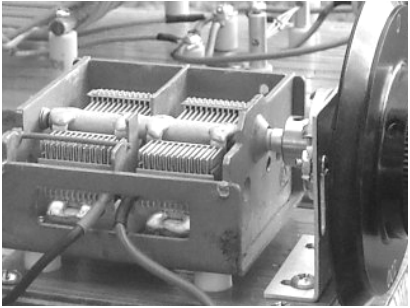
KEVIN'S WEBSURFER HANDBOOK IX
FOR CRYSTAL RADIO

CAPACITOR LOSSES and
WIRE RESISTANCE



Kevin Smith

2013

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http://www.lessmiths.com/~kjsmith/crystal/catalog.shtml

KJ Smith

NOTES:

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INTRODUCTION

This handbook stems from my studies of set resonance, LRC circuits and the quality of the tank components. My measurements on coil Q led me to believe I need to take the other components of the tank more seriously, noteably the capacitor and even the wire from which the coil is wound. In many radio articles I have read the Q-factor of the tank variable capacitor is generally assumed to be very high. High enough as to be disregarded as a significant contributor to the tank Q. Recent discussions on the Radio Board suggest that this may not be the case but in fact contribute significantly to tank resistive losses.

The first several chapters of this handbook then address capacitors, both from a general standpoint and then from the question of capacitor losses. Several articles on measurement techniques are included although it is noted that the capacitors they address appear to botton out at the 1uF range and do not include the smaller MW band picofarad world. Also they seem to be aimed at electrolyte caps. Still, this should stimulate the gray matter and suggest possible attack angles on the problem for crystal radio.

Next I address the question of wire resistance, primarily with a couple handy tables of wire data. High-End coils are wound from litz wire and so this was a partictular focus as I do not know much about this type of wire. A good page on litz wire resistance and data required to make your own calculations should help greatly, followed by papers on general litz considerations. With this edition I hope I will have covered most technical/theoritical aspects of crystal radio and can get back to the happy job of design and construction, this time with a more thorough understanding of what the heck I am up to.

Of general note, the web is a marvelous source of data and information. Many long-time crystal set builders, and many others have created dedicated sites to disseminate information and resources, to share their creations and knowledge. I am eternally in your debt. All of the material in this handbook is copyright for which I have not sought permission. Therefore this is not presented for publication or copy. It is only my personal resource. I encourage anyone finding this copy to pursue ON THE WEB the web pages identified within. I include the name of the author and web address of each section. I wish to sincerely thank every author presented for their excellent pages and ask forgiveness for my editing into this handbook.

Kevin Smith
2012

www.lessmiths.com/~kjsmith/crystal/cr0intro.shtml

to adopt wires of Litz construction to reduce the change in internal inductance and resistance as a function as frequency. However it would still probably be good advice to ensure that each wire bundle was of reasonably large diameter, and the bundles were closely spaced.

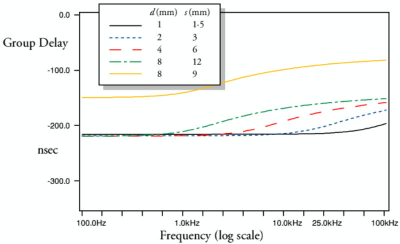


Figure 14 Delays as affected by wire diameter

Figure 14 plots the group delays for the same cables as were used as examples for Figure 13. Looking at these it can be seen that for the examples which share the same value the smallest diameter tends to push the variation in delay as a function of frequency up into the ultrasonic range. Hence a small diameter might be desired if this result is wished. However, by using the thickest wires with a closer spacing we obtain the result shown with the orange line. This reduces the overall level of delay.

In practice it is questionable whether delays of the magnitudes shown would ever be audible. If so, the general advice would seem to be to choose reasonable large diameter wires with a close spacing in order to minimise the effects of resistance and inductance. Since the characteristic impedance of the cables tend to be low compared to those of a typical load the cable series resistance and inductance seem to have far more effect than cable capacitance. This being so, it is the cable resistance and inductance which we should regard as the primary threat to reliable signal transfer. If the change in group delay over the audio band is felt to be significant, a suitable remedy would be

CONTENTS:

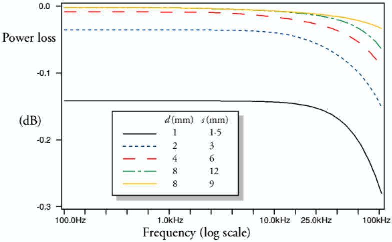


Figure 13 Signal loss as a function of wire diameter

Having examined the effect of varying the wire spacing we can now look at the effect of the choice of wire diameters. Figure 13 shows the power loss as a function of frequency for our standard twin feeder, but using various wire diameters and spacings. Four of the plotted lines have the same wire diameter (d) to spacing (s) ratio. This is to ensure that they all actually have the same characteristic impedance and hence the same nominal inductance and capacitance per length. The variations between these four examples are therefore only due to changes in the internal resistance and inductance of the wires.

Comparing the four plots which share the same value we can see that the lowest losses occur when the wire diameter (and hence spacing) is largest. This seems to clash with the desire to choose a low wire spacing. However the final plot (shown as a solid orange line) is for the largest wire diameter used for the previous four examples, but with a much smaller spacing. This reduces the inductance and hence reduces the level of high frequency power loss.

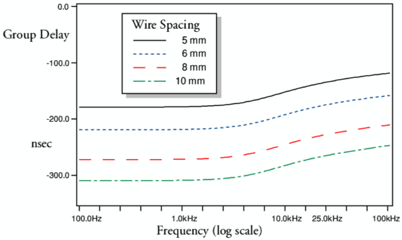


Figure 12 Delays as affected by wire spacing

Figure 12 shows the time delay plots for the same set of spacings, etc, as used for Figure 11. As we would expect, as the spacing is increased and the cable inductance rises, the delay also rises. From the results shown in Figures 11 and 12 we can reasonably expect that we can generally hope to reduce the high frequency losses and overall group delay by choosing a small wire spacing and minimising the cable inductance.

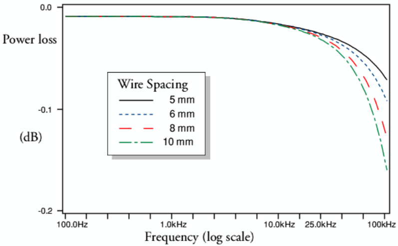


Figure 11 Signal loss as a function of wire spacing

Figure 11 shows the effect of various choices of wire spacing, with a common wire diameter of 4mm (i.e. as previously, a wire radius of 2mm). Note that this has a different vertical scale to earlier power/frequency plots to accommodate the greater losses produced by the wider-spaced cables. Looking at the results we can see that the effect of moving the wires apart is to increase the relative loss at high frequencies. This is simply a consequence of the increase in wire spacing producing a corresponding rise in the cable's inductance per unit length. The combination of the cable inductance and the load resistance act as a conventional low-pass filter. Since the wire diameter is the same for all the lines shown in figure 11 they all have the same value of wire internal impedance.

