touching or separated. This approach is valid at all frequencies through VHF, but reliable measurements are difficult at higher frequencies.

Impedance is also proportional to A_e/ℓ_e (equation 11) and this may be used to estimate the parameters for various cores.

Higher impedances can also be obtained by winding the wire through the core more than once. Resistance and inductance are proportional to the number of turns squared. Because of capacitance between turns this technique is most effective at lower frequencies. Also, since a greater length of smaller cross section wire is used, dc resistance will increase. A different approach can be taken at low frequencies where there is significant inductance. The filter can be tuned for maximum attenuation at a specific frequency by simply connecting a resonating capacitor from the output side to ground. Because of the high ac resistance, oscillation is rarely a problem and attenuation is also present at other frequencies.

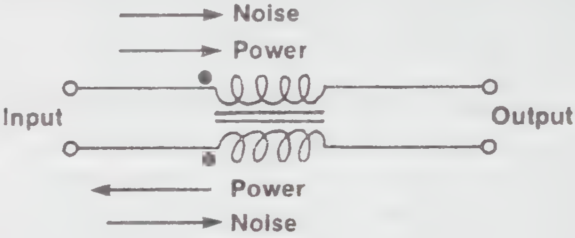
EXCITATION LEVEL

High currents, which are most likely to occur at dc or low frequencies because of the low impedance, can cause significant magnetizing force.

Equation 13. $H = \frac{0.4\,\pi\,N\,I}{\ell_e}$ Oerstreds

This can reduce the impedance and suppression. Since beads are often used with only one turn, fairly high currents can be tolerated before saturation is approached. At saturation, inductance and resistance will be low, but will recover upon removal of the high field. Curves will show the effect of DC current on impedance for certain beads. If the magnetizing force (H) of low frequencies is too great, it will be necessary to increase the effective magnetic path length (ℓ_e). Parts listed in the TOROID section generally have larger ℓ_e for similar A_e/ℓ_e ratios. For further increases in ℓ_e see the discussion on Slotted Toroids in the Toroid section. Another solution to problems concerning low frequency current takes advantage of the fact that much conducted RFI is common-mode. Then it is practical to wind the core as a common-mode choke. The dots in figure 11 indicate the winding sense, that is both windings are put on in the same way (bifilar). Then the magnetic fields of the two windings cancel for normal mode power currents but aid for common mode noise currents. High RF levels can cause excitation greater than that used for the data in this catalog. Often these will increase the effective resistance because of the contribution of hysteresis losses.

FIGURE 11.
Common-Mode
Choke.



ENVIRONMENT

Ferrites are inert ceramics free of any organic substances. They will not be degraded by most environments, including temperatures up to a few hundred degrees centigrade. Magnetic properties vary somewhat with temperature. Generally, inductance increases with increasing temperature while the effect on resistance is small. Above the Curie temperature the bead is nonmagnetic and no suppression can be expected. This effect is completely reversible and once the temperature is reduced below that point, normal performance is regained.

COATING

Because of the high volume resistivity of nickel zinc ferrites, these beads may be considered insulators in most applications. Manganese zinc ferrites are semiconductors and may need to be insulated if they are free to short circuit two or more conductors. Insulating coatings may be applied. This coating should be soft enough to not stress the core upon curing or during temperature cycling, withstand normal environments (including cleaning solvents) and provide insulation. A vapor deposited conformal coating produces an exceptionally uniform coating, normally about 0.0006" thick. Standard minimum voltage breakdown is 500V. A higher level of protection is offered by most manufacturers.

Chokes

Multi-Hole beads and chokes are for situations where a single hole bead does not provide the impedance level desired. An industry standard is the six-hole bead which can be wound to meet design requirements. The standard chokes are offered in 1 1/2, 2 1/2 and 3 turn versions and a common-mode choke version with 2 x 1 1/2 turns.

Manufacturers have also developed a line of ferrite chokes that are specifically engineered for maximum impedance to more effectively suppress high frequency noise or oscillations. These chokes begin where simple beads leave off. They have two to three times the impedance of a standard bead occupying the same space. This is a major benefit when you consider the value of space on a circuit board.

Most chokes are both color coded and marked to facilitate inventory control and production. This feature alone translates into substantial savings.

Special chokes have a unique capability of providing a very flat impedance profile across a broad range of frequencies, from 5 to 500 MHz.

Preassembled chokes take the guesswork out of noise suppression. You no longer need to know the precise frequencies of the noise causing EMI/RFI problems. By using a choke right off the shelf, you can dampen a full range of frequencies necessary to eliminate most noise. To suit your needs even further, manufacturers can adjust the amplitude of impedance to suit many practical requirements.

Slugs

DESIGN CONSIDERATIONS

The majority of slug applications consist of a winding around the core diameter with the magnetic flux parallel to the axis, exiting and entering near the ends. This has two results: The effective magnetic path length is considerably longer than other similar size parts, and the external field is available for useful functions. Slugs are available either pressed to size or, where tight dimensional control is required, with the outside cylindrically ground.

To a great extent the inductance of a coil/slug combination is dependent on the dimensions of the slug, the shape and location of the coil, and, of course, the number of turns, but only slightly on the core material. The material of the slug and its losses at specific frequencies influence the Q or losses of the coil/slug.

Because much of the magnetic path is in air - from one end of the slug around to the other - these devices do not lend themselves to mathematical analysis. The inductance of a coil surrounding a slug is greater than the same air coil by a constant called apparent permeability (μ_a). For a coil covering the whole slug, μ_a depends largely on the length to diameter ratio (l/d) of the slug. When l/d = 7 (about the maximum without resorting to grinding or extrusion) μ_a is about 20 to 40. When the coil is bunched near the center of the slug μ_a also increases- about 1.5 times when the coil is 60% of the slug length, about 3 times when 10%. These guidelines may be used as a starting point for estimating inductance. Also, the magnetizing force of dc current is reduced by the ratio of apparent permeability to material permeability (μ_a/μ). Temperature effects are similarly decreased by μ_a/μ.

POWER TRANSFORMERS AND INDUCTORS

Slugs are useful when the winding must carry large currents, usually dc. This is frequently the case with higher power blocking oscillators or filter inductors for switching power supplies. Using the guidelines above, the design of these devices is very much a matter of experimentation. When there is considerable ac voltage across the coil, it is also important to check the flux density:

$$B = \frac{E}{K A_e N f} \times 10^{-8} \text{ Gauss}$$

Here E is the rms voltage, A_e the slug cross section in cm^2, N the number of turns and f the frequency in Hz. The constant (K) depends on the peak of the voltage integral and is 4.44 for a sine wave, 4.0 for a symmetrical square wave and 1.0 for a unidirectional pulse (then $f = 1/t_w$, the reciprocal of pulse width). This equation will give the peak flux density under the center of the coil, but it is lower elsewhere along the length, so power loss calculations are somewhat difficult.

CHOKES

Slugs make convenient cores for chokes such as those used to limit di/dt and RFI in triac or SCR circuits, to separate RF from power in cable systems, or as stable and accurate RF inductors. Again, the large effective air gap helps to reduce the magnetizing force produced by dc or power frequency currents and the variability effects caused by material permeability tolerances and temperature effects. The specific design will have to be established by trial and error.

TRANSDUCERS

The open magnetic circuit can be used to produce or intercept external fields. A common example is the "loop-stick" antenna where the μ_a of a ferrite rod is used to proportionately shrink the diameter of a loop antenna. The same principle can be used to make transducers for proximity detectors (where a conductive target changes the circuit Q or a ferro-magnetic one changes the inductance), magnetic pick-ups (such as for moving magnets or a magnetic circuit of changing reluctance), to generate localized fields for position detectors, or other kinds of sensors. Sensitivity can be increased by making those changes which raise μ_a - increasing l/d or concentrating the coil near the center.

TUNING CORES

By physically moving a slug into or out of a coil, its inductance can be adjusted. This principle is used in many variable inductors and transformers. Often a stud or wire is cemented into a hole in the slug to facilitate movement. Material selection must often be determined experimentally since the best combination of Q, stability, inductance, cost, etc, is not always apparent from basic material properties.

How to Choose Ferrite Components for EMI Suppression

Introduction

The following pages will focus on soft ferrites used in the application of electromagnetic interference (EMI) suppression. Although the end use is an important issue and some applications are mentioned, this section is not intended to be a design manual, but rather, an aide to the designer in understanding and choosing the optimum ferrite component for the particular application. The reader will discover that ferrite suppressor cores are not too difficult to understand, are very simple to use, in either initial designs or retrofits, and are comparatively economical in both price and space. Ferrite suppressors have been successfully employed for attenuating EMI in computers and related products, switching power supplies, electronic automotive ignition systems, and garage doors openers, to name just a few.

Use of Ferrite Suppressor Cores

The United States was one of the first countries to recognize the potential problems caused by electromagnetic pollution. As a result, the FCC was charged with the responsibility of promulgating rules and regulations to control and enforce limits on high frequency interference.

Current radiation limits are defined by FCC Rules Part 15, for class A (industrial) and class B (mass-market equipment).

Contrary to a decade ago when these regulations were first enforced and designing for EMI protection was often an after thought rather than a forethought, a major portion of today's circuitry is incorporating EMI safeguards in the initial design. Many approaches can be used to comply with design or specification limits for EMI. Attention to basic circuit design, component layout shielded enclosures and other use of shielding materials may be considered. For reducing or eliminating interference on printed circuit boards in wiring and cables, ferrite components have been used very successfully for decades.

There are basically three different ways to use ferrites as suppressors of unwanted signals, conducted or radiated. The first, and least common, is as actual shields where ferrite is used to isolate a conductor, component or circuit, from an environment of radiated stray electromagnetic fields. In both the second and the third application, ferrites are used as components to protect against conducted EMI. The second application is in conjunction with a capacitive element to create a low pass filter that is basically inductance-capacitance (LC) at low frequencies and dissipative at higher frequencies. The third and most common use will be addressed in this section. In this case the cores are used alone on component leads or in board level circuitry either to prevent any parasitic oscillations or to attenuate unwanted signal pickup or transmissions which might travel along component leads or interconnecting wires, traces, or cables.

Variety of EMI Suppression Cores are:

- **Beads on Leads** - Shield beads are supplied assembled on tinned copper wire to aid automated circuit assembly.
- **Split Round Cable Suppression Cores and Cases** - Installed around a cable to attenuate any form of EMI emission. Available for a range of cable diameters. Nylon cases make the assembly of the core halves a snap. Cores are easily installed in equipment where a retrofit proves necessary.
- **Split Flat Cable Suppression Cores and Cases** - Used to attenuate radiated EMI emissions from ribbon cables. These cores can accommodate a range of cable sizes and conductors. Nylon cases and steel clips are available to assist in the assembly of the split cable core halves.
- **Printed Circuit (PC) Beads** - The beads are supplied with tinned copper jumper wires which complete the desired winding configuration on the printed circuit board. Multiple single turn printed circuit beads or multi-turn printed circuit beads are available in various sizes.
- **Toroidal Type Shield Beads** - Used over a wide frequency range.
- **Surface Mount Beads** - These beads are constructed with a solid flat, oxygen free copper conductor with a 30-60 μinch nickel barrier and a minimum 95/5 tin/lead coating thickness of 200 μinch. This rugged constructed decreases dc resistance and increases current carrying capacity compared with plated beads. Surface mount beads are available in various sizes.
- **Multi-Aperture Cores** - Used in balun (balance-unbalance) transformer and as broadband transformers in communication and CATV circuits.

- **Wound Beads** - Multi-hole beads are wound with tinned copper wire in several winding configurations.
- **Connector Suppression Plates** - Used to reduce conducted EMI, "D" type and DIP/Connector suppression plates are available in several sizes and pin layouts.
- **Rods** - Rods are extruded and supplied in long lengths. A diverse range of applications include antennas, impeder cores for HF welding, transducers and transponders. Some rods can be supplied as Slugs (pressed parts).
- **Slugs** - The simplest form or pressed cores used extensively for inductive devices when inductance tolerances of ±10% are permissible. Applications include coils for differential input filters, chokes for SCR and triac circuits, inductors in audio crossover networks and pulse transformers.
- **Discs** - Discs can be used with certain slugs to increase inductance and improve shielding.
- **Tiles** - Tiles are available for partial, semi-anechoic and fully-anechoic chamber applications. Tiles may have center hole for mounting to chamber wall. The tiles are ground on all sides to precise mechanical dimensions and tolerances. These mechanical requirements are necessary to minimize the spacing between adjacent tiles to ensure optimum performance.

The Magnetics

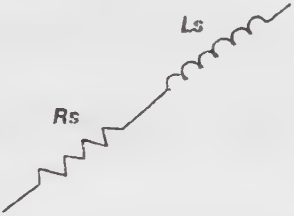
Although permeability and quality factor play a role in the performance of a ferrite EMI suppressor, in virtually all instances, the frequency of usage puts these parameters beyond the point of meaningful definition, and as will be discussed later, the cores impedance is specified instead.

The permeability of a material is a complex parameter consisting of a real and an imaginary part. The real component represents the reactive portion and the imaginary component represents the losses. These may be expressed as series components (μ_s', μ_s'') or parallel components (μ_p', μ_p'').

Manufacturers supply the graphs of the permeability (μ_s') vs. frequency for various ferrite materials. The curves are plots of loss factor ($\tan \delta / \mu$) in parts per million (ppm) vs. frequency. The total loss tangent ($\tan \delta$) is the reciprocal of the Q factor and is a measure of the energy lost or incurred as the magnetization alternates. In almost all cases μ_s' of the material first remains constant with frequency, then rises to a maximum value after which it falls off sharply. The loss component (μ_s'') rises to a peak as μ_s' falls. This is principally due to the ferromagnetic resonance or spin precession resonance. It should be noted that the higher the permeability the lower the frequency at which this occurs. This was first observed by J. L. Snoek.

The impedance of a wound ferrite core is considered to be a series combination of the inductive reactance ($j\omega L_s$) and the equivalent loss resistance (R_s), both of which are frequency dependent. At low frequencies the impedance of the suppressor core is primarily the inductive reactance, which is a function of the material's permeability, and unwanted signals are mostly reflected. As the frequency increases, the inductive reactance decreases. Even so, the total impedance increases due to the increasing losses, and unwanted signals are absorbed. Figure 12 is the simplest representation of the equivalent circuit of a ferrite core, an inductive reactance in series with a resistor.

FIGURE 12.



How to Choose the Right One

The designer or the retrofitter choosing the ferrite for optimum performance in combating EMI, should know the following:

- The frequency of the unwanted signals.
- The source of the EMI.
- Environmental conditions, temperature, and field strengths, AC and DC.
- If high resistivity is required because of multiple turns, conductor pins in a connector filter plate or position in the circuit.
- The circuit source and load impedances.
- How much attenuation is required.
- The allowable space on the board.

The Material

The first task is choosing the best ferrite material. This is based on the frequency or frequency range of the interference that must be suppressed.

Although most manufacturers of ferrite cores can produce more than ten materials, only a select few have been offered for use in the suppression of EMI. The impedance vs. frequency curves for materials are usually shown in manufacturer's catalogs.

Manganese zinc material, which is the material of choice for suppressing signals of 30 MHz or less, is a material of a high permeability 2500µ, low volume resistivity (10^2 ohm-cm).

Nickel zinc material is a "wideband" suppressor material, a medium permeability, 850µ, higher resistivity (10^5 ohm-cm). Material is designed for suppression of signals between 25 and 200 MHz.

A low permeability, 125µ, high resistivity (10^8 ohm-cm) nickel zinc material is recommended for use at frequencies above 200 MHz.

Manufacturers have limited production to several choice materials. Simply stated, why offer eight materials, when three or four will cover the same frequency range effectively?