Secondly, the above assumes we have maintained the diameter of the wire when changing to a lower conductivity material. As we can see from figure 9 this would produce quite high signal losses, and would make the response more sensitive to any frequency dependence of the load impedance.

Thirdly, we can interpret the reduction in the dependence of the delay upon frequency as being due to the relatively high cable resistance 'swamping' the inductance. i.e. the wire resistance is now so high that the cable/load system begins to look more like a simple resistive potential divider with a relatively negligable inductance.

In practice we would probably choose to increase the wire diameter to keep down the overall level of wire resistance. This increase in wire diameter would also increase the value at any frequency, thus lowering the frequency at which phase changes may occur.

7. Effects of wire diameter and spacing

For all the examples considered so far we have used the same values of wire diameter and spacing. This has allowed us to compare various choices of conductor, etc, on a 'like for like' basis. In practice, however, one of the most common types of variation from cable to cable is in the choice of the diameters of the wires, and the spacing between them. We can now use a convenient standard material (Copper) and explore the effects of varying the spacing and diameter of the wires. As before we will assume a 3 metre length of twin-feeder, and an 8 Ohm resistive load as the standard for the purposes of comparisons.

Capacitor Theory
https://ece.uwaterloo.ca/~lab100/ls1notes.pdf

Any arrangement of two conductors separated by an electric insulator (i.e., dielectric) is a capacitor. An electric charge deposited on one of the conductors induces an equal charge of opposite polarity on the other conductor. As a result, an electric field exists between the two conductor surfaces and there is a potential difference between them. The electric field anywhere between the conductor surfaces is directly proportional to the magnitude of the charge Q on the conductors. And the potential difference V is also directly proportional to the charge Q. The ratio Q/V is thus a constant for any electric field distribution as determined by the shape of the conductors, the distance of separation, and the dielectric in which the field exists.

The ratio Q/V is called the capacitance, C, of a particular arrangement of conductors and dielectric. Thus, C = Q/V, where Q and V are in units of coulomb and volt. C has the units farad (F).

$$C = \frac{\varepsilon_r \quad \varepsilon_o \quad A}{d} \quad (F) \qquad (farads)$$

The simple theoretical expression for the capacitance value of a parallel plate capacitor is where
A = plate area [m2] = cross section of electric field,
d = distance between plates [m],
Eo = permittivity of free space = 8.854 x 10-12 F/m and
Er = relative permittivity of the dielectric between the plates [dimension less].

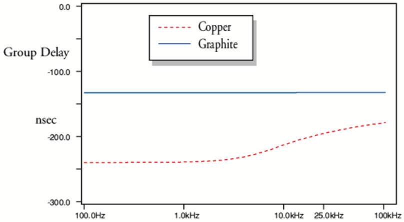


Figure 10 Delays of Graphite and Copper

Figure 10 shows the phase delay behaviour of graphite and copper. In this case the behaviours of the two materials are clearly very different. At low frequencies the inductance of the wires is not dependent upon the resistance level. Hence we do not see the ‘scaling’ effect we encountered when looking at signal losses which there tends to offset the reduction in at any frequency produced by the lower conductivity. From figure 10 therefore we can say that the variation in phase delay with frequency in the audio band is much smaller with graphite than with copper. The graphite also shows a lower overall delay as well as better uniformity. Hence low conductivity materials may be useful if we wish to reduce differential phase/time effects.

However we must draw such a deduction with care for three reasons.

Firstly, the phase changes at audio frequency are very small, even using copper. It is therefore questionable that they would be audible.

$$\frac{100R_0}{100f_0} = \frac{R_0}{f_0}$$

i.e the same in both cases!

This example is not exact and the behaviour is more complex than described here. However this example shows why it turns out that both materials show a noticeable change in signal loss at much the same frequency. Hence for a given wire diameter changing the choice of material has much less effect upon the frequency dependence of the signal loss than we might expect. In fact, we have already seen this to a lesser extent. If we examine the earlier results for fill factor we can see that the main effect of a fill factor of less than unity is to displace the loss curve to a higher or lower overall loss, not to changes its shape.

Using the same wire diameter, etc, as before, do see a marked increase in the low frequency loss due to the lower conductivity. This is simply due to the wires having a much higher d.c. resistance when we change to a material like graphite which has a relatively low conductivity. In practice the level of loss in the example chosen here (around -5 dB) would be far too high for a loudspeaker cable. When using such a lower conductivity material we would therefore probably choose to employ wires of much greater diameter (and hence cross-sectional area) to ensure a low level of loss. However it should be noted that this will mean a change in the value at any given frequency and alter the cable's behaviour in other ways.

This calculated value is based on the assumption that the charge density on the plates is uniformly distributed. In practice there is always a concentration of charge along the edges. This charge concentration is at the sharp corners of the plates. Thus for a given voltage, the actual total charge is always greater than the theoretical total charge.

Practical Capacitor Model

The low frequency lossy model is introduced to allow you to measure capacitance with a
capacitance measuring bridge instrument.

Discrete component measurements and models of the physical capacitor need to be considered here. Consideration of a finite subdivision of the physical device may assist in arriving at a lumped discrete equivalent circuit model of the practical physical device (ie: a capacitor).

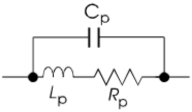


Figure 1.
Capacitor Model
With Parasitic L & R

First consider the physical device, the capacitor. We have two conducting surfaces separated by some distance. Hence a defined capacitor . Assume there is a voltage applied to the capacitor. The dielectric has some loss1 and will conduct a small urrent. One can consider this as a distributed loss within the length of the dielectric material between the two conductors. Hence a resistance in parallel with the capacitor.The lossy dielectric has length. Hence one could conceive a distributed inductance2 in series with the distributed resistance. Hence the inductor as shown in figure 1.

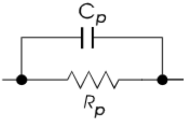


Figure 2. Capacitor Model With R Only

At very low frequencies (say 1 kHz) the parasitic inductance can be ignored3 for the parallel plate capacitor you will be using. Hence the model will only have the capacitor and resistor in parallel as seen in figure 2.

In this experiment the parallel model resistance is very large. Hence the dissipation of energy is small. The resistor will dissipate very little energy. Thus we will have a capacitor with a low dissipation factor, D. One can calculate the resistance value knowing the value of D measured after balancing the bridge.