Covering the broadest frequency range with several materials is economical for both the manufacturer and the consumer, but if the application frequency is specific, rather than over this range, the most effective material may not be one of those offered. The impedance vs. frequency curve for the same bead in different materials should be compared at the proper frequency.

Generally speaking, the higher the permeability, the lower the optimum attenuation frequency, and conversely, the lower the permeability, the higher the attenuation frequency. This is frequency limited at both ends, since low frequency attenuation is reflective and high frequency attenuation is limited by the core and circuit resonance.

The Core

Once the material has been chosen, the size and shape of the core should be selected.

Comparing a set of curves of impedance vs. frequency for different beads, in the same material, will show that increasing the cores volume does not guarantee an increase in impedance.

The same relationship that exists between the low frequency inductance factor (A_L) of a core and the core's initial permeability, can be used to approximate a core's impedance. This relationship is the core's air core inductance or L_o. It can be used as a proportion between two cores of the same material at the same frequency to approximate the impedance of one, knowing the impedance of the other. It is valid only below the resonant frequency, which is a function of the material and geometry of the core and/or winding/circuit conditions. The relationship is:

$Z = K \times L_o$

Toroidal $L_o = 0.046 \times N^2 \times \mathrm{Log}_{10}(OD/ID) \times Ht \times 10^{-8}$ Henries

$Z = 0.046\,K \times N^2 \times \mathrm{Log}_{10}(OD/ID) \times Ht \times 10^{-8}$ Ohm.

All dimensions are in mm.

where:
$K = Z/L_o$ (Ohm/Henry) 10^8
N = Number of turns = 1.

This is a very useful tool for maximizing the impedance of a core within the allowable space, after the best material has been chosen. It reconfirms the fact that in most cases greater impedance will be obtained by increasing the height of the core rather than the diameter for the same increase in volume.

Graphs of the average impedance per unit L_o, and phase angle (ϕ) are used in order to determine the impedance of a core. When using these graphs, the designer needs to know the physical dimensions of the core, the frequency of concern, and the material.

In most instances, maximum impedance is achieved by using the smaller OD with the longer length. If the ID was larger, the reverse would have been true. The same approach may be used for different materials, dimensions, and frequencies.

The previous discussion does assume that the core of choice is cylindrical. If the ferrite core being used is for flat ribbon, or bundled cable, or a multi-hole plate, the calculation for the L_o becomes more difficult, and fairly accurate figures for the cores path length and effective area must be obtained in order to calculate the air core inductance. In all cases though, the increase in impedance will be directly proportional to the height/length as long as the increase in height/length does not cause the core to be in resonance.

The Environment

Ferrite's magnetic parameters can be affected by temperature and field strength. Graphs of impedance vs. temperature may be used to adjust the specified room temperature impedance if desired attenuation is to be at elevated temperatures.

Biases

As in the case of temperature, dc and 50 or 60 Hz power current will also affect the same intrinsic ferrite characteristics which in turn will result in lowering of the impedance of the suppressor core.

A graph shows the effect of dc bias on the initial impedance. The curves depict the degradation in impedance vs. field strength, H in Oersted, for selected frequencies.

In the design example cited earlier, the user chose a certain material ferrite core with a 5 mm outside diameter, 0.8 mm inside diameter and a 10 mm length. Suppose this core will be exposed to 1 Amp dc, and the resulting impedance must be determined.

In order to calculate the field strength, H, the path length of the core must be established. This is a lengthy procedure that involves first calculating the core constants, C_1 and C_2, then using these to find the path length.

These calculations reduce to:
path length $\ell e = 2\pi\,[\log_e (r_2/r_1)] / (1/r_1 - 1/r_2) = 0.55$ cm

and $H = 0.4\,\pi\,NI/\ell e = 2.28$ Oersted

where: $r_1 = \frac{ID}{2} \qquad r_2 = \frac{OD}{2}$

The curves can confirm that the greatest degradation in impedance occurs at the lower frequency.

Resistivity
The dc resistivities of ferrites can range from 10 Ohm-cm to more than 10^9 Ohm-cm. Generally, high permeability manganese zinc ferrites have low resistivities, below 10^2 Ohm-cm, while nickel zinc ferrites have resistivities that range from 10^5 - 10^{12} Ohm-cm. In most instances this parameter is measured using low voltage values, and since a ferrite's resistivity is voltage sensitive, these values may not be valid if the body of the core itself is exposed to high dc or AC values. There are ferrite materials available that have resistivity of 10^9 Ohm-cm measured with 1000 volts. If the frequency that must be attenuated is lower than 25 MHz, requiring a manganese zinc ferrite, coating the core with a special coating, epoxy or polyurethane varnish or using insulated wires, may be the only solution.

Helpful Hints......

Increasing Impedance

Impedance can be increased significantly by adding turns to a bead or a coil on a slug. A graph of impedance vs. frequency for a large toroidal core with one, two and four turns, will show that the impedance will increase in direct proportion to the turns squared, but the frequency at which the maximum impedance is reached is lowered due to the additional capacitive effects.

A graph of different plots of impedance for the same size slug, same number of turns, in different materials, will illustrate that the impedances gained by this type of configuration can be significant. Again there is a tradeoff; increasing the permeability lowers the resonant frequency at which the impedance becomes maximum, therefore narrowing the effective frequency band. The lower the initial permeability of a material, the higher the frequency where this occurs.

Increasing DC Handling

Introducing an airgap in the cores path length can decrease the degrading effects of dc bias. The larger the gap the less effect the bias will have on the impedance. Gaps vary from finishing gaps in mated parts to the gaps found in open magnetic circuits, for example, a slug. Note that the higher the frequency the less the effect of the gap and the current.

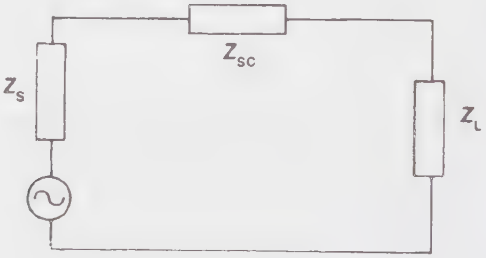
Specifying the Core

Most ferrite manufacturers specify their EMI product line in terms of impedance at select frequencies.

Some manufacturers specifies impedances at two frequencies for each material. The impedance of the cores are controlled to meet a minimum at the lower frequency and at the higher frequency is toleranced. The measuring equipment is specified by the manufacturer.

Figure 13 shows the equivalent circuit of an interference source with an internal impedance of Z_S, generating an interference signal through the series impedance of the suppressor core Z_{SC} into the load impedance Z_L.

FIGURE 13.



Most manufacturers are specifying cores for EMI applications in terms of impedance, but often the end user needs to know the attenuation. The relationship that exists between these two parameters is:

Attenuation = $20 \log_{10} Z_s + Z_{sc} + Z_L / Z_s + Z_L$ dB

where:
Z_s = source impedance
Z_{sc} = suppressor core impedance
Z_L = load impedance.

This relationship is dependent on the impedance of the source generating the noise and the impedance of the load receiving it. These values are usually complex numbers that can be infinite in scope and not easily obtained by the designer therefore it is recommended that the attenuation calculations be verified in the actual circuit by appropriate measurements

Selecting a value of one ohm for both the load and the source impedance, as may be the case when the source is a switched-mode power supply and the load has many low impedance circuits, simplifies the equation and allows comparison of ferrite cores in term of attenuation

Under these conditions, the equation reduces to:

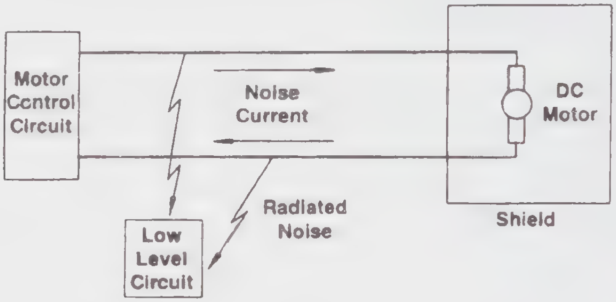
$A = 20 \operatorname{Log}_{10} |Z_{sc}|/2$ dB

$Z_{sc} >> 1$ Ohm

Some Examples

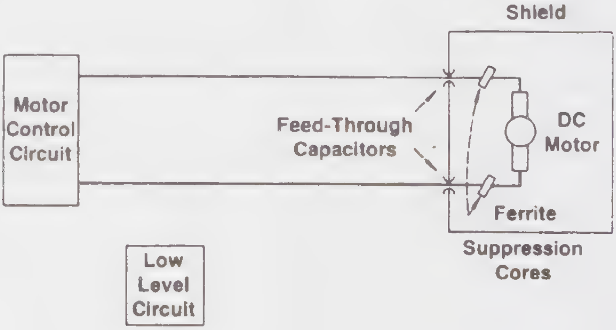
In the circuit shown in Figure 14, a metal enclosure is used to eliminate radiated interference generated by the commutation of the dc motor. This enclosure, however, will not prevent the conduction of noise on the power cables of the dc motor. This conducted noise in turn might affect other circuits with either conducted or radiated interference.

FIGURE 14. - Commutation noise is interfering with low-level circuits.



When a metal enclosure is used in conjuction with ferrite suppression cores and feed-through capacitors, as shown in Figure 15, the conducted noise is eliminated from the circuit.

FIGURE 15. - Ferrite suppression cores used in conjuction with feed-through capacitors to eliminate interference.



In applications that must accommodate dc, ferrite materials are influenced by the level of dc magnetization. This might require the use of a suppressor core with an airgap in its magnetic circuit.

In Figure 16, a ferrite core is used to suppress common-mode noise, effectively reducing ground-loop coupling without affecting the circuit load current.

FIGURE 16. - Ferrite core used for suppressing common-mode noise.
I_{cm} = Common-mode current
Z_W = Impedance of wire

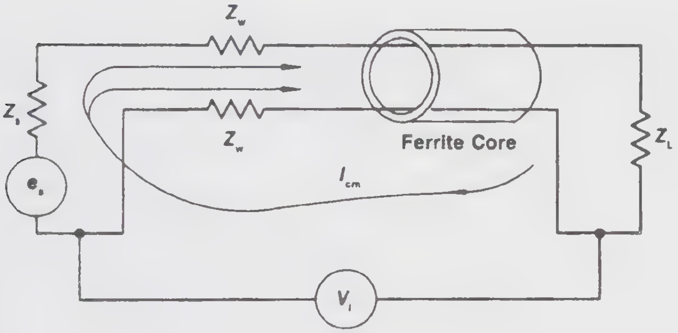
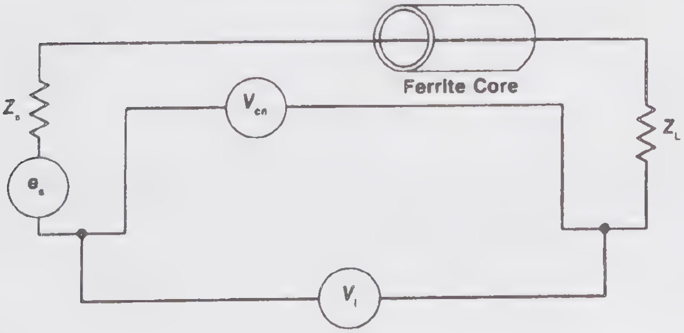


Figure 17 illustrates the use of a ferrite core to suppress conducted noise, without introducing significant circuit losses at low frequencies.

FIGURE 17. - Ferrite core used to suppress conducted noise.
Z_L = Load Impedance
V_I = Common-mode voltage
V_{cn} = Conducted noise voltage



The Use of Ferrite as an Absorber in Anechoic Chambers

Introduction

As a consequence of the continuing growth of high speed electronic systems and equipment in the marketplace and the associated problems of electromagnetic interference (EMI), recent national and international regulatory acts have been instituted. Worldwide organizations have been established, FCC (U.S.), VCCI (Japan), VDE (Germany), and CISPR (international), for the purpose of imposing maximum acceptable levels of noise that can be radiated from electronic systems and equipment.

Testing to ensure compliance to these specifications is being performed over the frequency range of 30 to 1000 MHz by laboratories all over the world, either in open field test sites or inside anechoic chambers. Open test sites can be effective, but since they are at the mercy of the environment, both natural and manmade, the testing can be time-consuming and expensive. Another drawback is that suitable sites are difficult to find and are usually located in remote areas far removed from manufacturing facilities, where the need for quick testing and solutions to EMI problems may be crucial. Although the development of indoor anechoic chambers eliminated most of the problems that are associated with open field test facilities, they do have their own obstacles. To simulate an open test site, designers of these chambers must consider controlling reflection within the chamber and shielding the chamber from transmitting or receiving radio wave interference. There is a variety of these chambers, depending on application, but most are using (or investigating the use of) ferrite tiles as an absorber material.

Basic Theory

When an electromagnetic wave traveling in free space encounters a different medium, the wave is either reflected, transmitted/refracted and/or absorbed. In this application, which is a lined metallic enclosure, when excited by the source, the field that will be measured by the antenna will depend not only on the energy reflected from the absorber and the scattering effects, but also the energy that is transmitted through the absorber, then reflected back by the metal sheet through the absorber again along with all the resultant scattering effects.

In typical absorbing structures the dissipative material is arranged in the form of wedges, pyramidal cones, ribs, foils, thin layers, or homogeneously distributed within a thick layer. There is a constant search to develop the best material, or combination of materials, so that when placed in front of the metal surface, absorption of incident electromagnetic energy is maximized over the widest frequency range. Unfortunately, almost all these dissipative materials produce reflections at every plane surface. So, when designing anechoic EMC test facilities one of the main concerns is to reduce reflections from the shielded room walls, particularly at the lower frequencies, down to 30 MHz. Thus, the parameter of major concern is the reflectivity of the composite "metallic surface + absorber system."

It was researched and shown over 25 years ago that the application of ferrites for homogeneous single layer absorbers is effective for the frequencies of about 20 to 2000 MHz; many ferrites will act as thin magnetic absorbers in the same frequency range with layer thickness of 5 mm to 6.5 mm.

The factors that determine whether a ferrite tile will be a good absorber are its shape, including thickness, and design, along with its intrinsic magnetic parameters. The determining magnetic parameters are the relative complex permeability, and permittivity.

Over the stated frequency range the relative permittivity of ferrite is almost constant, while the complex permeability exhibits dispersion. When designing chambers using ferrite tiles, the goal becomes finding the best combination of size, shape and material characteristics providing the closest impedance match over the broadest frequency range, with the least reflection.

Measuring Techniques

Although there are several methods of measuring the effectiveness of shielding and/or absorbing materials, the most reliable is testing the end product in the final configuration. Unfortunately, this is not always a feasible design approach.

In most cases, the source of the EMI determines the best measuring technique. Conducted EMI is interference that is coupled between circuits, equipment or systems as a result of being conducted along an interconnecting power or signal wire or cable. The units of measure for conducted EMI are usually expressed in terms of voltage or current-lumped elements.

Previous articles written by manufacturers detailing the use of ferrite components for the suppression of EMI, have taken this circuit theory approach. The graphs and specifications are usually given in terms of impedance and phase angle.

Radiated EMI is interference that is coupled between circuits, equipment or systems by means of electromagnetic fields that are radiated from an EMI source and picked up by susceptible circuits, equipment or systems. The units of measure for radiated EMI are expressed in terms of power density or field strength. Measurement of the intrinsic parameters are made on toroids that are precision cut from randomly selected ferrite absorber tile and then closely fitted into a coaxial airline fixture. Utilizing this sampling technique, an S-parameter test set and a Network Analyzer, the incident signal is compared with transmitted signal, data is collected and the S-parameters are used to compute the complex parameters for the graphs of complex permeability and complex permittivity vs. frequency.

The Effect of Direct Current on the Inductance of a Ferrite Core

Introduction

If ferrite cores are used in the design of transformers, chokes or filters, which are required to carry direct current, it is necessary to predict the degree of inductance degradation caused by the static field. When dc flows through the winding of a ferromagnetic device, it tends to pre-magnetize the core and reduce its inductance. The permeability of a ferrite material measured with superimposed dc might increase slightly for very low values of dc ampere-turns, but then it progressively decreases as the dc field is increased and the core approaches saturation. This permeability is referred to as the incremental permeability, μ_Δ. If an air gap is introduced into the magnetic path of a core, the reluctance is increased hence the inductance is decreased. However, the core's capacity for dc ampere-turns without a degradation in inductance is significantly improved, albeit at the expense of a lower effective permeability.

DC Bias in Gapped Cores

The use of graphs such as the Hanna curves has simplified the tedious trial and error methods are often employed when designing inductors with superimposed dc. A Hanna curve is created by measuring the inductance vs. dc bias of various core sizes and gap lengths of the same material grade. The measured data is used to create curves such for a set of E cores. A line is drawn connecting the individual curves through the point of tangency. The graphs are then normalized by dividing the vertical scale by the effective core volume, V_e, and the horizontal scale and the gap lengths by the effective path length, ℓ_e, of the core set. The individual curves, once normalized, overlay creating the Hanna curve.

For a typical output choke application, the designer knows a number of design criteria such as the required inductance, the direct current, alternating ripple current and allowable dc resistance. He will also have requirements for core size, ambient temperature an often a preference for a particular core geometry.

DC Bias in Open Magnetic Cores

The discussion so far has been on core types that have a closed magnetic path, in which a small air gap has been inserted by either a ground gap or the use of shims. An open magnetic core can be thought of as a core with a very large fixed air gap. Since the air gap is determined by the core geometry and cannot be changed, the Hanna curves can not be used for these types of cores. Such cores as rods, slugs and bobbins can be used quite successfully in inductor designs that have relative low inductance values and can accommodate significant amounts of static currents.

The large air gap will forestall the saturation of this type of core, hence the inductance will not as easily decrease as a function of the dc ampere-turns. For an inductor design the number of turns can be calculated from the required inductance, L, and the inductance index of the bobbin. $N = \sqrt{L/A_L}$ (L in nH). The turns, N, times the direct current, I, will give the NI product, which should be less than the value quoted for the bobbin. For winding fit and dc resistance check, the same procedure is used as outlined in the example above, except here the W_a of the bobbin is the total available winding area. The graphs can show the effect of temperature on the inductance factor vs. dc bias characteristics of the bobbin. The NI values in manufacturer's catalogs are at room temperature, and must be derated when operating at elevated temperatures. Open magnetic cores, rods, slugs and bobbins are used and designed into SCR and triac controls, speaker crossover networks and differential-mode input filters. They are also utilized for EMI suppression applications where relative large direct currents are present and also for output chokes in switched-mode power supplies.

Use of Ferrites in Broadband Transformers

Introduction

Most of the magnetic information in this chapter is data obtained from cores wound with a single multi-turn winding which forms an inductor. When a second winding is added on the core, the inductor becomes a transformer. Depending on the requirements, transformers can be designed to provide dc isolation, impedance matching and specific current or voltage ratios. Transformers designed for power, broadband, pulse, or impedance matching can often be used over a broad frequency spectrum.

In many transformer designs ferrites are used as the core material. This section will address the properties of ferrite materials and core geometries which are of concern in the design of low power broadband transformers.

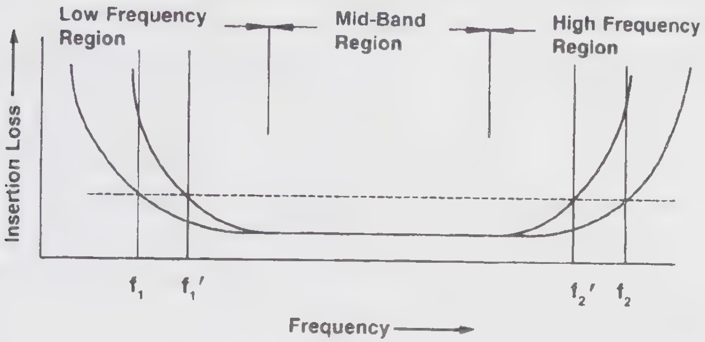
Brief Theory

Broadband transformers are wound magnetic devices that are designed to transfer energy over a wide frequency range. Most applications for broadband transformers are in telecommunication equipment where they are extensively used at low power levels.

Figure 18 shows a typical performance curve of insertion loss as a function of frequency for a broadband transformer. The bandwidth of a broadband transformer is the frequency difference between f_2 and f_1, or between f_2' and f_1', and is a function of the specified insertion loss and the transformer roll-off characteristics.

It can be seen that the bandwidth is narrower for transformers with a steep roll-off (f_2'-f_1') than for those with a more gradual roll-off (f_2-f_1). Also in figure 18, the three frequency regions are identified.

FIGURE 18. - Typical Characteristic Curve of Insertion Loss vs. Frequency for a broadband transformer.



The cutoff frequencies are determined by the requirements of the individual broadband transformer design. Therefore, f_1 can be greater than 10 MHz or less than 300 Hz. Bandwidths also can vary from a few hundred hertz to hundreds of MHz. A typical broadband transformer design will specify for the mid frequency range a maximum insertion loss and for the cutoff frequencies, f_1 and f_2 maximum allowable losses. A typical schematic diagram shows the lumped element equivalent circuit of a transformer, separating the circuit into an ideal transformer, its components and equivalent parasitic resistances and reactances. The secondary components, parasitics and the load resistance have been transferred to the primary side and are identified with a prime.

To simplify this circuit, the primary and secondary circuit elements have been combined. In the low frequency region the roll-off in transmission characteristic is due a lowering of the shunt impedance. The shunt impedance decreases when the frequency is reduced, which results in the increased level of attenuation. The impedance is mainly a function of the primary reactance X_{LP}, with a negligible contribution of the equivalent shunt loss resistance R_p. The insertion loss may therefore be expressed in terms of the shunt inductance.

For most ferrite broadband transformer designs, the only elements that are likely to effect the transmission at the mid-band frequency range are the winding resistances.

In the higher frequency region the transmission characteristics are mainly a function of the leakage inductance or the shunt capacitance. It is often necessary to consider the effect of both of these reactances, depending upon the circuit impedance.

Reviewing the insertion loss characteristics for the three frequency regions, it can be concluded that the selection of ferrite material and core shape should result in a transformer design that yields the highest inductance per turn at the low frequency cutoff f_1. This will result in the required shunt inductance for the low frequency region with the least number of turns. The low number of turns are desirable for low insertion loss at the mid-band region and also for low winding parasitics needed for good response at the high frequency cutoff f_2.

Low and Medium Frequency Broadband Transformers

For broadband transformer applications the optimum ferrite is the material that has the highest initial permeability at the lower cutoff frequency f_1. Manganese zinc ferrites are very suitable for low and medium frequency broad- band transformer designs. As stated before, the transformer parameter that is most critical is the shunt reactance (ωL), which will increase with frequency as long as the material permeability is constant or diminishing at a rate less than the increase in frequency. This holds true even if a transformer is designed using a manganese zinc ferrite where f_1 is at the higher end of the flat portion of the permeability vs. frequency curve. Although the whole bandpass lies in the area where the initial permeability is decreasing, yet the bandpass characteristics will be virtually unaffected. For broadband transformers that use a manganese zinc ferrite material the core geometry should be such as to minimize the R_{dc}/L ratio. In other words, the ratio of dc resistance to the inductance for a single turn should be a minimum. The range of pot cores, standardized by the International Electrotechnical Commission in document IEC 133, has been designed for this minimum R_{dc}/L ratio. Other core shapes such as the EP cores and PQ cores can also be used in the design of these broadband transformers. Often the final core selection will also be influenced by such considerations as ease of winding, terminating, and other mechanical design constraints of the transformer.

Broadband Transformers with a Superimposed Static Field

In transformer designs that have a superimposed direct current, gapped cores can be employed to overcome the decrease in the shunt inductance. Hanna curves can be used to aid in the design of inductive devices that carry a direct current. For more information see section "The Effect of Direct Current on the Inductance of a Ferrite Core."

High Frequency Broadband Transformers.

Although there is no clear division between the frequency regions, for this section it is assumed that the high frequency broadband transformer designs use nickel zinc ferrites as the preferred core material. This will typically occur for transformer designs where the bandpass lies wholly above 500 kHz. At these higher operating frequencies it becomes more important to consider the complex magnetic parameters of the core material, rather than use the simple core constants, such as A_L, recommended for low frequency designs.

Another important consideration is that high frequency transformers are generally used in low impedance circuits, which means that these designs require low shunt impedances. This can often be accomplished with a few turns, hence winding resistances are no longer an issue, and the design concept of minimizing R_{dc}/L is no longer required. The design will instead become focused on core shape and material for the required shunt impedance at f_1 along with reducing leakage inductance of the winding. Since the material characteristics permeability and losses affect the shunt impedance these parameters need to be considered in high frequency broadband transformer designs. For high frequency broadband transformers the toroidal core shape becomes an attractive core geometry. The few turns that are often required can easily be wound on the toroid. However, windings that require only a few turns may give rise to problems in obtaining the desired impedance ratios. To minimize leakage inductance it is suggested that the primary and secondary windings be tightly coupled and where possible a bifilar winding be used.

An improvement in core performance over toroids can be obtained by the use of multi-aperture cores, which can be considered as two toroidal cores side by side. This core shape has a lower single turn winding length than the equivalent toroidal core with the same core constant C_1, and will result in a wider bandwidth of the transformer design.

Many broadband transformers have been designed utilizing nickel zinc ferrite toroids with good results. If bandwidth requirements cannot be met using toroids, multi-aperture nickel zinc cores should be considered.

Summary

The low cutoff frequency f_1 is the single most important factor in the ferrite material selection. The material with the highest initial permeability at f_1 is the recommended choice.

Manganese zinc ferrites, can be used to a cutoff frequency f_1 of 500 kHz. Above this frequency use a nickel zinc ferrite, again depending upon the frequency.

For low and medium frequency transformers the optimum core shape should provide the lowest dc resistance per unit of inductance. If there is a superimposed dc present the use of gapped cores and Hanna curves is suggested. For high frequency designs, use nickel zinc ferrite. The toroidal and multi-aperture cores are the recommended core configurations.

The number of turns should be kept to a minimum to reduce leakage inductance and self-capacitance of the windings. Wind primary and secondary windings tightly coupled or as bifilar windings to lower leakage inductance.

Application Summary

Ferrites can be used in an ever widening range of electronic component applications as inductors, chokes, and transformers in televisions, CATV taps and splitters, power supplies, computers, and computer peripheral devices. Some of the more common applications include the following:

MAGNETIC DEVICES

- Power transformer and chokes
- Inductors and tuned transformers
- Pulse and wide band transformers
- Magnetic deflection
- Recording heads
- Rotating transformers
- Shield beads and chokes
- Transducers

APPLICATIONS

- HF Power supplies
- Frequency selective circuits
- Matching devices
- TV sets and monitors
- Storage devices
- VCR's
- Interference suppression
- Vending machines

Ferrites can be pressed in block form and then machined into intricate shapes. Where large sizes are required, it is possible to assemble them from two or more smaller machined or pressed sections; the variety of sizes and shapes becomes limitless.

Four major variables that influence performance of magnetic components are:

- FREQUENCY
- TEMPERATURE
- GEOMETRY
- STABILITY

These reference charts will help the designer select the magnetic component that precisely meets the designers needs. Cores and laminations offers freedom of choice.

Selection, of course, depends on the specific application involved. Tape wound and bobbin cores are generally used in square loop applications such as inverters or magnetic amplifiers. Permalloy powder cores and ferrites, having linear characteristics, are primarily used in chokes, coils, inductors, filters, resonant circuits, and transformers; there are certain varieties of tape cores that could be used in these applications.

FREQUENCY RANGE

MAGNETIC MATERIAL	MATERIAL SPECIFICATIONS	FREQUENCY RANGE
Transformer Laminations (Nickel - Iron)	$\mu_o = 5,000 - 60,000$ $Bm = 8,000 - 15,000$	1 KHz to 10 KHz
Tape Wound Cores (Nickel - Iron)	$\mu_o = 5,000 - 100,000$ $Bm = 8,000 - 20,000$	1 KHz to 100 KHz
Cut Cores (Nickel - Iron)	$\mu_o = 1,000 - 25,000$ $Bm = 8,000 - 20,000$	1 KHz to 100 KHz
Tape Wound Cores (Amorphous)	$\mu_o = 3,000 - 20,000$ $Bm = 5,000 - 16,000$	1 KHz to 1 MHz
Bobbin Cores (Nickel - Iron)	$\mu_o = 5,000 - 100,000$ $Bm = 8,000 - 15,000$	1 KHz to 1 MHz
Permalloy Power Cores (80% Ni)	$\mu_o = 14 - 550$ $Bm = 7,000$	1 KHz to ≈ 2 MHz
High Flux Power Cores (50% Ni - 50% Fe)	$\mu_o = 14 - 160$ $Bm = 14,000$	1 KHz to 2 MHz
Ferrites (MnZn)	$\mu_o = 750 - 15,000$ $Bm = 3,500 - 5,000$	10 KHz to 2 MHz

TEMPERATURE RANGE (°C)

MAGNETIC MATERIAL	TEMPERATURE RANGE (°C)	NOTES
Laminations	- 60 to 200	
Tape Wound Cores	- 60 to 200	
Bobbin Cores	- 60 to 125	
Permalloy Powder Cores (μ_o)		Anticipated Induction Changes (Δ L)
14 - 550	- 60 to 140	± 1.5%
60 - 200	0 to 55	± 0.1%
60 - 200	- 60 to 135	± 0.25%
60 - 300	- 55 to 75	ΔF = < 0.05% with polystyrene capacitor
Ferrite Cores (μ_o) (Filter Materials)		
750	- 30 to 80	1.0 to 3.0 **
2,000	- 30 to 80	0.9 to2.1 **
2,300	20 to 80	- 0.7 to + 0.7 **
(Power Materials)		
1,500	- 30 to 150	
2,300	- 30 to 150	
2,500 *	- 30 to 150	
3,000 *	- 30 to 150	
5,000	- 30 to 110	
10,000	- 30 to 110	
15,000	- 30 to 100	

* Bm = 4900

** T.F. = $\frac{\Delta \mu_o}{\Delta T \mu_o^2}$

STABILITY CRITERIA

MAGNETIC MATERIAL	STABILITY
TAPE WOUND CORES and BOBBIN CORES	• Will withstand high shock and vibration.
PERMALLOY POWDER CORES	• Excellent DC bias stability. • Excellent AC flux density stability. • Good frequency stability, (Q values up to 250). • Narrow inductance tolerances, (± 8% in 2% groups). • Superior temperature stability.
FERRITES (MnZn)	• Narrow inductance tolerances, (± 3% in gapped pot cores). • Excellent time stability, (Disaccommodation factors as low as 1.5×10^{-6}). • Very good frequency stability. • Good DC bias stability.
LAMINATIONS	• μ_o stability depends on materials and shape. • Not as stable as powder cores.
NICKEL-IRON CUT CORES	• Excellent DC bias stability. • Will withstand high shock and vibration.