The capacitance measuring bridge has to be adjusted so that both the capacitance and the resistance of the practical model are balanced. This is the reason for adjusting both the CRL and the DQ dials of the Capacitance Measuring Bridge.

1) The dielectric has a finite resistivity, so when an electric field is across it, a small but finite electric current flows through it. This current, or charge flow, is referred to as a loss (from what an ideal device would do).
2) Because the plates of the capacitor are conductors with current flowing through them, they have a small amount of self inductance.

around 0·05dB greater than the loss at d.c. This is lower than the the drop of around 0·1dB we would obtain using materials like copper, but the improvement is surprisingly small given the factor of 1000 in conductivity.

To understand this behaviour we can consider an example where we have two rods of material which have the same diameters and lengths. One material has a conductivity which is 100 times lower than the other. As a result the end-to-end d.c. resistances of the rods also differ by a factor of 100. The frequencies at which the a.c. resistances of the rods double due to the effects of finite skin depth (internal impedance) also differ by a factor of 100. We can now define a specific value of the ratio of skin depth to diameter (and hence) which is the value for which the a.c. resistance is double the d.c. resistance. For the rod of lower conductivity material this value of occurs at a frequency 100 times higher than for the rod of higher conductivity material.

As a result of the above we would expect the effect of skin effect to be shifted up in frequency by a factor of 100 when we change from the higher to the lower conductivity rod. However, we must then also take into account another difference which turns out to be significant. From the above we can say that the higher conductivity rod can be said to have a d.c. resistance, , which increases to at a frequency we can label . The lower conductivity rod will have a d.c. resistance of which rises to at a frequency . Another way of looking at what happens is therefore that the higher conductivity rod changes its resistance by an amount when we change the frequency by an amount . Whereas the lower conductivity rod changes resistance by when we change the frequency by . This means that in each case the overall rate of change of resistance as we change the frequency will be

6. Effect of low conductivity material

Having noticed the effect of the relatively low conductivity of aluminium some manufacturers and writers have argued that it would be better to deliberately use a material with a particularly low conductivity. Carbon (graphite) has frequently been mentioned in this context. To estimate the effect of using such a material we can now use figures 9 and 10. These show the relative loss and phase delay for the two materials and our standard example of cable.

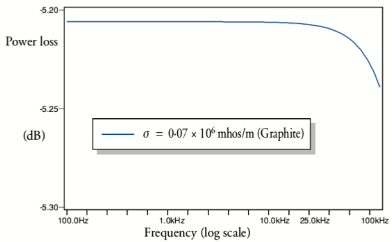


Figure 9 Signal loss for Graphite

Figure 9 shows a result which at first examination seems strange. The graphite has a conductivity around three orders of magnitude lower than that of copper. The skin depth in graphite at any frequency is therefore around 30 times greater. For a given wire diameter this might lead us to expect any increase in loss as the frequency rises to be shifted up to much higher frequencies in graphite than in copper. Yet figure 9 does not seem to show this. The increase in signal loss at 100kHz is

3) The change in current and direction is so slow, the self inductance effect is almost too small to measure.

http://www.avx.com/docs/Catalogs/cbasic.pdf

Basic Capacitor Formulas

AVX

I. Capacitance (farads)

English: $C = \frac{.224\,K\,A}{T_D}$

Metric: $C = \frac{.0884\,K\,A}{T_D}$

II. Energy stored in capacitors (Joules, watt - sec)

$E = \frac{1}{2} CV^2$

III. Linear charge of a capacitor (Amperes)

$I = C \frac{dV}{dt}$

IV. Total Impedance of a capacitor (ohms)

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

V. Capacitive Reactance (ohms)

$x_C = \frac{1}{2 \pi fC}$

VI. Inductive Reactance (ohms)

$x_L = 2 \pi fL$

VII. Phase Angles:

Ideal Capacitors: Current leads voltage 90°
Ideal Inductors: Current lags voltage 90°
Ideal Resistors: Current in phase with voltage

VIII. Dissipation Factor (%)

D.F.= tan δ (loss angle) = $\frac{E.S.R.}{X_C}$ = (2 πfC) (E.S.R.)

IX. Power Factor (%)

P.F. = Sine δ (loss angle) = Cos ϕ (phase angle)
P.F. = (when less than 10%) = DF

X. Quality Factor (dimensionless)

Q = Cotan δ (loss angle) = $\frac{1}{D.F.}$

XI. Equivalent Series Resistance (ohms)

E.S.R. = (D.F.) (Xc) = (D.F.) / (2 π fC)

XII. Power Loss (watts)

Power Loss = $(2 \pi fCV^2)$ (D.F.)

XIII. KVA (Kilowatts)

$KVA = 2 \pi fCV^2 \times 10^{-3}$

XIV. Temperature Characteristic (ppm/°C)

$T.C. = \frac{Ct - C_{25}}{C_{25}(T_t - 25)} \times 10^6$

XV. Cap Drift (%)

$C.D. = \frac{C_1 - C_2}{C_1} \times 100$

XVI. Reliability of Ceramic Capacitors

$\frac{L_o}{L_t} = \left(\frac{V_t}{V_o}\right)^X \left(\frac{T_t}{T_o}\right)^Y$

XVII. Capacitors in Series (current the same)

Any Number: $\frac{1}{C_T} = \frac{1}{C_1} + \frac{1}{C_2} \cdots \frac{1}{C_N}$

Two: $C_T = \frac{C_1 C_2}{C_1 + C_2}$

XVIII. Capacitors in Parallel (voltage the same)

$C_T = C_1 + C_2 \cdots + C_N$

XIX. Aging Rate

A.R. = %Δ C/decade of time

XX. Decibels

$db = 20 \log \frac{V_1}{V_2}$

METRIC PREFIXES

Pico	$X\,10^{-12}$
Nano	$X\,10^{-9}$
Micro	$X\,10^{-6}$
Milli	$X\,10^{-3}$
Deci	$X\,10^{-1}$
Deca	$X\,10^{+1}$
Kilo	$X\,10^{+3}$
Mega	$X\,10^{+6}$
Giga	$X\,10^{+9}$
Tera	$X\,10^{+12}$

SYMBOLS

K	= Dielectric Constant	f	= frequency	L_t	= Test life
A	= Area	L	= Inductance	V_t	= Test voltage
T_D	= Dielectric thickness	δ	= Loss angle	V_o	= Operating voltage
V	= Voltage	ϕ	= Phase angle	T_t	= Test temperature
t	= time	X & Y	= exponent effect of voltage and temp.	T_o	= Operating temperature
R_S	= Series Resistance	L_o	= Operating life		

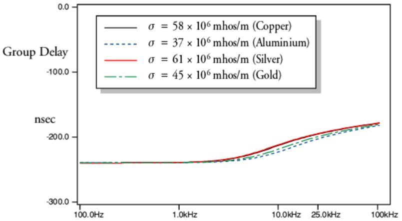


Figure 8 Effect of material upon delay

As with variations in fill factor, the variations between the materials used as examples here seem quite small. Hence it is not obvious that changing, say, from copper wires to silver ones of the same diameter and spacing is likely to produce an audible change as a result of differences in the internal impedance of the wires. Curiously, although aluminium is generally regarded as a 'poor' choice for wiring in audio systems, its relatively low conductivity does mean that the effects of internal impedance tend to occur at slightly higher frequencies than copper.