FERRITE SHAPES

APPLICATIONS	DESIRED PROPERTIES	AVAILABLE SHAPES
FILTER INDUCTORS	• High μQ. • High stability, adjustable and fixed.	Pot cores, toroids, E, U, and I cores, RM cores.
NARROW BAND TRANSFORMERS	• Moderate Q. • High μ. • High stability.	Pot cores, toroids.
POWER TRANSFORMERS AND INDUCTORS	• High μ. • Low losses at high flux densities and temperatures. • High saturation.	Ungapped pot cores, E, U, and I cores, toroids, EP cores, RS cores, PQ cores, beads, slugs.
BROAD BAND TRANSFORMERS	• Low loss. • High μ.	Pot cores, toroids, E, U, and I cores, RM cores, EP cores.
PULSE TRANSFORMERS	• High μ. • Low loss. • High Et product.	Toroids, beads, baluns, multi-holes.
CONVERTER AND INVERTER TRANSFORMERS	• Low losses. • High saturation.	Toroids, E, U, and I cores, pot cores, RS cores.
NOISE FILTERS	• Very high μ.	Toroids, beads, baluns, multi-hole.
MACHING APPLICATIONS	• High μ. • Low losses. • High saturation.	Ferrite blocks for machined parts.
HIGH Q INDUCTORS	• High Q.	Toroids
LINE FILTERS	• Very high μ.	Toroids
SPECIAL APPLICATIONS	• Controlled temperature properties.	Toroids

Application Considerations - Ferrite Advantages and Disadvantages

Application	Advantages	Disadvantages
Low Frequency **(<1 KHz)** **High Flux Applications** Generators Motors Power Transformers	• Ease of forming shapes allows possible use in inexpensive, high loss applications such as relays, small motors.	• Flux density low. • Relative cost high. • Limited size of parts.
Medium Frequency **(1-100KHz)** **High Flux,** **Non-Linear, Applications** Flyback Transformers Deflection Yokes Inverters Wide Band Transformers Recording Heads Pulse Transformers Memory Cores	• Cost much lower than Nickel-Iron alloys, especially thin tapes. • Moderately high permeabilities available. • Low losses, especially in upper half of this range. • Inherent shielding in pot cores. • Good wear resistance. • Easily adapted to mass production.	• Flux density lower than Nickel-Iron alloys. • Permeabilities lower than Nickel-Iron alloys. • Curie Temperature fairly low. • Good mating surface necessary for high inductance. • Smaller flux change than bobbin cores.
Medium Frequency **(1-100 KHz)** **Low Flux,** **Linear Applications** Loading Coils Filter Cores Tuned Inductors Wide Band Transformers Antenna Rods	• Permeabilities higher than powdered iron or Permalloy cores. • Gapped pot cores provide: 1. Adjustability. 2. Self-shielding. 3. Stability - temperature, time, AC flux density, and DC bias. • Q Products higher than other materials. • Wide choice of Inductance and Temperature Coefficient.	• Brittleness. • Low Curie point. • Need precision grinding of air gap. • Mounting hardware needed.
Higher Frequencies **(>200 KHz)** **Low Flux,** **Linear Applications** Filters Inductors Tuning Slugs	• Low losses, (especially eddy currents). • Ferrites, powdered iron and moly-permalloy can operate at higher frequencies. • Medium frequency advantages apply.	• Permeability decreases with frequency. • Medium frequency disadvantages apply. • Poor heat transfer.
Microwave **Frequencies** **(>500 MHz)**	• Low dielectric losses. • Good gyromagnetic properties. • Only bulk materials available.	

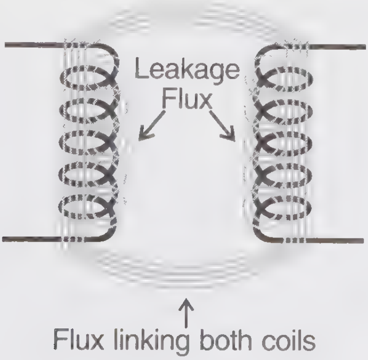
Chapter 3

Transformers

Signal Transformer Specification and Application

A transformer in its simplest sense is two or more windings coupled by a common magnetic field (flux density). It is this magnetic field that provides the means to pass voltages and currents from one winding (primary) to the other winding (secondary). Magnetic flux is made when an alternating current flows through a winding (see Figure 1). This magnetic field (represented by magnetic flux lines) follows the path of least magnetic resistance (reluctance).

FIGURE 1.



The path that the flux follows then proceeds through to the other winding where a current is induced. However, not all of the flux created by the first winding flows through the second winding. The flux that does not connect the two windings is called leakage flux and is energy lost. There are other forms of energy losses such as core and winding losses which also reduce the efficiency of the transformer but will not be discussed in detail here. One way of reducing the amount of leakage flux is to shorten the path between the two windings or place the windings on top of each other. Properly configuring the winding allows more of the flux to link both windings and limit losses. Lowering leakage flux (leakage inductance) is a primary goal of the signal transformer designer since fewer losses means that a greater amount of the signal will be transferred in an undistorted manner. There are many reasons for using a transformer including: isolation of DC currents, voltage and current transformations, and impedance matching. The type of signals to be transferred from the primary to secondary winding(s) relate directly to the type of transformer that should be used in an application. For instance, a transformer needed to provide DC isolation between two windings carrying large amounts of currents would be designed differently than a transformer that needs to provide an impedance match to a small signal communications network. In this section, the emphasis will be on two types of signal transformers that are designed specifically for the transmission of data at low power levels with DC isolation (the wide band transformer) and those designed to be non-isolating (the autotransformer). These types of transformers are not limited to low power capabilities, but the emphasis here is for use in communication and small signal applications.

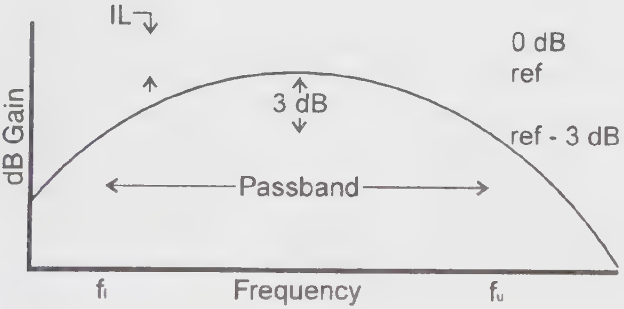
The Wide Band Transformer

Many transformers are efficient in their ability to carry large amounts of voltage and current over a narrow range of frequencies. However, these transformers can become inefficient in terms of transferring energy that is free of distortion. A wide band transformer can transmit a clean, undistorted low power signal over a wide range of frequencies. This is important when transmitting a signal such as a square wave since each individual pulse is made up of a wide range of harmonics. These harmonics are what give the square wave its shape and are thus necessary to preserving the original wave shape.

A wide band transformer is used for many different applications including: impedance matching, phase shifting, isolation, coupling, balanced to unbalanced transitions, and current and voltage conversion, while still providing DC isolation between circuits. The major emphasis on a wide band transformer is usually placed on the ability to transmit small signal information over a wide range of frequencies. Wide bandwidths are possible through tight coupling between the primary and secondary winding(s). Due to tight coupling, the frequency range that can be achieved for efficient operation is much higher than a typical power transformer.

When speaking of the operable range of frequencies for a wide band transformer, one usually refers to the transformer's bandwidth. The bandwidth is the range of frequencies that are allowed to pass while still maintaining at least 1/2 of the signal power. The 1/2 power point is usually measured in dB and is specified as being 3 dB down from the insertion loss (see Figure 2).

FIGURE 2.

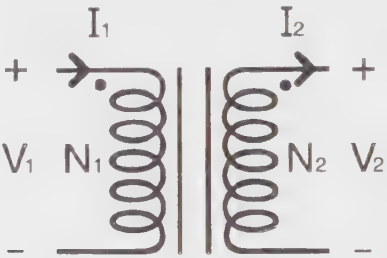


Other parameters also need to be considered when evaluating a wide band transformer. Depending on the application, the maximum or minimum inductance on the primary or secondary winding, DC resistance of the windings, rise time, or a number of other factors might be considered when designing or specifying a wide band transformer. One primary parameter that needs to be determined is the turns or inductance ratio. The ratio is used for matching the primary and secondary impedance of the transformer to the circuit into which it is inserted. This ratio is also crucial for "stepping up" or "stepping down" a voltage or current from one side of the transformer to the other.

Design Equations

As mentioned earlier, one of the main uses of the wide band transformer is to step-up or step-down voltages or currents. The same techniques can be applied to matching impedances between unbalanced circuits. For the purpose of design simplicity, a lossless transformer will be used to derive the equations (see Figure 3).

FIGURE 3.



For a lossless transformer, the following equations are true:

$$\frac{N_1}{N_2} = \frac{V_1}{V_2} = \frac{I_2}{I_1} = a$$

Where:
N_1 = number of turns on primary
N_2 = number of turns on secondary
V_1 = voltage across primary
V_2 = voltage across secondary
I_1 = current through primary
I_2 = current through secondary
a = turns ratio.

The inductance of a winding can be approximated by using the following relation:

$$L = N^2 A_L$$

where A_L is the equivalent inductance (per turns squared) of the core material. This is generally specified by the core manufacturer.

To find the required turns ratio necessary to match two impedances Z_1 and Z_2, the following relations are used (note that the impedance is proportional to the square of the turns):

$$Z_1 = \frac{V_1}{I_1} = \frac{V_2}{I_2}(a^2) = Z_2\,(a^2) \quad \text{where} \quad \left(\frac{N_1}{N_2}\right)^2 = a^2$$

For example, if a 100 Ohm source Z_1 is to be connected to a 150 Ohm load Z_2, the resulting turns ratio necessary would be:

$$100 = 150a^2 \quad \sqrt{\frac{100}{150}} = a \qquad a \equiv 0.816 \qquad \text{thus } N_1 = 0.816N_2$$

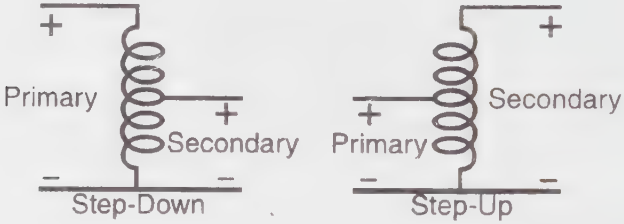
The Autotransformer

Another type of transformer that is applicable in communication systems is the autotransformer. The autotransformer has its own unique characteristics as well as having many of the benefits of the wide band transformer. It can be used in many of the same applications as the wide band transformer such as voltage dividers, impedance matching, phase shifting, and voltage and current transformations. It cannot, however, provide any DC isolation between the primary and secondary(s). What the autotransformer lacks in DC isolation, it gains in efficiency, lower leakage inductance, and better voltage regulation.

The autotransformer is ideal when DC isolation is not necessary and optimum efficiency is needed. Additional efficiency is gained by way of greater coupling, which in turn leads to less leakage inductance. This is partially due to the autotransformer passing part of the energy from primary to secondary by voltage division. Whereas the DC isolating wide band transformer delivers all of its energy by transformer action, the autotransformer delivers part of the energy directly through the winding and the other part through means of a magnetic medium. This allows for better coupling between the primary and secondary and thus less leakage inductance and increased bandwidth.

An autotransformer also requires less windings to achieve the same result as an isolation transformer. This is due to the secondary using part of the primary winding to achieve the necessary turn count for a step-up or step-down transformer (see Figure 4). Since fewer turns are needed to achieve the same primary to secondary impedance ratio, total copper loss will be reduced - thus it is more efficient than a comparable wide band transformer. Note that the autotransformer is not limited to a single secondary (or "tap"), but may have multiple taps off of the primary winding.

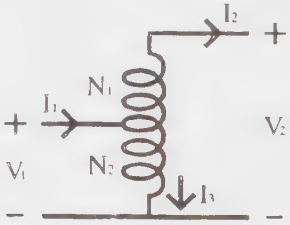
FIGURE 4: Autotransformer (Step-down, step-up).



Design Equations

The techniques used in determining the appropriate number of turns to step-up or step-down a voltage, or match impedances is similar to that of the isolation wide band transformer.

FIGURE 5.



For an autotransformer (see Figure 5), the following equations should be used:

$$\frac{V_1}{V_2} = \frac{N_1 + N_2}{N_2} = \frac{I_2}{I_1}$$

Where:
N_1 = number of turns common to secondary
N_2 = number of turns common to primary and secondary
V_1 = voltage across primary (common winding)
V_2 = voltage across secondary
I_1 = current in tap
I_2 = current in secondary
a = turns ratio.

Again, the inductance of a winding can be approximated by using the following relation:

$$L = N^2 A_L$$

To use the autotransformer as a voltage divider network, use the following equation:

$$V_1 \left(\frac{N_1 + N_2}{N_2} \right) = V_2$$

To use the autotransformer as an impedance matching network, use the following equations:

$$Z_1 = \frac{V_1}{I_1} = \frac{V_2}{I_2} \left(\frac{1}{(a+1)^2} \right) = \left(\frac{1}{(a+1)^2} \right) Z_2$$

$$100 = 150 \left(\frac{1}{(a+1)^2} \right) \qquad \sqrt{\frac{100}{150}} = \left(\frac{1}{a+1} \right)$$

$$a \equiv 0.225 \quad \text{thus } N_1 = 0.225 N_2$$

Conclusion

An introduction to the basic design and application of the wide band transformer and autotransformer has been covered. It can be used as a guide to help in the development and specification of wide band transformers and autotransformers, which are available in both PC and surface mount configurations.

How To Specify Power Transformer and Filter Ratings

The purpose of this section is to provide a practical guide for the selection of a power supply transformer and filter components. A number of basic assumptions are made to avoid an academic discussion of unnecessary material. For those interested in a rigorous theoretical analysis, there are a number of fine references available. Additionally, circuit analysis using appropriate analysis software (SPICE or its equivalent) is recommended in the rare circumstances where a better understanding of a particular situation is needed, or when it becomes necessary to optimize some aspect of the design. Computer analysis is particularly useful in understanding areas that are difficult to approach using traditional circuit analysis methods, areas such as capacitor RMS ripple current.

One of the more esoteric problems encountered by the circuit designer is the selection of power transformer ratings for a particular DC power supply. The designer is immediately confronted with a number of rectifier circuits and filter configurations. For the sake of simplicity, we will make some assumptions which should be valid for 99% of the average designer's applications, and present some useful rules of thumb appropriate for conservative design.

Filters

We will immediately discard the consideration of choke input filters and confine our choice to capacitor input filters because of the following:

- It is desirable to eliminate the weight and cost of chokes.
- It can be assumed that if a regulator is used it will provide sufficient extra ripple reduction so that an L-C section is not required. In addition, the regulator will compensate for the poor output voltage regulation with load, inherent in capacitor input systems.

The remaining disadvantages of the capacitive input filter system are caused by the discontinuous secondary current flow (high peak-to-average ratio of forward diode current). Current is drawn in short, high amplitude pulses to replace the charge of the filter capacitor which discharges into the load during diode off time. This results in higher effective RMS values of transformer secondary current. However, the transformer average VA rating is the same as the choke input filter because the higher DC output voltage obtained at the capacitor compensates for this effect. In addition, except perhaps for supplies handling very high currents, average semiconductor diodes will meet most of the peak or surge current requirements of capacitive filters.

Rectifier Circuit

The remaining choice is that of a rectifier circuit configuration. The most common single phase circuits are:

- Half-Wave (single diode).
- Full-Wave Center-Tapped, FWCT (two diodes).
- Full-Wave Bridge (four diodes).
- Dual Complementary Supply - "Full-Wave Center-Tap" - FWCT (four diodes).

The only advantages of the half-wave rectifier are its simplicity and the savings in cost of one diode. Its disadvantages are many:

- Extremely high current spikes are drawn during the capacitor charging interval (only one current surge per cycle). This current is limited only by the effective transformer and rectifier series impedance, but it must not be too high or it will result in rectifier damage. This short once-per-cycle current spike also results in very high secondary RMS currents.

- The unidirectional DC current in the transformer secondary biases the transformer core with a component of DC flux density. As a result, more "iron" is needed to avoid core saturation.

About the only time it would pay to consider using the half-wave rectifier is for very low DC power levels of about 1/2 watt or less. At these levels a power transformer cannot be reduced very much in size (at reasonable cost) and a small filter capacitor will be large enough for adequate DC smoothing.

The remaining single-phase rectifier circuits are of the "full-wave" type. Secondary current surges occur twice per cycle so that they are of smaller magnitude and the fundamental ripple frequency is double the supply frequency (i.e., 120 Hz rather than the 60 Hz of a half-wave system). All full-wave rectifiers also have the same basic rectified waveform applied to the filter capacitor.

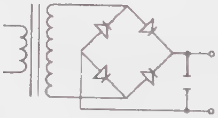
Other Factors

- **Full-Wave Center-Tap**
  Uses 1/2 of secondary winding at a time.
  Requires center-tap.
  Uses 2 diodes.

- **Full-Wave Bridge**
  Uses full secondary winding continuously.
  No center-tap required.
  Uses 4 diodes.

As can be seen above, the choice between FWCT and Bridge configurations is a tradeoff. The bridge rectifier has the best transformer utilization but requires the use of 4 diodes. The extra diodes result in twice the diode voltage drop of a FWCT circuit so that the FWCT is usually preferable in low voltage supplies. See Figure 6.

FIGURE 6. Full Wave Bridge.



The "dual complementary rectifier circuit" is the combination of two FWCT circuits and is a very efficient way of obtaining two identical outputs of reversed polarity sharing a common ground. It is also called a "center-tapped bridge rectifier." See Figure 7.

FIGURE 7. Dual Complementary Rectifier.

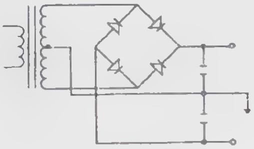
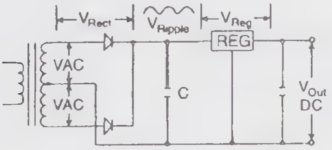


Figure 8 below represents a full-wave center-tapped rectifier using a capacitive filter and is the most common selection for moderate power, regulated DC supplies. The following assumptions can safely be made:

- V_{REG} is approximately 3 volts DC or greater.
- V_{RECT} is about 1.25 volts DC.
- V_{RIPPLE} is about 10% VDC peak.

FIGURE 8. Full Wave Center Tap.



This equation may be used for determining the transformer secondary voltage:

$$V_{AC} = \frac{V_{Out} + V_{Reg} + V_{Rect} + V_{Ripple}}{0.9} \times \frac{V_{Nom}}{V_{Lowline}} \times \frac{1}{\sqrt{2}}$$

Where: 0.9 = rectifier efficiency (typical).

$\frac{V_{Nom}}{V_{Lowline}}$ is the ratio of the nominal AC line voltage to the required low line conditions.

A sample illustration of the above will be shown for a supply requiring an output of 5V DC at 2A DC to operate down to an input voltage of 95V RMS.

$V_{OUT} = 5V$
$V_{REG} = 3V$
$V_{RECT} = 1.25V$
$V_{RIPPLE} = 0.5$ (1V p-p)

$$V_{AC} = \frac{9.75}{0.9} \times \frac{115}{95} \times \frac{1}{\sqrt{2}} = 9.27\ V_{AC}$$

Therefore, the transformer secondary voltage can be specified as about 18V CT. For a bridge rectifier of the same output requirements, the only change is that:
$V_{RECT} = 2 \times 1.25 = 2.5V$

As a result VAC will be reformulated as:

$$V_{AC} = \frac{11}{0.9} \times \frac{115}{95} \times \frac{1}{\sqrt{2}} = 10.46\ V_{AC}$$

so that the transformer secondary voltage now becomes about 10.5V.

Transformer Secondary Current

The remaining step is to determine the transformer RMS secondary circuit. This can be accurately determined only by complex analysis. However, for practical engineering purposes the following chart may be used:

Rectifier Type	**Filter Type***	**Required RMS Current**
Full-Wave Center-Tap	Choke Input	0.7 x DC Current
Full-Wave Center-Tap	Capacitor Input	1.2 x DC Current
Full-Wave Bridge	Choke Input	DC Current
Full-Wave Bndge	Capacitor Input	1.8 x DC Current

*Even though we have have dropped choke input filters from this discussion, they are included for reference.

For instance, in our particular example (5 V, 2A DC supply) the transformer RMS current would be:

for FWCT 1.2 x 2 = 2.4 A
for bridge 1.8 x 2 = 3.6 A

The total transformer specification would then be:

Circuit	Secondary Rating
FWCT	18.5 CT @ 2.4A RMS = 43.2 VA
bridge	10.5 @ 3.6A RMS = 36 VA

Dual Complementary Supply

One more common example will be given, i.e., a dual complementary supply for ± 15 V @ 100 mA DC. See Figure 9.

$V_{OUT} = \pm 15$
$V_{RECT} = 1.25$
$V_{REG} = 3$
$V_{RIPPLE} = 0.75$ (1.5 V p-p)

$$V_{AC} = \frac{(15 + 3 + 1.25 + 0.75)}{0.9} \times \frac{115}{95} \times \frac{1}{\sqrt{2}} = 19V$$

1AC = 1.8 x 100 mA = 180 mA RMS.

The transformer secondary rating is 38 V CT @ 180 mA RMS. A precautionary calculation remains to be made. That is, the increase in voltage at the filter capacitor (into the regulator) caused by a high line condition. If we assume our highest line voltage to be 130 V AC then the transformer output (compared to low line) would rise by the ratio 130/95. In the 5V supply, for instance, the following would happen:

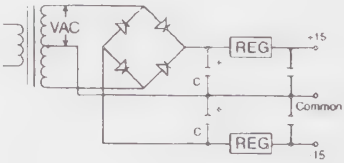
$$V_{AC} = \frac{130}{95} \times 9.27 = 12.7V$$

In the dual complementary +15 V supply:

$$V_{AC} = \frac{130}{95} \times 19 = 26V$$

The increase in output must be absorbed by the regulator, which results in higher regulator power dissipation. The illustrated values are safe for the typical IC regulator but should be checked in any specific application.

FIGURE 9. Dual Complementary Supply.



ADDITIONAL FACTORS TO BE CONSIDERED IN TRANSFORMER SELECTION

Load Regulation

It has been assumed in the previous discussion of the change in transformer secondary voltage with line voltage that no change has been occurring in load current. Therefore, the transformers would seem to be ideal and the transformer secondary voltage (VAC) will always be the same. Actually, all the voltages calculated are assumed to be full load.

Since transformers are not ideal and have an internal impedance or "regulation" characteristic, variations in load current may cause a problem. If the load should be "light" at "high line," then there will be an additional rise in secondary voltage, beyond that due to the rising line voltage, caused by the decreasing voltage drop in the transformer windings.

Most smaller VA transformers (<10VA) have a load regulation of 20% or higher. This means that the transformer no-load voltage will be 20% or more higher than rated full-load voltage. This must then be taken into account in the calculation of maximum VAC (and DC voltage into regulator) with low-load currents.

Due to the inherent design characteristics of transformers, "regulation" will vary inversely with size (or VA rating). In larger transformers size is determined primarily by the heat generated by internal losses. In smaller transformers (low VA rating) size is determined by the maximum permissible no-load to full-load regulation. Even though this is an important design limitation, most transformer manufacturers do not publish load regulation data.

It is possible to estimate the output voltage at intermediate loads since load regulation varies in an almost linear manner. For example, a transformer has a full-load rating of 16 V @ 6.25 A and a regulation of 10%. Its no-load output would be 10% more than 16 or 17.6 V. At half-load (3.12 A) its output would be 5% more than 16 or 16.8 V. Similar estimates can be made for any percentage load.

Another factor to bear in mind is that it is possible to safely exceed the VA rating of many small power transformers. If the added regulation (drop in output voltage) is acceptable, an "overload" may be permissible because the design is regulation-limited rather than heat rise-limited. If such a choice is being considered, the decision should be checked with the transformers manufacturer's Design or Engineering department.

Temperature Rise

In power transformers over 25VA, temperature rise becomes a factor. The transformer may be constructed with materials capable of withstanding higher temperatures and be a perfectly valid design. However, the extra power dissipated may cause heating of nearby components. This added power loss increases the total power dissipated in the circuit area. The problem is not the internal temperature of the transformer, but the actual increase in watts lost.

Shielding

Certain AC power line noise and transients will be fed through to the transformer secondary because of the capacitance between windings. This is a problem which is very difficult to analyze. Whether or not it is a problem in a particular application can be best determined empirically.

If such feedthrough is a problem, the most common first step is to use an electrostatic shield between windings. This effectively reduces the interwinding capacitance. An equal and sometimes superior approach is to choose transformers with nonconcentric windings, i.e., with primary and secondary wound side-by-side rather than one over the other. Both result in at least order of magnitude reductions in capacitance. The "nonconcentric" approach, however, gives superior reductions. It also results in higher insulation resistance and makes it simpler to obtain higher insulation test voltages.

Certain types of feedthrough cannot be much affected by the transformer design and other approaches such as line filters or MOVs or ZNRs (transient surge suppressors) may have to be considered.

Summary

This has been an attempt to provide a simple, practical method of determining transformer ratings. Certain basic assumptions have been made and this section is not meant as a rigorous academic analysis.

COMMENTS ON CAPACITOR & DIODE SELECTION

Capacitor Selection

For low current supplies ($I_{OUT}<I_A$) capacitor selection is relatively straightforward. Capacitance is found by the simple formula:

$$C = \frac{I_L \times (6 \times 10^{-3})}{\Delta V}$$

where:
I_L = DC load current
ΔV = peak-to-peak ripple voltage
ripple frequency = 120 Hz.

This yields 2000µF/amp for 3V p-p ripple. At DC currents below 1 Amp, capacitor heating is usually not a problem and peak-to-peak ripple voltage is the determining factor in capacitor size.

At higher values of capacitance, where the ratio of capacitor outside surface area to volume is significantly lower, internal heating becomes a problem. Ripple current rating may be a determining factor in capacitor selection, rather than ripple voltage. In many cases, capacitor size will have to be increased to prevent excessive internal heating. Manufacturer's data sheets should be consulted (after an initial selection is made) to ensure that capacitor ripple current ratings are met. Remember that the RMS ripple current ratings shown on capacitor data sheets are not the same as DC load current. RMS ripple current in a capacitor input filter is 2 to 3 times the load current. In addition, the time-to-failure used to rate capacitors on data sheets is often 10,000 hours. For five-year life (40,000 hours), ambient temperature may have to be derated 30°C from the data sheet rating. Capacitor life roughly doubles for each 15°C reduction in operating temperature. The following calculations illustrate a typical design example:
Assume $I_L = 3A$, $\Delta V = 4V$ p-p, and VDC = 12V.

$$C = \frac{(3A) \times (6 \times 10^{-3})}{4V} = 4{,}500\mu F$$

Manufacturer's rating on a 4,600µF/20V capacitor @ $T_A = 65^\circ C$ is 3.1 A RMS. Dividing by 2.5 to convert from RMS ripple current to output current yields a maximum DC load current of 1.24 Amps. Obviously either a large capacitor is required or ambient temperature must be reduced.

As a final note, be sure to check whether the data sheet ratings are for still or moving air. Computer grade capacitors are often rated only for moving air. Other types may be rated for still air and are, therefore, actually more conservatively rated.

Remember that capacitors are the number one cause of power supply failure. Don't let your supplies dominate the statistics column!

Diode Selection

The RMS value of the current flowing into a capacitor input filter is 2 to 3 times the DC output current because the current is delivered in short pulses. Assuming a full-wave center-tap or bridge, this means that although each diode is conducting only on alternate half cycles, it should be rated for at least the full output current. To ensure adequate surge capability during turn-on, a diode rating of at least twice the output current is recommended, especially for higher current supplies where the ratio of filter capacitance to output current is somewhat higher. Keep in mind that axial lead diodes achieve most of their heat sinking through the leads. Short leads soldered to large area standoffs or printed circuit pads are definitely recommended.

For "short circuit proof" IC regulated supplies using three-terminal regulators, an additional diode derating may have to be used. Long-term output shorts do not harm the regulator, which goes into a current limit or thermal limit mode to protect itself. The diodes, however, may experience a substantial current increase during the short. Regulator data sheets should be consulted for current limit values, keeping in mind that current limit is a function of input-output voltage differential. At high input voltages, the short circuit current of IC regulators is often less than full load current, tending to alleviate this problem.

METHOD OF DETERMINING SECONDARY CURRENT RATINGS

The secondary currents shown in the tables are RMS ratings. Depending upon rectifier circuit configurations, the RMS secondary current is different from the DC output current. This is indicated in the chart below:

Rectifier Type	FilterType	RMS Secondary current is
Full-Wave Center-Tap	Choke Input	= 0.7 x DC Amps
Full-Wave Center-Tap	Capacitor Input	= 1 to 1.2 x DC Amps
Full-Wave Bridge	Choke Input	= DC Amps
Full-Wave Bridge	Capacitor Input	= 1.6 to 1.8 x DC Amps

For example:
In a F.W. Bridge circuit with a capacitive filter, if the load is 1 Amp DC, the RMS Secondary current is 1.6 to 1.8 Amp RMS.

References
For more detailed theoretical analysis the following are recommended:

1. Reuben Lee, Electronic Transformers & Circuits, 1947,
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2. EE Staff - MIT, Magnetic Circuits & Transformers, 1943,
John Wiley & Sons
3. O.H. Schade, Proc. IRE, vol 31, p. 356, 1943

Application Information

Wide Band Transformers applications include: impedance matching, voltage or current transformation, DC isolation, balanced/unbalanced mixing, matching, power splitting, coupling and signal inversion.

Power Factor Controller Transformers are intended for use in AC-DC pre-converter circuits for electronic ballast and off-line switching power converter applications.

Current Sensing Transformers incorporates the current and sense windings in one package thus requiring minimum board space. The sensitivity may be changed easily by adjusting the value of the sense resistor. Applications include feedback control, overload sensing, detecting load drop or shutdown, and proportional output.

Current Sensors function as encapsulated or molded secondaries of a current transformer. The conductor carrying the current to be measured is the "one turn primary." Sensitivity may be enhanced by increasing primary turns. Sensitivity can be doubled by looping the conductor through the hole twice. Applications include sensing branch circuit overload, switcher feedback, and detecting load drop or shutdown.

Switchmode Transformer Assemblies are applicable in all types of switching power supply circuits including forward converter, flyback, and bridge types. Powdered iron E cores can also be used for power inductors as well as switching power transformers. The high saturation levels of powdered iron combine with the economy of the E core shape to provide inductors that have larger energy storage capacity and are straightforward to design.

Audio Output, Line Matching Transformers are used to couple speakers to line outputs of amplifiers. Designed for Hi-Fi, stereo, Guitar amplifiers, and P.A. system applications. High Wattage units are designed for push-pull tube output circuits. Molded or cast construction produces a sealed unit that can withstand rugged to severe requirements.

Audio Transistor Transformers are used in amplifiers and transmitters, in input, driver, and output circuits.

Audio Telephone Coupling Transformers are designed to meet FCC Part 68 requirements to provide line isolation, impedance matching and line balance.