It is interesting to note that the nominal behaviours of copper and silver are quite similar compared to their difference from aluminium or gold. However it is worth bearing in mind that most of the metals used for wires are far from being of elemental purity. This means that the actual conductivity of a wire may differ by perhaps 5% or more from the standard values assumed for the above calculations. This in itself is unlikely to be significant compared to the much larger range in conductivities for the materials used here as examples.

5. Effect of material conductivity

We can now examine the effect of various choices of conductive material. Figure 7 illustrates the effect of using four different choices of metal for the wires. The same wire diameter and spacing is assumed as previously, and the fill factor is assumed to be unity.

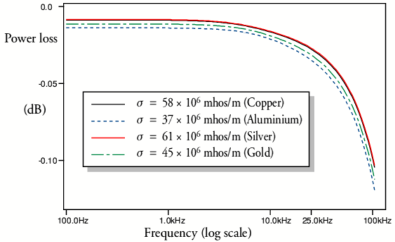


Figure 7 Effect of material upon loss

Figure 8 shows the related results for signal delay values.

What is the best variable capacitor for my crystal radio?
By Dave Schmarder
http://makearadio.com/misc-stuff/capacitors.php

This is a question that has been asked of me many times. It would be easy for me to just say, "Just buy some holy grail capacitors.". Unfortunately, they are not so common anymore, and when you see how much they cost at the large auction venue, you'll forget about the high cost of the litz wire that goes with them.

The purpose of this paper is to discuss various types of variable capacitors, showing their good and bad points. I have come up with some answers that I have not seen discussed elsewhere. Perhaps this work will help you find the right variable capacitor for your radio.

This paper deals nearly exclusively with a crystal radio detector unit in a two section radio. Discussion of an antenna tuner section can be found on my antenna tuner page.

Before we get down to the nitty-gritty brass tacks of the actual capacitors that you should get, let's set the basics for this page. First, anyone that wants to build a dx set should visit the crystal radio DX'ers best friend, Ben Tongue's website. Specifically his paper dealing with variable capacitors as used in crystal sets. In his article 24, section B1, his charts plainly show the difference between a holy grail and an everyday capacitor.

One even wonders how a crystal set works with such a lousy capacitor. Well, it really isn't that bad. But with losses highest at the high end of the band, plus the very squeezed tuning (10% of the capacitor rotation covers 30 to 40% of the

Looking at figure 5 we can see that changing the fill factor from being unity down to 0·7 has a surprisingly modest effect upon the signal loss behaviour.

Figure 6 shows the delay behaviour for the same set of fill factor values.

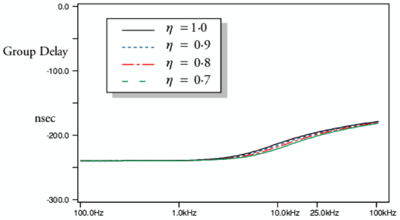


Figure 6 Effect of fill factor upon delay

Again, the changes are relatively modest. On the basis of these results it does not seem as if the fill factor value of stranded wires is likely to have a noticeable effect for twin feeders of this type. That said, it may be more significant for wire diameters and spacing other than those used for this example. More generally, however, it seems that a Litz construction (with the strands individually stranded) is probably usually required for multistranded wires to have audio properties that differ noticeably from solid wires. That said, multistrand is often physically more flexible, and may be easier to connect, so may be more convenient than solid-core for other practical reasons.

Again it is not clear that this would be noticeable in a practical situation.

4. Effects of fill-factor

For the sake of example, the cables considered until now were assumed to be made of copper with an effective fill factor value of unity. To assess the typical effects of a change in fill factor, or of the chosen conductive material, we can recalculate the behaviour of a cable similar to that used before, but with various fill values or materials. As before we therefore assume a 3 metre length of twin-feeder with 2mm diameter wires spaced with their centers 6mm apart.

Figure 5 shows the power loss for four differing values of the fill factor, using copper wires.

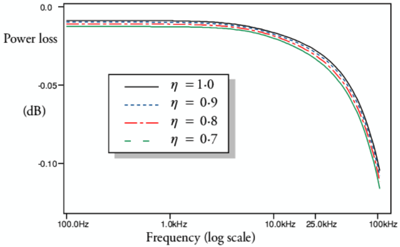


Figure 5 Effect of fill factor upon power loss

frequency range), there is a lot of bad karma at the high end of the band.

Below are some examples of capacitor traits. They are not in any particular order of importance or the goodness of one capacitor over another.

A variable capacitor is made up of two major sections, a stator or stationary part, and a rotor or moveable section. These sections are made with plates, also called vanes or blades. The frame is usually connected to the rotor and is one connection of the capacitor. There are some instances where the rotor is isolated from the frame, but not too often.

Dielectric

This is the material that separates the stator plates from the rotor plates. An air dielectric is most common, and is very low loss. Air has a standard dielectric constant of 1.0. Sometimes other materials are placed between the rotor and stator plates, mainly so the plates can be closer together. This allows for a smaller capacitor size, but it is at the expense of higher losses. I like using air capacitors the best.

Insulators

Insulators are the non-conducting material that holds the stator plates to the capacitor frame. The insulator material is considered by most builders the prime criteria in selecting a good capacitor. Two major types of insulating materials used are ceramic and phenolic. Ceramic is consider the very best and the only thing to use in a dx crystal set.

However, new ceramic insulated variable capacitors are expensive. For less aggressive radios, and very aggressive budgets, phenolic is the choice.

Plate Plating

Some of the variable capacitors that you see will have a silver plating. Ben warned us in his article that what appears to be silver plating might not be. The new capacitors sold have bare aluminum plates. They are higher resistance than the silvered plates. Of all the capacitor spec's, this is the least important. But most of the older capacitors that have ceramic insulators will also have a better plating.

Wiper Arms

The wiper arms are those little springy thin pieces of metal that attach to the frame and contact the rotor. Variable capacitors are rated by the number of wipers and how much surface area is in contact. The tightness of the contact should also be considered. Tight is good.

Inexpensive capacitors sometimes have no wiper arms. While they do work, their use in a good crystal set is discouraged. Other cheaper capacitors have just one small wiper at the back frame. The contact area is small, and the contact pressure is fairly light. These are ok for TPT sets. (Radios that are low performance because the builder uses coil forms made with cardboard Toilet Paper Tubes.)

The maximum number of wipers is determined by how many frame surfaces are reachable. In a single gang capacitor, two, one at the front and one at the back is the maximum. In a two gang, it's four (front, back and two in the middle), and so on.

Most hobbyists don't consider the wiper arm quality when they are looking for the capacitor for their dream radio. I consider wipers to be almost as important as the insulator material. I have tested phenolic insulator capacitors that had very nice

$$v = \frac{c}{\sqrt{\varepsilon_r}}$$

where is the velocity of light in vacuo, and is the relative permittivity. Here so we would expect a velocity of around 2·44 × 108 m/s. Hence we might expect a propagation delay when using 3 metre cable of only around 12 - 13 nanoseconds. Yet we see values over ten times greater than this.

The reason for this discrepancy is that microwave/RF engineers usually arrange to match the source and load impedances to the cable's characteristic impedance. This is far from being the case here. The load resistance is 8 Ohms, but the cable has a characteristic impedance of around 75 Ohms. Since the cable is also very 'short' in terms of the signal wavelengths it primarily acts as an inductor and resistor placed in series with the 'low' load impedance. At low frequencies the signal delay is therefore approximately equal to which – ignoring internal effects – in the example equals around 164 nanoseconds. Hence the group delays we see are in broad agreement for what we would expect as a result of cable inductance in this example. The dashed line (equivalent to Litz wire) shows a value of 164 nanoseconds as internal inductance has been suppressed. This indicates that although the solid conductor (or a similar stranded wire) have a value of around 240 nanoseconds at low frequencies only around 80 nanoseconds of this is due to the wire's internal inductance.

Despite the above, to an acoustic engineer the delays seem very small. Since sound propagates at a velocity of around 330 m/sec it will only travel around 16 microns in 50 nanoseconds. In effect, therefore, the above differential group delay is equivalent to a sound source that seems 16 microns nearer at high audible frequencies than at low audible frequencies.

lines we can see that the internal impedance causes the group delay to depart from being uniform. If no ‘skin effect’ were present the group delay would be almost steady at 164 nanoseconds across the whole band plotted. However when the internal impedance of the wires are taken into account the group delay varies from around 240 nanoseconds at low frequency, to around 212 nanoseconds at 10kHz and around 195 nanoseconds at 25 kHz. The effect of the internal impedance is therefore to produce a differential group delay of around 50 nanoseconds across the audio band. Interestingly, high frequencies will tend to arrive slightly earlier than low frequencies.

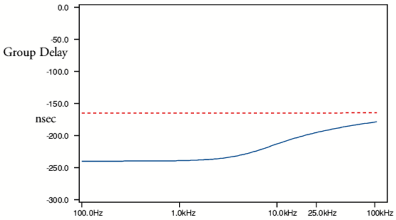


Figure 4 Group delay versus frequency

Curiously, the above time delay values can be seen as ‘large’ or ‘small’ depending upon your starting point. To a microwave/RF engineer with experience of transmission lines a delay of around 200 nanoseconds seems oddly large for a twin feeder only three metres long. Since we are assuming that no magnetic materials are present and a twin feeder propagates a TEM wave we might expect a propagation velocity of

wipers and while they didn't test as well as ceramic types, they were still very decent.

Capacitance Minimum to Maximum Ratio

Here is something that almost nobody considers when buying their capacitors, until the builder tries to tune the entire band with a low maximum to minimum capacitance ratio and finds out that it doesn't work. The US-Canadian broadcast band is from 530 to 1710 kHz (while the US is 540 - 1710 and Europe only goes to 1606 kHz). This means that if you are going to tune the entire band, without switching in additional capacitance or with coil taps (yeech), close attention has to be paid to the maximum to minimum capacitance ratio.

An example of a capacitor that will not tune the entire MW band in one swoop is a capacitor that is available on the surplus market. This capacitor has everything that one would want in a capacitor, except a good maximum to minimum capacitance ratio! A picture of it is shown below. Each section measures from about 300 to 35 pF. This is ratio of about 8.6:1. The medium wave band at my location needs a total ratio of at least 20:1 if it is expected to tune without band switching. This includes the distributed capacitance of the coil, and the diode input. At 20:1 there is very little leeway and the coil has to be constructed with extra care. The cheap capacitors fall well within this range, as they have a ratio of about 24:1.

The way you can tell about a capacitors ratio at a glance is to look at the plate structure. If you see a many plates and they are small, this means that the ratio is low. This is because when completely open, the capacitor still has a fair amount of surface between the plates.

Capacitor Frequency Linearity

Those of you that have tuned a radio using a capacitance linear capacitor such as the inexpensive ones mentioned earlier, noticed the top of the band dial squeeze. This is because the capacitance is linear to the rotation. If the capacitor shaft is rotated half way, the capacitance is approximately at half the total value. This means that the center of the dial is at around 800 kHz, about 370 kHz out of the total of nearly 1200 kHz that comprises the medium wave broadcast band. The other rotation half has to cover 3/4 of the band. It gets a lot worse the closer you get to the top of the band.

But what if you could move the capacitor 50% and have half of the band covered? This is what is called a straight line frequency capacitor. In practice having a completely straight line frequency capacitor is nearly impossible, but we can come close.

There are three basic types of capacitor construction. One is where the shaft is centered on the capacitor frame, right to left. The capacitor plates are semicircular. This operates as a straight line frequency capacitor (because if you plot the capacitance vs. the rotation, there is a straight line result).

Another type which was very popular in the old superhet tube radio days was with the shaft and plates offset from the center. These types can also have a modified plate shape, which produces a slight "S" curve when the capacitance vs. rotation is plotted. These, mostly dual gang capacitors are widely sold, but not as popular as the inexpensive single gang linear variable capacitors.

You can expect to pay twice as much for these as the inexpensive capacitors, but most have the advantage of a dual wiper, which lowers the losses. The tuning of this capacitor

Figure 3 shows the power loss in the cable as a function of frequency. The solid (blue) line shows the loss when the wire's internal impedance is taken into account. The broken (red) line shows the loss that would occur if we ignored the internal impedance. Hence the difference between the two shows the results in this case of 'skin effect'. In principle, the broken line shows what we might expect when using wires of 'Litz' construction with a fill factor approaching unity. The solid line shows what we might expect from either a solid wire or a multistranded wire of closely packed thin strands.