Broadcast Quality Audio Transformers with or without magnetic shielding can be used for the following applications:

- Emitter or mic to Line
- Isolating
- Collector to Line
- Line to Base
- Matching
- Bridging
- Hybrid

Audio Input Broadcast Quality Transformers have maximum shielding and are sealed for long life and stable characteristics. Humbucking construction with electrostatic shield between primary and secondary connected to core and terminal provides excellent noise and hum attenuation. For radio and TV broadcast applications such as:

- Line to Line
- Mic to Grids
- Mic to Base
- Line to Grids
- Line to Base
- Bridging
- Plates to Line
- Mixer to Base

Voice Transformers should have a high cross-talk isolation dB factor.

Automatching Transformers are suitable for driving from solid state amplifiers where isolated windings are not required.

Pulse Transformers find uses in digital and data processing, line coupling, matching and isolation, wide band and pulse applications including thyristor/triac firing circuits, and for trigger source isolation in half and full wave SCR power control circuits, including motor speed controls, heater controls, and incandescent lighting controls.

Trigger Transformers requires switching circuit to trigger primary. This is used to trigger a capacitive discharge in an electronic photo flash or strobe tube.

Power Transformers are for power applications in solid state control and instrumentation, and isolated DC supplies requiring single or multiple outputs. Some uses are in console and portable Stereo AM-FM, CD, and Tape Players. They can be designed for use with selenium or silicon rectifiers in capacitor or choke input circuits for power supplies, test equipment, pre-amps, etc. Suitable for single phase, half wave, full wave center tapped, full wave bridge, or voltage double circuits. Industrial applications include power supplies, controls and instrumentation.

Rectifier Transformers are for laboratory supplies, test equipment, or experimental power supplies. Note: If used for intermittent service (i.e.: 2 min. on, 6 min. off), increase ratings 100%. They can be used with selenium or silicon rectifiers in capacitor or choke input circuits for power supplies, test equipment, pre-amps, etc. Suitable for single phase, full wave center tapped, or bridge circuits.

400 Hertz Transformers may be used in rectifier, signal reference, and filament heating circuits, etc.

Plate Transformers for power supply applications are rated for Continuous Service and for Intermittent or Amateur Service.

DC to AC Inverters are for inverter power supplies with DC input and AC square wave output.

Line Adjusting Autotransformers are ideal for line voltage adjustment for electronic equipment or small appliance operation.

Line Isolating Transformer provides circuit isolation with an electrostatic shield between primary and secondary.

Control Transformers are used to adjust a supply voltage to match a load requirement. Applications include:

- supply machine tool control circuits
- machine lighting and circuit isolation
- actuating relays, contactors, bells, signal and alarm systems
- operating small motors, solenoid operated valves and dampers.

Buck-Boost Transformers are control transformers with low voltage secondary windings. By field connecting the primary and secondary windings in an autotransformer configuration, Buck-Boost transformers offer an economical solution to the adjustment of line voltages that are slightly above or below normal. When a Buck-Boost transformer is connected as an autotransformer, only a portion of the load kVA is actually transformed. The majority of the load kVA is passed directly from the source to the load and is not transformed. For this reason, a Buck-Boost transformer may be used to supply a much larger load kVA than is indicated on the nameplate. Buck-Boost transformers can be used to adjust stable voltages only. Fluctuating line voltages should be regulated with a Computer Regulator or line Voltage Conditioner.

Voltage (Potential) Transformers are used (with AC voltmeters) to measure the voltage of high voltage systems.

Energy Limiting Transformer (Class 2) applications include:

- Chimes and door bells
- Appliances
- Decorative lighting
- Furnaces and humidifiers
- Security systems
- Vending machines
- Machine and CAM applications
- Industrial applications
- Intercom systems
- Central vacuum systems

Autotransformers are the most economical alternative to distribution transformers of adjusting a supply voltage when isolation from the line is not necessary and local electrical codes permit. Units can be used in either step-up or a step-down application. Units are designed to withstand motor inrush currents.

Distribution Transformers (Single and Three Phase) are suitable for a variety of lighting, resistive and inductive loads. They are generally used to provide power for lighting, resistive heating and many motorized machine applications. These transformers are provided either ventilated or totally enclosed and can be wall or floor mounted.

Shielded Distribution Transformers (Single and Three Phase) provide a copper electrostatic shield between the primary and secondary windings. The shield is grounded and thus shunts the noises and transients to the ground path rather than passing them along. For most applications, a common-mode noise attenuation of 60 dB is recommended.

Low Temperature Rise Transformers (Single and Three Phase) have high continuous overload capacity.

Encapsulated Transformers are suitable for both indoor and outdoor use. The core and coils are totally enclosed in electrical grade silica and epoxy then baked to seal out moisture and air-borne contaminants. The encapsulated design is especially suited for installations in harsh environments where dust, lint, moisture, and corrosive contaminants are present.

High Voltage Distribution Transformers (2400V and up) are designed primarily for use in stepping down incoming high voltage power to a lower operating voltage for commercial, institutional, or industrial applications. The ventilated construction is suitable for indoor installations.

Transformers for Nonlinear Loads are powered by either switching power supplies or some type of rectifier circuit. Used in computers, fax machines, copiers, printers, cash registers, UPS, and solid-state ballasts, etc. In industrial plants, one will find other electronic devices like variable speed drives, HID lighting, solid-state starter and solid-state instruments, etc. They all contribute to the distortion of the current waveform and the generation of harmonics. As the use of electronic equipment increases and it makes up a significant portion of the electrical load, many concerns are raised about its impact on the electrical power system. K factor is defined as a ratio between the additional losses due to harmonics and the eddy losses at 60 Hz. It is used to specify transformers for nonlinear loads. For existing facilities, the actual harmonic content can be measured. With the measured results, ANSI/IEEE recommended procedure (ANSI/IEEE C57.110-1986) is used to determine the required K factor.

Motor Starting Transformers are for use with magnetic or manually operated motor controllers. Transformers are thermostatically controlled.

Drive Isolation Transformers are designed to meet rugged demands of AC and DC variable speed drives and also to provide the required voltage change to match the load operating level. The windings are torque wound to yield low operating looses, high overload capacity and excellent mechanical stress withstand. Electrically balanced windings and high density fiberglass duct sticks in combination with the coil support structure render high axial short circuit withstand. The separate primary and secondary windings provide both magnetic and electrical isolation between the incoming line and the SCR load. Windings are designed to withstand overcurrent of the rated load. Typical operation may allow a duty cycle.

Design of switching power magnetics can be accomplished in a relatively simple manner by limiting magnetic configurations to a few structures. These structures have been chosen not only for their versatility, but for their low cost. Manufacturer's catalogs contain dimensional information as well as design information in the form of design curves. By using these curves, the complete transformer can be designed.

STRUCTURE SIZE

The first step in the design consists of choosing a minimum structure size consistent with the output power required. The approximate power capabilities of each structure are given in the specifications. If five or six outputs are required, a larger structure may be required to allow the copper to fit in the available winding area.

TURN COUNT

For a given core size, the ability of an inductor to operate without saturating is directly proportional to its turn count N_P. The normal saturation specification is E-T or volt-time rating. The E-T rating is the maximum voltage, E, which can be applied over a time of T seconds. (The E-T rating is identical to the product of inductance L and peak current I). Equation 1 defines a minimum value of N, for a volt-time product of E-T:

Equation 1. $$N_P = \frac{E\text{–}T \times 10^8}{BA_E}$$

Where:
E-T = minimum volt-time rating in volt-seconds
B = maximum allowable flux density
A_E = effective cross sectional core area.

Equation 1 can be plotted as a Volt-Time vs. Primary Turns curve for choosing N_P. Locate the required E-T rating on the vertical axis. Move horizontally to the curve. From this point drop vertically to the horizontal axis and read N_p. This value for N_P should allow nonsaturating operation to 100°C. For lower operating temperatures, a slightly smaller value for N_p can be used.

Secondary turn count is a function of duty cycle and primary turn count. For a flyback system, use equation 2:

Equation 2. $$N_S = N_p \frac{(V_S + V_D)}{V_L D} (I - D)$$

For a forward converter use equation 3:

Equation 3. $$N_S = N_p \frac{(V_S + V_D)}{V_L D}$$

Where:
N_S = secondary turn count
V_S = secondary output voltage
V_D = voltage drop across the rectifier and choke in the secondary
D = duty cycle
V_L = voltage across the primary.

For the flyback system, D is seldom greater than 0.5. For the forward converter, D is the duty cycle of the rectified output, and can approach 0.9 for a wave rectified output. Known conditions should be used to calculate N_S. At minimum input voltage and maximum output power, the supply will operate at maximum duty cycle. This is therefore a good point to use to determine N_S.

WIRE SIZE

Once all the turn counts have been determined, wire size must be chosen for each winding. Power losses in the transformer windings cause a temperature rise ΔT in the transformer. The amount of loss will depend on how much current is being drawn from the winding, how many feet of wire and what wire size is used. The power loss is a function of the amount of resistance in the wire. This resistance is composed of a DC resistance, R_{DC}

and an AC resistance, R_{AC}. At 20 kHz, wire smaller than about 20 AWG will have $R_{DC} >> R_{AC}$, and R_{AC} can effectively be ignored. For larger wire sizes, it may be necessary to use stranded wire or foil. Let's assume R_{AC} is small and can be ignored. Then all of the copper losses will be due to R_{DC}. This power loss should be distributed among the primary and secondary windings to minimize hot spots in the transformer. The temperature rise in a given structure is a function of thermal resistance and surface area.

For flyback transformers use equation 4:

Equation 4. $$R_{DC} = \frac{3\,P_L}{D_I\,I^2}$$

For the forward converter use equation 5:

Equation 5. $$R_{DC} = \frac{P_L}{I_{av}{}^2}$$

Where:
P_L = maximum allowable power loss in the winding
I = peak current in the winding
D_I = duty cycle of the current ramp
I_{av} = average current in the winding.

Once R_{DC} for all the windings has been calculated, the wire size can be determined using the average length per turn listed for each coil form. Divide the R_{DC} by the product of the average length per turn times the turn count. This value of R_{DC} per foot can be found listed in any wire manufacturer's data sheet. From their data, choose a wire size that has the same or slightly less R_{DC} per foot.

WILL IT FIT?

After the wire sizes have been determined, it is necessary to check fit, to see if the available winding area will accommodate the copper calculated in the previous steps. The available winding area for the structure chosen is given on its data sheet. The area each winding occupies is determined. Then the areas occupied by each winding are added. To this total, an area allowing for interwinding insulation and shielding is added. The grand total is then compared to the available winding area. If the area of the windings is greater than the allowed structure area, either wire size must be reduced, or a larger structure must be chosen. Of course, a reduction in wire size will result in increased losses in the transformer.

To determine winding area, use a Turns per Layer vs. Wire Size curve. The first step is to determine the turns per layer, n_l. Using the curve, draw a horizontal line from the required wire size on the vertical axis. Where this line intersects the curve, drop vertically to the horizontal axis and read n_l. Next you find the number of layers, l, by dividing turns N by n_l using equation 6.

Equation 6. $l = \frac{N}{n_l}$

If l contains a fractional part, round l up to the next integer value.

Next find the area per layer, a_l. Multiply the winding width as shown on the individual coilform drawing by the wire diameter. Multiply this value by the number of layers to find the area occupied by the winding by using equation 7:

Equation 7. $a_m = l a_l$

Repeat this procedure for each winding and add to find the total area occupied by the windings. See equation 8.

Equation 8. $a_l = a_{m1} + a_{m2} + a_{m3} + \ldots$

The total winding area, a_l, added to the area of insulation and shielding, a_i, must be less than the allowed structure area, a_w. See equation 9.

Equation 9. $a_w \geq a_l + a_i$

a_w is given in the structure data sheet.

AIR GAP

Now that the structure size and winding turn counts have been found, the air gap must be determined. For most forward converters, inductance should be as large as possible and no air gap is required. However for flyback transformers and chokes, an air gap is necessary. For economic reasons, E cores are usually gapped in all three legs. The amount of gap per leg can be found from a Inductance Factor vs. Air Gap curve. To use this curve it is necessary to convert inductance, L, to inductance factor, A_L, by using equation 10.

Equation 10. $A_L = \frac{L}{N^2}$

Draw a horizontal line from the vertical axis at the A_L value obtained from equation 10. At the intersection of this line and the curve, drop vertically to the horizontal axis and read g, the required gap per leg.

SUMMARY

The transformer should now be ready to wind. Sufficient insulation must be used for the required voltage isolation. Windings must be uniform and carefully wound to minimize leakage inductance. Some small adjustments may be necessary to obtain the exact results desired, but this procedure will provide a good approximation.

Glossary

Air Core Inductance - L_o (Henries) - The inductance that would be measured if the core had unity permeability and the flux distribution remained unaltered.

Amplitude Permeability - μ_a - The quotient of the peak value of the flux density and the peak value of the applied field strength at a stated amplitude of either, with no static field present.

Coercive Force - H_c (Oersted) - The magnetizing field strength required to bring the magnetic flux density of a magnetized material to zero.

Core Constant - C (cm^{-1}) - The summation of the magnetic path lengths of each section of a magnetic circuit divided by the corresponding magnetic area of the same section.

Core Constant - C_2 (cm^{-3}) - The summation of the magnetic path lengths of each section of a magnetic circuit divided by the square of the corresponding magnetic area of the same section.

Curie Temperature - T_c (˚C) - The transition temperature at which a ferromagnetic material loses its ferromagnetic properties and becomes paramagnetic (μ approaches 1).

Disaccommodation - D - The proportional decrease of permeability after a disturbance of a magnetic material, measured at constant temperature, over a given time interval.

Disaccommodation Factor - DF - The disaccommodation factor is the disaccommodation after magnetic conditioning divided by the permeability of the first measurement times $\log_{10}$ of the ratio of time intervals.

Effective Cross-Sectional Area - A_e (cm^2) - In a structure containing a nonuniform cross section, the effective magnetic cross section is the area of a structure with uniform cross section which is equivalent to the first for purposes of magnetic calculations.

Effective Dimensions of a Magnetic Circuit -
Area A_e (cm^2), Path Length ℓ_e (cm) and Volume V_e (cm^3)
For a magnetic core of given geometry, the magnetic path length, the cross-sectional area and the volume that a hypothetical toroidal core of the same material properties should possess to be the magnetic equivalent to the given core.

Effective Disaccommodation Coefficient - DF_e - The actual disaccommodation of a magnetic circuit whose material permeability has been reduced to μ_e by gapping.

Effective Hystersis Coefficient - $C_{h(e)}$ (/Gausses) - The actual hysteresis loss in a magnetic structure whose permeability has been reduced to μ_e by gapping.

Effective Magnetic Path Length - ℓ_e (cm) - In a structure containing a nonuniform cross section, the effective magnetic path length is that length of a similar structure with uniform cross section which is equivalent to the first for purposes of magnetic calculations.

Effective Magnetic Volume - V_e (cm^3).

Effective Permeability - μ_e - For a magnetic circuit constructed with an air gap or air gaps, the permeability of a hypothetical homogeneous material which would provide the same reluctance.

Effective Temperature Coefficient - TC_e - (/°C) - The actual temperature coefficient of a magnetic structure whose material permeability has been reduced to μ_e by gapping.

Federal Communications Commission - FCC - regulates commercial (common-carrier) interstate and international telegraph and telephone and all radio and television operations in the United States and its territories, excluding agencies of the federal government.

Field Strength - H (Oersted) - The parameter characterizing the amplitude of alternating field strength.

Flux Density - B (Gauss) - The corresponding parameter for the induced magnetic field in an area perpendicular to the flux path.

Flux Density, saturation - B_s (Gauss) - The maximum intrinsic induction possible in a material.

Hysteresis coefficient - C_h (/Gausses) - The coefficient in the Legg equation which separates the hysteresis losses from the eddy current and residual losses.

Incremental Permeability - μ_Δ - Under stated conditions the permeability obtained from the ratio of the flux density and the applied field strength of an alternating field and a superimposed static field

Inductance - L (Henries) - The magnetic flux linkages in Maxwells-turns per Ampere of magnetizing current.

Inductance Factor - A_L (nH) - Inductance of a coil on a specified core divided by the square of the number of turns. (Unless otherwise specified the inductance test conditions for inductance factor are at a flux density Gauss.)

Initial Permeability - μ_i - The permeability obtained from the ratio of the flux density, kept at Gauss, and the required applied field strength. Material initially in a specified neutralized state.

International Special Committee on Radio Frequency Interference - CISPR - from the French abbreviation for Comite International Special des Perturbations Radioelectriques. Operating under the auspices of the International Electrotechnical Commission (IEC), the stated objective is to promote the international agreement of EMI limits and measurement techniques to facilitate international trade.

Loss Angle - tan δ - Deviation from ideal phase angle (90°) due to losses.

Loss Factor - $\tan \delta/\mu_i$ – The phase displacement between the fundamental components of the flux density and the field strength divided by the initial permeability.

Magnetic Constant - μ_o - The permeability of free space.

Magnetic Hysteresis - In a magnetic material, the irreversible variation of the flux density or magnetization which is associated with the change of magnetic field strength and is independent of the rate of change.

Magnetically Soft Material - A magnetic material with a low coercivity.

Permeability, Absolute Complex - μ - The magnetic flux density divided by the magnetic field strength.

Permeability, amplitude - μ_a - The quotient of the peak value of the flux density and the peak value of the applied field strength at a stated amplitude of either, with no static field present.

Permeability, effective - μ_e - For a magnetic circuit constructed with an air gap or air gaps, the permeability of a hypothetical homogeneous material which would provide the same reluctance.

Permeability, incremental - μ_Δ - Under stated conditions, the permeability obtained from the ratio of the flux density and the applied field strength of an alternating field and a superimposed static field.

Permeability, initial - μ_i - The permeability obtained from the ratio of the flux density, kept at Gauss, and the required applied field strength. Material initially in a specified neutralized state.

Permittivity - ε - Often referred to as the dielectric constant of a material.

Power Loss Density - P (mW/cm^3) - The power absorbed by a body of ferromagnetic material and dissipated as heat, when the body is subjected to an alternating field which results in a measurable temperature rise. The total loss is divided by the volume of the body.

Q Factor - Q - The efficiency of an inductor, that is the ratio of series inductive reactance to loss resistance.

Relative Complex Permeability - μ_r - The relation of the absolute permeability of a substance or medium to that of a vacuum.

Relative Hysteresis Factor - C_h/μ_i^2 (/Gausses) - The hysteresis coefficient normalized to unit permeability so that it is strictly a material property.

Relative Loss Factor - $\tan \delta/\mu_i$ - Losses per unit of permeability. Figure of merit of a material.

Remanence - B_r (Gauss) - The flux density remaining in a magnetic material when the applied magnetic field strength is reduced to zero.

Return Loss - R_L - The return loss of a particular system is the scalar reflection coefficient expressed in decibels.

Saturation Flux Density - B_s (Gauss). The maximum intrinsic induction possible in a material.

Scatter Parameter - S-parameters, used in a square array of complex numbers consisting of the transmission and reflection coefficients of a waveguide component.

Standing Wave Ratio - SWR - At a given frequency in a uniform transmission line or waveguide, the ratio of the maximum to the minimum amplitudes of corresponding components of the field (or the voltage or the current).

Temperature Coefficient - TC - The relative change of the quantity considered, divided by the difference in temperatures producing it.

Temperature Factor - TF - The fractional change in initial permeability over a temperature range, divided by the initial permeability.

Tesla - T - 1 Tesla = 10^4 Gausses.

Verband Deutscher Elektrotechniker - VDE - is the Society of German Electrical Engineers which generates EMI specifications which become legal documents when they are accepted by the Ministe of Postal and Telecommunication Service, Bundesminister für Post und Telekommunikation (BMPT).

Voluntary Control Council for Interference by Data Processing Equipment and Electronic Office Machines - VCCI - was established by four Japanese organizations for the purpose of taking voluntary control measures against interference from data processing equipment and electronic office machines.

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Appendix A

Data Line Filtering

Noise is a major concern among designers of electrical equipment because failure of an electronic design due to unexpected or indeterminate noise sources is not uncommon. The FCC and other regulatory agencies are rigorously enforcing their noise requirements and this gives noise an often pivotal role in the ultimate success of a piece of electronic equipment. This section will address some of the important attributes of the noise associated with electrical and electronic equipment, particularly low signal level data line equipment of which digital equipment is one type. Some simple filtering solutions will also be discussed.

Noise Sources

Power converters are notorious sources of noise; they normally produce both common mode (occurring equally on all signal and common lines which refer to earth ground) and differential mode noise (occurring between signal and return paths). Noise from power converters generally predominates at the harmonics of the switching frequency, but some wide band noise is also produced. Because power converters are often required within electrical proximity of low signal level circuitry, they can be a major factor in determining the overall dependability of the data line system involved. Semiconductor devices may act as noise sources because of temperature (thermal noise), the adjoining of differing materials (contact noise), and the electron-hole movement of junction devices (shot noise). The magnetic components of modules which often support a data line system (for example, the transformer of a power converter, or some other high Q component) may themselves produce noise from ringing.

The logic circuits of equipment are yet another significant source of noise. At the switching of logic levels, the local power source can momentarily short to ground, introducing noise directly to both the ground plane and the dc power supply and affecting the entire electrical system. Internal noise may be generated from a number of sources which are electrically near the data line circuitry such as adjacent printed circuit boards or local magnetics (transformers, mixers, etc.). Semiconductor noises can also propagate to the threshold sensitive logic lines, to and from digital system clocks, and onto data lines. Externally created noise may emanate from a number of sources which are beyond design control. High frequency noise may appear on the power line, riding the 60 Hz signal usually without affecting the power circuitry, but providing a real danger to data line transmissions. Radiated noise may impinge on many parts of a piece of electronic equipment, especially through unshielded cabling, or through an ineffectively grounded case. Cabling is an important factor in the generation of EMI. Cables are generally the longest paths between circuit components and modules, thus providing an optimum situation for creating a loop antenna for radiating and receiving externally generated noise fields.

Noise Signal

Figure 1 shows an actual sine wave of 60 Hz line voltage (110 Vrms) occurring in a typical factory situation. The waveform contains random noise voltages of much higher than line frequencies.

FIGURE 1. Line source in a factory.

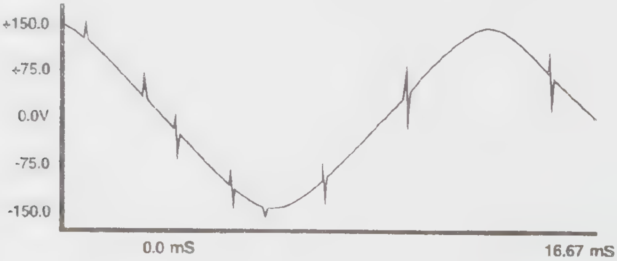


Figure 2 shows a 2 mS section of the 60 Hz wave with noisy spikes shorter than 5 µS and attaining 50 V peaks. Much higher frequency noise can also exist beyond the resolution of the waveform plot.

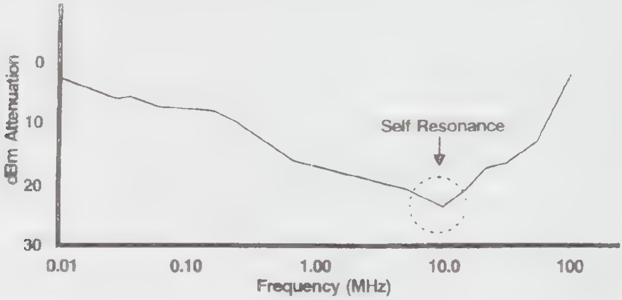
FIGURE 2. 2 mS section of the line.

Contaminated line currents are conducted directly into machines and devices connected to the same line as the offending machines; noise voltages can also be measured radiating through the air near these machines.

Generally, line filters are placed ahead of sensitive equipment to improve the situation of the 50 V high frequency factory noise mentioned above. Typically, a line filter is composed of a low frequency common mode magnetic structure which allows (differential) line current to pass freely without saturating the magnetic material (differentially induced flux lines cancel one another; this is the primary aspect of a common mode magnetic design). However, the noise common to both the source and return lines is attenuated.

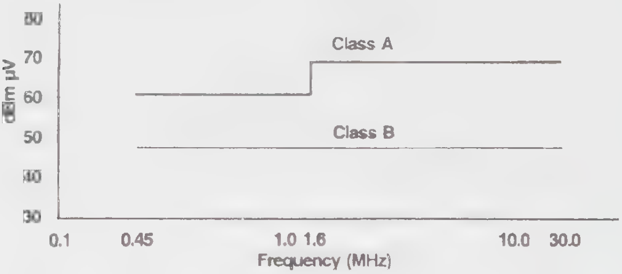
Line filters clean the incoming 60 Hz line signal by substantially attenuating frequencies through approximately 10 MHZ, with increasing failure to attenuate beyond this frequency (Figure 3).

FIGURE 3. Typical line filter response.



A real possibility still exists for higher frequency line aberrations, which often evade line filtering, to affect the integrity of the low voltage data transmission of data line equipment. Especially vulnerable is inter-equipment data transmission. An additional problem lies in the fact that equipment designated for factory use need only meet relatively less stringent noise requirements, although these machines have the potential for affecting their more sensitive counterparts (Figure 4).

FIGURE 4. FCC noise limits.



Certainly many data line devices and equipment are not intended to be situated in the severe environment of a factory. For example, the equipment required to meet FCC Class B limits do not usually have to deal with 50 V noise occurring from the power line; yet, Class B devices are becoming more and more susceptible to the usually low level/very high frequency noise common to contemporary data line equipment environment (Figure 4). Especially true is the sensitivity to noise of equipment which runs at high clock speeds and baud rates.

The system clocks of data line equipment are one concern of the noise conscious design engineer. Figure 5a shows a 20 MHZ clock signal with a spurious high frequency content and potential problem regions.

FIGURE 5a. 20 MHz clock with high frequency noise.

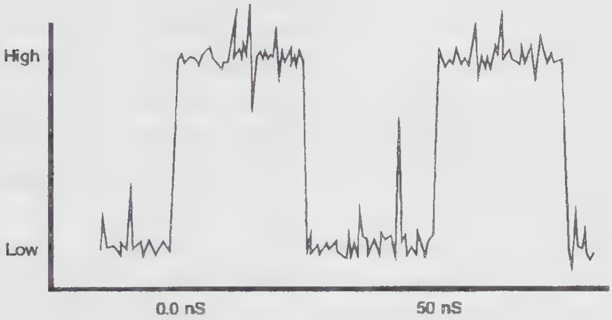
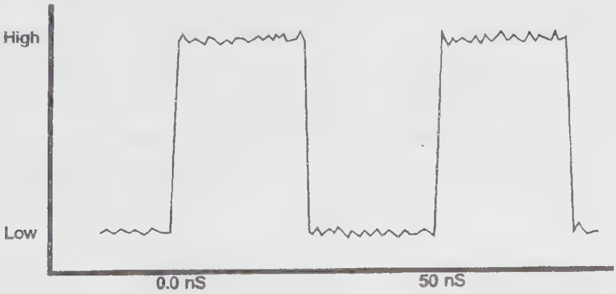


FIGURE 5b shows a noise-free 20 MHZ clock signal.

FIGURE 5b. A clean 20 MHz clock signal.



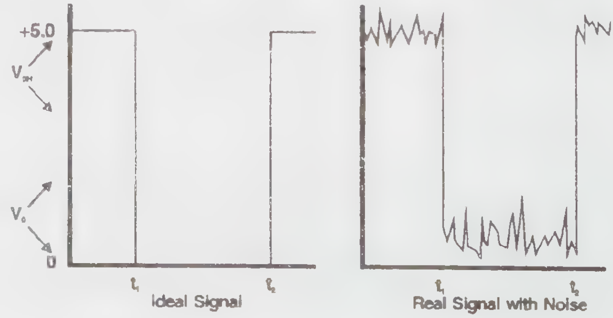
The transmission of data between equipment, whether within a local networking scheme or machine-to-machine communication through peripheral equipment like modems, can be disrupted and contaminated by unchecked noise.

Filtering in General

The obvious solution to the problem of noise is to eliminate the source, but this is often out of the question. Noise emanating within a piece of equipment is almost always a side effect of a critical module of a design. The next obvious solution is to eliminate the noise at or as near the source as possible. Noise sources may sometimes be obvious. A power supply, for example, is a logical place to expect ample noise to emerge. Sometimes the sources of noise are not at all predictable; moreover, the precise effect of a noise source is often illusive and only sporadically evident. The achievable solution to the problem of noise is to prevent it from affecting the systems and subsystems susceptible to it. After locating and filtering the obviously culpable noisemakers (e.g., the output of a power supply), the next step is to protect the inputs to the sensitive equipment and components. No more noise is then permitted into the sensitive equipment than the equipment may continuously tolerate. By filtering the inputs of sensitive components, the elusive and unmanageable sources of noise then become much less critical to the operation of the equipment.

Filtering is the ready solution to the problem of noise. Digital equipment and sensitive components can be made to withstand the subtle noise emanating from subsystems of the equipment as well as its external environment. Conducted noise often creates external fields, radiating noise to equipment and components which in turn induce conducted noise into these equipment and components. Also, the higher the frequency the less energy required for disruption. The cabling used to relay data line signals is particularly sensitive to external noise, both conducted and radiated. Because data line signals are essentially square waves with predetermined transition levels (e.g., 0.8V and 2.0V for TTL, Figure 6), the sudden encroachment of a noise spike in the electrical vicinity of the cable, or radiating near the cable, could cause a data error at the receiving end of the cable.

FIGURE 6.
TTL signals.



Ideal Signal

Real Signal with Noise

Filtering Techniques

A noise spike is caused by transient energy of a high frequency content. Because noise spikes are composed of such high frequency energy, the drastic attenuation of all high frequency noise just prior to reception of a logic signal would preserve the integrity of the original signal. Likewise, attenuating all high frequency components of the signal just ahead of a transmission cable would clean up the signal affected by noise within the data line system itself prior to any transmission. Also, the logic levels of the digital clock must be precise and unadulterated. Clock signals can benefit from high frequency noise attenuation, safeguarding the data line system from false state and logic transitions.

Magnetic Filtering

The solution to EMI problems typically arising in cabling is inherently magnetic. Cables are loop antennas created by a miss-match of a cable's signal and return line current. If the signal and return lines of a cabling scheme were positioned identically in space, the field created by the conducted current in the signal line would exactly cancel the field effected by the conducted current of the return line; of course this 180° perfect current reversal is not realizable, and we therefore must have a finite quantity of radiation due to any cabling.

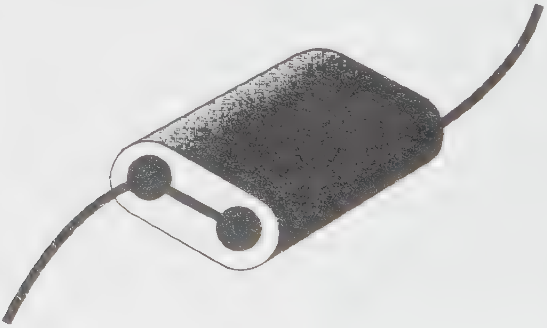
Twisting signal and return lines of a cable where possible breaks up the loop antenna structure into smaller loop antennas, achieving two useful results: 1) a substantial increase in the resonant frequency of the generated/received fields, and 2) allowing adjacent and inverted fields to cancel one another. Increasing the noise frequency is sometimes helpful, but generally not a sufficiently complete solution since applicable noise frequencies range into the hundreds of megahertz.