For the example chosen, at low frequencies the bulk resistance of the copper wire causes a power loss of around 0·008dB. At 10kHz the loss rises to 0·009dB if the internal impedance were absent, and 0·016dB with the internal impedance taken into account. At 25kHz these values rise to 0·012dB and 0·028dB respectively. Hence the change in relative signal level from near-d.c. to 25kHz, with internal effects taken into account is around 0·020dB. This is quite a small change so it is not obvious that it would be noticed in a practical audio system.

Figure 4 shows the group delay as a function of frequency. This measure is more convenient that plotting any change in phase since a propagation delay will manifest itself as a phase change that rises linearly with frequency. When signals have to propagate through a system the group delay is therefore a better indicator of any time domain effects that might alter the signal pattern. A uniform group delay value implies that all components will arrive with the same relationship as when they were transmitted.

As with figure 3, figure 4 shows two two lines, the solid (blue) one taking internal impedance into account, and the broken (red) line neglecting internal impedance. By examining these

For illustration we can use a case where we have copper cables (assume = 5·8 × 107 mhos/metre) which have wire radii of 2mm with their centers 6·5mm apart. We can also take as an example a twin feeder length of 3 metres. (i.e. 6 metres of wire, in total, in the circuit) and assume a nominal load of 8 Ohms. We can also assume a perfect voltage source and set the source resistance to be 0 Ohms. The cables are assumed to be insulated with a dielectric whose relative permittivity value is 1·5. Most practical dielectrics tend to have a higher value, but the effect is diluted since part of the surrounding E-field is in the air, so this value seems a reasonable one to assume for the purposes of an example.

The results obtainable for this situation are indicated in the following graphs. The dimensions chosen for the wires here are not meant to represent any specific cable, but seem typical for a relatively inexpensive form of wiring.

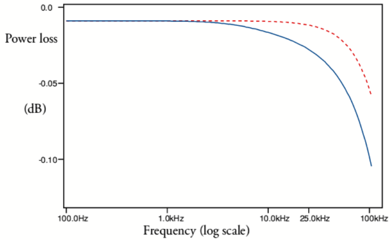


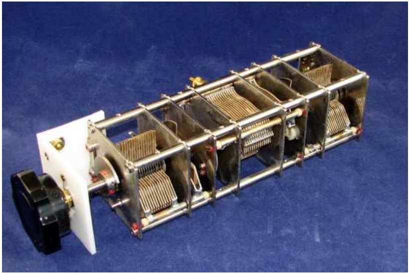
Figure 3 Power loss in transmission versus frequency

does help the high end of the band frequency crunch. It is at least tolerable.

The third type is a closer to a straight line frequency capacitor. This has highly shaped plates plus the offset shaft position. There is a much better frequency spread across the dial, especially at the top end of the band.

Insulator Loss vs. Capacitor Plate Loss Ratio

As Ben so clearly shows in his charts, as the capacitance is increased on a particular variable capacitor, the losses go down. In his example of the cheap capacitor, the losses go down by over a factor of 10 when the shaft is rotated to about 1/3 of the way up. The more expensive capacitor also trends up, but at less steep rate. This is because the losses of the dielectric and wiper arms are summed with the losses between the capacitor plates.
High quality capacitor with a lot capacitance minimum to maximum ratio.



This is the so called "waysalot" capacitor. I used a pair of these in my #75 dx crystal set. It is the best quality capacitor I have. This one has a vernier attached and is used as a test capacitor for determining unloaded Q of the coil under test.
The so called Holy Grail capacitor. The dream of every dx crystal radio builder.

conductor in each triangle is equal to . However the area of each triangle will be . Hence the fill factor will be approximately equal to

$$\frac{\pi}{2\sqrt{3}} \approx 0{\cdot}9069...$$

Real strands will tend to deform slightly when compressed and may not have perfectly smooth, circular cross sections. They may also not be perfectly packed. Hence we can expect the actual value of the fill factor to vary accordingly in practice. In Litz bundles the insulation layers on the strands will also move the conductors apart by a small amount, thus reducing the fill factor value. Usually, we can expect this effect to be small as the layer of insulation is likely to be very thin. In most cases we can therefore tend to assume that the fill factor is reasonably close to unity so this is a reasonable assumption for general analysis. That said, in some specific cases we can take stranding into account by modifying the effective conductivity by an appropriate fill factor value.

3. Twin Feeder Loudspeaker Connections – an example

These internal values can now be used when we wish to evaluate the effect of a cable upon the transmitted signals. We can treat the connecting cables as a form of transmission line, and modify the inductance and resistance per unit length to take the internal values into account. At audio frequencies the cables are normally many orders of magnitude shorter than the highest signal wavelength so it is reasonable to then use lumped values which just multiply the per-unit-length value by the chosen length. For the sake of example here we will consider loudspeaker cables nominally in the form of ‘twin feeder’. For a given cable length this means we must remember to double the values for internal resistance and inductance as a pair of wires are used.

illustrated in figure 2. This shows a close-packed array of strands, each of radius .

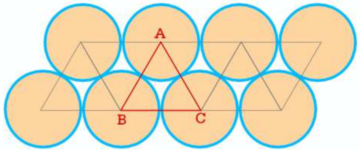
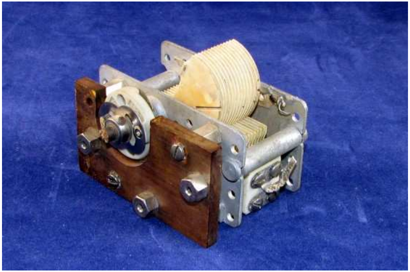


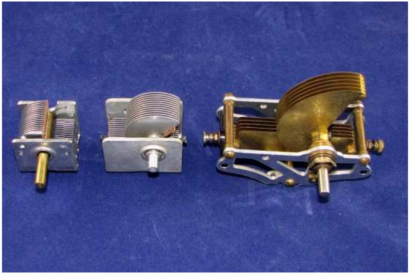
Figure 2 Cross section of a packed bundle of strands.

Since the strands all have a circular cross-section we find that even if they are tightly packed into a hexagonal array there will be spaces in between the places where they touch. Hence a bundle of small packed strands that are in electrical contact can be regarded as similar to a solid but which has some air inclusions which mean that overall wire cross section is only partly filled with conductor. Given a value for the 'fill factor', , (the fraction of the cross section which is filled with conductor) we can treat the wire as being equivalent to a solid conductor of the same wire diameter, , but having an effective conductivity of . Thus the main electronic effect of using a bundle of strands is to dilute the effective conductivity and lower its apparent value.

To estimate the value of the fill factor we can note that the array has a symmetry similar to a packed array of equilateral triangles whose sides all have a length of . Taking one of these elemental triangles we can see that it contains three 60 degree sectors of conductor. Hence the cross sectional area of



This is known as Holy Grail (everyone should bow and say "I'm not worthy, I'm not worthy"). This capacitor is about 475 pF can was removed from a high quality test instrument. If you can find one for less than seventy-five bucks, grab it.
Examples of capacitor styles with special attention to the plate shape.



Here are examples of three different variable capacitors frequency vs. capacitance plots. The one on the left is the inexpensive ones available everywhere. The shaft is centered left to right and the plates have a semicircular shape.

The middle one also has the semicircular shape, but the shaft is offset from the center of the frame and the plates. This offers better frequency linearity than the capacitor on the left.

Finally the capacitor on the right is a great example of a straight line frequency plot. The shaft is offset, and the plates are very long. It is only near the bottom of the tuning range is where a lot of capacitance is added with little turning of the shaft. This allows for a more even frequency spread across the tuning range.

Litz wire (as stranded but with insulation between the individual strands)

For solid-core wires the above analysis can be used immediately to compute the internal impedance and deduce the effects it may have in a given situation. For the stranded and Litz wires we need to take the stranding and the effect of inter-strand contacts or insulation into account.

Litz wires consist of a bundle of very thin, individually insulated conductors. The insulation ensures that the current flows in all of the wires in the bundle as the charge cannot migrate towards the surface of the bundle. The entire cross section of conductor bundle is therefore used by the charge transport. Provided that the individual strands are thin enough, the strands all have individual radii that are small compared to the skin depth at audio frequencies. Hence the overall properties of the Litz bundle tends to be similar to that of a single wire of the same diameter of the bundle but where 'skin effect' is apparently absent.

In practice, most of the multi-strand wires used for audio purposes have no insulation on the individual strands. This means they do not behave like a Litz wire. In stranded wires without insulation between the individual strands charge may cross from strand to strand. Hence current will tend to preferentially flow near the skin of the bundle of wires, just as it does with a single solid conductor of similar overall diameter. Hence when the strands are thin but in electrical contact with their neighbours we can expect the effect of internal impedance to be similar to that of a solid wire of a diameter similar to the bundle of strands. There is, however, one factor we should take into account. This arises when the strands do not 'fill' the bundle and their are air gaps. This is

The standard 'h.f.' approximation is that both and will be essentially equal to . However by inspecting the results shown in figure 1 we can see that it is possible to do better than this and a more reliable approximation would be

$$\frac{R}{R_0} \approx \frac{q}{2\sqrt{2}} + 0{\cdot}26 = \frac{r_0}{2\delta} + 0{\cdot}26 \qquad (9)$$

$$\frac{\omega L}{R_0} \approx \frac{q}{2\sqrt{2}} - 0{\cdot}02 = \frac{r_0}{2\delta} - 0{\cdot}02 \qquad (10)$$

This approximations are show in figure 1 by the broken lines. The approximation for is shown by a short-dashed red line, and that for by a longer-dashed green line. It can be seen that these approximations are likely to be reasonably accurate in the region .

2. Types of Wire

In general, electronic signals are conveyed using a 'pair of wires'. These are used to form a closed loop (path) between the signal source and the load around which charge may flow. Broadly speaking we can then define a 'cable' to consist of a pair of wires. The most common forms of cable used in audio are 'Twin Feeder' and 'Co-ax'. The basic properties of these are discussed elsewhere. Each of the wires may be a single, solid, length of conductor. More usually, however, each wire will consist of a bundle of conducting 'strands'. Multistrand wires have properties which may differ from that of a single, solid wire of similar cross-section. We can therefore treat wires as falling into three general categories as outlined below.

Solid core (i.e. just one strand of conductor per wire)
Stranded wire (a bundle of thin strands of conductors)

Types of capacitor
From Wikipedia
http://en.wikipedia.org/wiki/Types_of_capacitor#Ohmic_losses.2C_ESR.2C_dissipation_factor.2C_and_quality_factor

Theory of conventional construction

Dielectric is placed between two conducting plates (electrodes), each of area A and with a separation of d

In a conventional capacitor, the electric energy is stored statically by charge separation, typically electrons, in an electric field between two electrode plates. The amount of charge stored per unit voltage is essentially a function of the size, the distance and the plates' material properties and the material, the dielectric, placed between the electrodes, while the potential between the plates is limited by the Dielectric breakdown field strength .

Nearly all conventional industrial capacitors except some special styles such as "feed-through capacitors", are constructed as "plate capacitors" even if their electrodes and the dielectric between are wound or rolled to a winding. The capacitance formula for plate capacitors is:

$$C = \frac{\varepsilon A}{d}.$$

The capacitance C for conventional capacitors increases with area A of the electrodes and with permittivity ε of the dielectric material and decreases with the distance d. The capacitance is therefore greatest in devices made from

results of doing this are illustrated in figure 1. These are plotted versus so that the relevant nominal skin thickness is also normalised in terms of the wire radius. The solid lines plotted show the relevant values calculated from the above expressions.

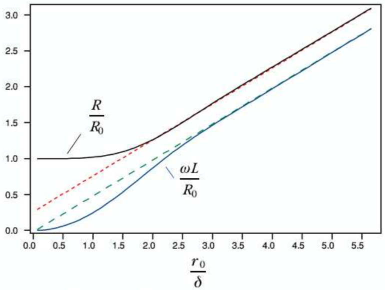


Figure 1 Wire impedance versus skin depth

Unfortunately, the expressions provided by A & S only cover the range for which roughly corresponds to . Above this value the Bessel functions become hard to evaluate and their combination tends to lead to a situation where a set of large values cancel to give a moderate result. Hence for computational simplicity we can use a simpler approximation for the situations where is ‘large’. Here this may be defined as where this ratio has a value greater than 5.

The detailed analysis in Ramo leads to the following expressions which can be used to determine the relevant wire resistance and inductance per unit length for a conductive wire of circular cross section at frequencies above d.c.

$$R = \frac{R_S}{\sqrt{2}\,\pi r_0}\left[\frac{\mathrm{Ber}\{q\}\,\mathrm{Bei}'\{q\} - \mathrm{Bei}\{q\}\,\mathrm{Ber}'\{q\}}{(\mathrm{Ber}'\{q\})^2 + (\mathrm{Bei}'\{q\})^2}\right] \quad (3)$$

$$\omega L = \frac{R_S}{\sqrt{2}\,\pi r_0}\left[\frac{\mathrm{Ber}\{q\}\,\mathrm{Ber}'\{q\} + \mathrm{Bei}\{q\}\,\mathrm{Bei}'\{q\}}{(\mathrm{Ber}'\{q\})^2 + (\mathrm{Bei}'\{q\})^2}\right] \quad (4)$$

where

$$R_s \equiv \frac{1}{\sigma\delta} \quad ; \quad q \equiv \frac{\sqrt{2}\,r_0}{\delta} \quad (5,6)$$

and is the ‘skin depth’ value which may be calculated via

$$\delta \equiv \frac{1}{\sqrt{\pi f \mu\sigma}} \quad (7)$$

where is the signal frequency (as opposed to the signal’s angular frequency).