Ribbon type cabling usually does not lend itself to the twisting method described, unless special pre-twisted ribbon cabling is used (where the ribbon lines are twisted in pairs) but running return lines directly adjacent to respective signal lines is necessary to decrease radiated/received noise. Determining the actual ribbon cable line assignments is seldom an option provided an EMI engineer; usually, a solution to noise problems is a post mortem activity, such basic layouts having already been made for an equipment's design.

Filtering of a data line system, particularly when cabling and other signal wires are the objects to be safeguarded, is generally and most successfully realized through external magnetics. Creating a magnetic field around a wire can divert the high frequency energy from the signal propagating through the wire. By increasing the inductance of the wire or cable at an appropriate point along its path, a magnetic field can be created and magnified. Forming the cable or wire into a number of loops is the simplest method of increasing inductance. However, to obtain any useful inductance (thus appropriate attenuation of noise frequencies), many tens, or even hundreds, of loops would be required. Many loops are required to obtain the useful magnetic field because air alone is utilized to contain the field. Fortunately, magnetic materials can be used that have tens of thousands of times the magnetic capacity of air. Magnetic materials significantly increase the inductance with minimal looping of associated wires and cables.

Magnetic beads are often utilized for filtering; binocular cores, often used in balance to unbalanced ("Balun") transformers are also used in common mode filtering. A wire through one hole in a balun effectively constitutes a single turn (Figure 7).

FIGURE 7. A binocular filter.

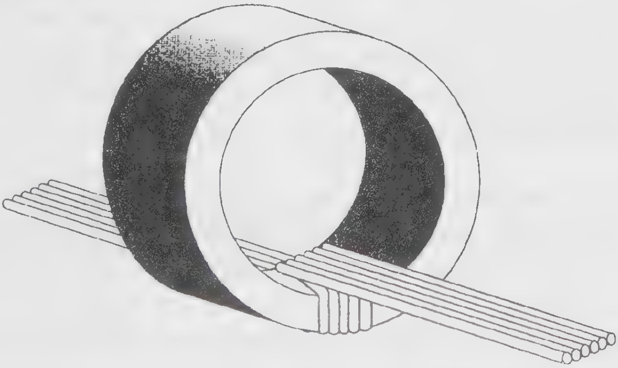


The advantage of using baluns is in their simplicity: one wire, one bead. Unfortunately, one bead may not suitably attenuate high frequency noise for many data line systems. Several beads may be strung together on the same wire for increased attenuation, but stringing beads on a signal wire, or multiple beads on a cable of multiple wires can result in both a difficult design and problems achieving repeatable design. Also, beads, each one magnetically isolated from all others, lack the ability to attenuate wide band common mode noise while supporting differential signals.

The problem of common mode noise is very real and somewhat elusive. Because it occurs equally and in phase on all signal lines with relationship to the ground plane, it becomes evident within an electronic system only when measured with reference to earth ground.

Within a single piece of equipment, common mode noise, though very real, may be difficult to detect. The subsystems of a piece of electronic equipment, each with semi-isolated ground planes, may display symptoms of common mode noise, a loss of signal integrity and a failing electrical reference. Likewise, when isolated pieces of equipment, having independent common references, are used in conjunction with one another (e.g., communicating through common signal lines), the common mode noise from each can become problematic. Fairly large magnetic toroidal structures may be used in a similar manner to a bead for filtering cables by fitting the structure around the entire cable. (Figure 8). While using a single, large structure allows common mode noise attenuation, it may lack the magnetic capability to provide effective protection of cabling because of the physical limitation of looping the cable through the large toroid more than once (thus increasing inductance).

FIGURE 8.
Filtering with
a large magnetic
structure.



Another method of protecting a data line system from noise is the use of special cable connectors which are either predominately capacitive or magnetic by design. The protection thus provided is limited to external connections. The capacitive connectors are typically more expensive than equivalent magnetics; magnetic connectors often have the additional advantage of being common mode by design. Multi-line common mode components are also available. Such components are magnetic, are compatible with ribbon cabling, and can be readily designed into a circuit. These multi-line common mode "Data Line" filters are available in standard DIP and surface mount configurations.

Rules of Thumb

A successful filtering program of a data line system should have the following features:

- The filter method must provide for the maintenance of the signal integrity.
- The filter method should be able to increase substantially the magnetic field at virtually any critical electrical location.
- The filter method must provide effective attenuation of all high frequency noise, as well as broadband common mode noise.
- The filter method must be compatible with standard cabling schemes, including ribbon-type cables, and individual signal wires.
- Ideally, the filtering method should be easy to implement, readily available, and successfully repeatable.

Conclusion

Noise sources are prevalent in a number of environments. Particularly severe is the factory situation where high frequency/high level noise transients intrude onto the power mains. To a lesser degree, such noise also may be found in the power stage of a piece of sensitive equipment. Power converter noise may also be evident outside the factory situation, where the more stringent regulatory agency noise limits come into effect.

The various noises which can destroy the repeatability and integrity of sensitive equipment, particularly data line equipment, are essentially of two types: common mode and differential mode. Common mode noise predominates. To successfully filter both types of noise, a magnetic scheme is optimum. Common mode noise reduction is effective when afflicted lines are coupled together through a common magnetic structure. Individual beads are only effective against differential mode noise.

A successful filtering scheme should have filters in place prior to transmission as well as prior to the reception of sensitive signals (e.g., at the beginning and at the end of a transmission cable). The filters should offer repeatable results and also be effective through a wide high frequency bandwidth.

Appendix B

Equations and Symbol Definitions

Basic Inductor Formulas

DC FORMULAS

1. Ohm's Law

$$E = IR = \frac{P}{I} = \sqrt{PR} \qquad I = \frac{E}{R} = \frac{P}{E} = \sqrt{\frac{P}{R}}$$

$$R = \frac{E}{I} = \frac{E^2}{P} = \frac{P}{I^2} \qquad P = I^2R = \frac{E^2}{R} = E\,I$$

E = Voltage in VOLTS $\quad$ I = Current in AMPERES

R = Resistance in OHMS $\quad$ P = Power in WATTS

2. Resistance in Series

$$R_T = R_1 + R_2 + R_3 + R_4 + \ldots$$

3. Resistance in Parallel

$$\frac{1}{R_T} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \ldots$$

$$R_T = \frac{R_1 \times R_2}{R_1 + R_2} \quad \text{(For 2 Resistors only)}$$

$$R_T = \frac{R_1}{\text{No. of Resistors}} \quad \text{(For Equal Resistors)}$$

AC FORMULAS

4. Ohm's Law

In a Purely Inductive Circuit (Neglect effect of Resistanace in circuit):

$$I = \frac{E}{X_L} \qquad X_L = \frac{E}{I} \qquad E = I X_L$$

In a Purely Capacitive Circuit (Neglect effect of Resistance in circuit):

$$I = \frac{E}{X_C} \qquad X_C = \frac{E}{I} \qquad E = I X_C$$

In a Circuit Containing Combinations of Circuit Elements:

$$I = \frac{E}{Z} \qquad Z = \frac{E}{I} \qquad E = I Z$$

$$P = I^2 R \quad \text{or} \quad P = IE \times (pf) \qquad pf = \text{powerfactor} \left(\frac{R}{Z}\right)$$

L = Inductance in HENRIES
C = Capacitance in FARADS
X_C = Capacitive Reactance in OHMS
X_L = Inductive Reactance in OHMS
Z = Impedance in OHMS.

5. Voltage and Current Relationship

RESISTOR - IN-PHASE
CAPACITOR - I LEADS E
INDUCTOR - I LAGS E

6. INDUCTANCE in SERIES

computed as RESISTORS in SERIES.

7. INDUCTANCE in PARALLEL

computed as RESISTORS in PARALLEL.

8. Inductive Reactance

$X_L = 2 \pi f L$ (f = frequency in Hz)

9. REACTANCE of Inductors in SERIES

computed as RESISTORS in SERIES.

10. REACTANCE of Inductors in PARALLEL

computed as RESISTORS in PARALLEL.

11. CAPACITANCE in SERIES
computed as RESISTORS in PARALLEL.

12. CAPACITANCE in PARALLEL
computed as RESISTORS in SERIES.

13. Capacitive Reactance

$X_C = \dfrac{1}{2\pi f C}$ (f = frequency in Hz)

14. REACTANCE of CAPACITORS in SERIES
computed as RESISTORS in SERIES.

15. REACTANCE of CAPACITORS in PARALLEL
computed as RESISTORS in PARALLEL.

16. Impedance

$$Z = \sqrt{R^2 + (X_L - X_C)^2}$$

17. Series and Parallel Resonance

$$X_L = X_C$$

SERIES - Current is Maximum
- Voltage across L or C may be many times line voltage.

PARALLEL - Line Current Minimum
- Circulating current (tank current) is Maximum.

18. Frequency (f)

$$f = \frac{1}{2\pi \sqrt{LC}}$$

19. AC Voltage or Current

Maximum = 1.414 x Effective
Effective = 0.707 x Maximum

20. Transformer Relationships

$$\frac{N_P}{N_S} = \frac{E_P}{E_S} = \frac{I_S}{I_P} = \sqrt{\frac{Z_P}{Z_S}}$$

21. Efficiency

$$\% \text{ EFF} = \frac{\text{OUTPUT}}{\text{INPUT}}$$

MAGNETIC DESIGN FORMULAS

22. Effective Core Parameters

$$C_1 = \sum \frac{\ell}{A} \ (\text{cm}^{-1}) \qquad C_2 = \sum \frac{\ell}{A^2} \ (\text{cm}^{-3}) \qquad \ell_e = \frac{C_1^{\,2}}{C_2} \ (\text{cm})$$

$$A_e = \frac{C_1}{C_2} \ (\text{cm}^2) \qquad V_e = \frac{C_1^{\,3}}{C_2^{\,2}} \ (\text{cm}^3)$$

23. Flux Density (Peak)

$$\hat{B} = \frac{E\, 10^8}{4.44\, f\, N\, A_e\ *} \ (\text{Gauss})$$

*for a nonuniform core set substitute A_{min} for A_e.

Where:

E = RMS sine wave voltage (V)

f = frequency (Hz)

A_e = Effective cross-sectional area (cm^2)

ℓ_e = Effective path length (cm)

I_p = Peak current

N = Number of turns.

24. Field Strength (Peak)

$$\hat{H} = \frac{0.4\, \pi\, N\, I_p}{\ell_e} \ (\text{Oersted})$$

25. Air Core Inductance

$$L_o = \frac{4\, \pi\, N^2\, 10^{-9}}{C_1} \ (\text{Henries}) \qquad C_1 \text{ in cm}^{-1}$$

26. Inductance

$$L = N^2 A_L \quad \text{(nanoHenries)} \qquad L = \mu_1 \frac{4\pi N^2}{C_1} 10^{-9} \quad \text{(Henries)}$$

$$L = \mu_e \frac{4\pi N^2}{C_1} 10^{-9} \quad \text{(Henries)} \qquad C_1 \text{ in cm}^{-1}$$

27. Number of Turns

$$N = \sqrt{\frac{L\, 10^9}{A_L}} \quad L \text{ in Henries}$$

28. Effective Permeability

$$\mu_e = \frac{\ell_e}{\ell_e / \mu_i + \ell}$$

Where: ℓ_e = Effective path length
ℓ = Air gap length.

29. Attenuation

$$20 \log_{10} \frac{| Z_s + Z_L + Z_{sc} |}{| Z_s + Z_L |} \quad \text{(dB)}$$

Where: Z_s = Source impedance
Z_L = Load impedance
Z_{sc} = Suppression core impedance.

30. Quality Factor

$$Q = \frac{2\pi f L_s}{R_s} = \frac{R_p}{2\pi f L_p}$$

Ferrite Material Constants

Ferrite Material Constants*

Specific Heat	0.25 cal/g/°C
Thermal conductivity	$10\ \ 10^{-3}$ cal/sec/cm/°C
Coefficient of Linear Expansion	$8 - 10\ \ 10^{-6}$/°C
Tensile Strength	$7\ \ 10^{3}$ lbs/in^{2}
Compressive Strength	$60\ \ 10^{3}$ lbs/in^{2}
Young Modulus	$18\ \ 10^{3}$ lbs/in^{2}
Hardness (Knoop)	650
Specific Gravity	≈ 4.7 g/cm^{3}

* The above quoted properties are typical for MnZn and NiZn ferrites.

Conversion Tables

SI Units	**CGS Units**
1 T (Tesla) = 1 Vs/m^{2}	= 10^{4} Gauss
1 mT	= 10 Gauss
1 A/m = 10^{-2} A/cm	= 0.01257 Oersted
0.1 mT	= 1 Gauss
79.55 (80) A/m	= 1 Oersted

Multiply number of:	by:	to obtain number of:
Oersteds	79.5	Ampere-turns/m
Oersteds	0.795	Ampere-turns/cm
Gausses	10^{-4}	Teslas (Webers/m^{2})
Gausses	0.1	milliTeslas
in^{2}	6.452	cm^{2}
circular mils	5.07×10^{-6}	cm^{2}
mWatts/cm^{3}	0.094	watts/lb.

Metric Prefixes

Pico - x 10^{-12}
Nano - x 10^{-9}
Micro - x 10^{-6}
Milli - x 10^{-3}
Deci - x 10^{-1}

Tera - x 10^{12}
Giga - x 10^{9}
Mega - x 10^{6}
Kilo - x 10^{3}
Deca - x 10^{1}

Symbols

A_E - Effective cross sectional core area
A_L - Inductance factor (nanoHenries per turn squared)
a_i - Area of insulating system
a_l - Winding area of one layer
a_m - Winding area of m^{th} winding (m = 1,2,3 etc.)
a_t - Total winding area
a_w - Available winding area
B - Maximum flux density (Gauss)
D - Duty cycle
D_I - Duty cycle of current ramp
E-T -Volt-time product (Volt-Seconds)
g - Air gap (per leg)
I - Peak current
I_{av} - Average current
L - Inductance
l - Layers
N - Turns
N_p - Primary turn count
N_s - Secondary turn count
n_l - Turns per layer
P_L - Power loss
R_{AC} - AC Resistance
R_{DC} - DC Resistance
ΔT - Temperature rise
V_D - Output rectifier and choke voltage drop
V_L - Input rectified line voltage
V_S - Output voltage

Index

Notes

There is more to passive component selection than just specifying the electrical ratings. This is the one and only reference book that has answers to **ALL** of your daily application questions. This book tells you what the equations don't specify.

This handbook is one in a series of technical books. The Passive Trilogy is comprised of:

The Capacitor Handbook
The Inductor Handbook
The Resistor Handbook

* Adopted by Universities and Colleges in their Electrical Engineering, Physics, and Electronics Degree Programs. As stated by a College Electronics Technology Coordinator:
"This trilogy fills that gap in academics where the theory ends and the applications of actual components begins. It completes one's education on this subject."

* Used by General Motors Instructors and Electrical Engineers.

* Book Reviews by Electrical Engineering Industry magazines say:

" There's a wealth of useful technical information to be found, ..."
"... loads of practical circuit-application information."
"... a handy all-in-one reference ..." Technology Editor, *Electronic Design*

"With its extensive descriptions of the electrical characteristics and applications of a fundamental electronic device, [this book] would be a useful addition to the bookshelf of any engineer, technician, or hobbyist."
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Editor, *Soun*

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