In fact, using the above expressions we can show that

$$\frac{R_s}{\sqrt{2}\,\pi r_0} \equiv \frac{R_0 q}{2} \quad (8)$$

and specifying this factor in terms of and may be more convenient when performing calculations.

Now, , etc are Bessel functions. We can find numerical expressions for evaluating these in a text like Abramowitz and Stegun. Using these we can compute values. For the sake of clarity is is useful to plot values normalised in terms of . Some

materials with a high permittivity, large plate area and small distance between plates.

Theory of electrochemical construction

Besides the conventional static storage of electric energy in an electric field, two other storage principles to store electric energy in a capacitor exist. They are so-called electrochemical capacitors.in In contrast to ceramic, film and electrolytic capacitors, supercapacitors, also known as electrical double-layer capacitors (EDLC) or ultracapacitors do not have a conventional dielectric. The capacitance value of an electrochemical capacitor is determined by two high-capacity storage principles. These principles are:

* electrostatic storage within Helmholtz double layers achieved on the phase interface between the surface of the electrodes and the electrolyte (double-layer capacitance) and the

* electrochemical storage achieved by a faradaic electron charge-transfer by specifically adsorpted ions with redox reactions (pseudocapacitance)

Double-layer capacitance and pseudocapacitance combine to a form a supercapacitor.

Common capacitors and their names

Capacitors are divided into two mechanical groups: Fixed capacitors with fixed capacitance values and variable capacitors with variable (trimmer) or adjustable (tunable) capacitance values.

The most important group is the fixed capacitors. Many got their names from the dielectric. For a systematic classification these characteristics can't be used, because one of the oldest, the electrolytic capacitor, is named instead by its cathode construction. So the most-used names are simply historical.

The most common kinds of capacitors are:

* Ceramic capacitors have a ceramic dialectric.

* Film and paper capacitors are named for their dielectrics.

* Aluminum, tantalum and niobium electrolytic capacitors are named after the material used as the anode and the construction of the cathode

* Supercapacitor is the family name for:
 * Double-layer capacitors were named for the physical phenomenon of the Helmholtz double-layer
 * Pseudocapacitors were named for their ability to store electric energy electro-chemically with reversible faradaic charge-transfer
 * Hybrid capacitors combine double-layer and pseudocapacitors to increase power density

* Seldom-used Silver mica, glass, silicon, air-gap and vacuum capacitors were named for their dielectric.

Capacitors in each family have similar physical design features, but vary, for example, in the form of the terminals.

misunderstood, and hence people occasionally tend to invoke this frequency dependent behaviour as the implied basis for all kinds of claims regarding the ‘sounds’ of different types of cables. The purpose of this analysis is to throw some light into this area and help provide some understanding of the effects of using conductors of finite conductivity.

In engineering textbooks, the consequences of finite conductivity and wire size are treated in terms of an ‘Internal Impedance’. This term is probably more useful that ‘skin effect’ as it acts as a reminder that the effects arise due to the fields internal to the conductor. The internal impedance per unit length of a wire is considered in Ramo et al. From this was may draw the following results as a starting point.

The d.c. (i.e. very low frequency) impedance of a wire which has a circular cross-section and is uniform may be said to consist of a resistance per unit length of

$$R_0 = \frac{1}{\pi r_0^2 \sigma} \qquad (1)$$

where is the bulk conductivity value appropriate for the material used to manufacture the wire, and is the radius of the wire. The resistance is in Ohms/metre if we are using S.I. units (which will be assumed from now on).

The wire will also exhibit an effective inductance per unit length at very low frequency due to its internal fields. At very low frequencies this has the value

$$L_0 = \frac{\mu}{8\pi} \qquad (2)$$

where is the permeability of the material. In general we can assume that this equals the value for free space

$$\mu_0 = 4\pi \times 10^{-7} \quad \text{Henries/Metre}$$

Skin Effect, Internal Impedence, and types of wire

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http://www.st-andrews.ac.uk/~www_pa/Scots_Guide/audio/skineffect/page1.html

1. Skin Effect and cable impedance

There is a considerable amount of discussion (not to say argument!) amongst those interested in audio and Hi-Fi about the possible effects of cables upon 'sound quality'. This tends to lead to some people adopting almost 'theological' viewpoints that differ fundamentally from the views held by others. One of the technical factors which is sometimes claimed to affect sound quality is what is usually called 'Skin effect'.

In most electronics textbooks, the properties of cables and wires are considered as a form of transmission line. The text may mention briefly the skin effect without exploring this in detail. More often, however, the only parameters that tend to be considered are the capacitance per length, inductance per length, and their relationship with the signal's nominal propagation velocity and the characteristic impedance of the system. The fact that normal materials have a finite conductance (or resistance) is not usually considered beyond its effect upon the d.c. (and low frequency) resistance of the cables and the resulting implied signal power losses.

In reality, when we transmit alternating signals along conductive lines we may experience effects due to what is generally called 'skin effects'. This subject is widely

In addition to the above shown capacitor types, which derived their name from historical development, there are many individual capacitors that have been named based on their application. They include:

* Power capacitors, motor capacitors, DC-link capacitors, suppression capacitors, audio crossover capacitors, lighting ballast capacitors, snubber capacitors, coupling, decoupling or bypassing capacitors.

Often, more than one capacitor family is employed for these applications, e.g. interference suppression can use ceramic capacitors or film capacitors.

Specialized devices such as built-in capacitors with metal conductive areas in different layers of a multi-layer printed circuit board and kludges such as twisting together two pieces of insulated wire also exist.

Dielectrics

The most common dielectrics are:

* Ceramics
* Plastic films
* Oxide layer on metal (Aluminum, Tantalum, Niobium)
* Natural materials like mica, glass, paper, air, vacuum

All of them store their electrical charge statically within an electric field between two (parallel) electrodes.

Beneath this conventional capacitors a family of electrochemical capacitors called Supercapacitors was developed. Supercapacitors don't have a conventional dielectric. They store their electrical charge statically in

* Helmholtz double-layers (Double-layer capacitors)

and additional electrochemical with faradaic charge transfer

* with a pseudocapacitance (Pseudocapacitors)
* or with both storage principles together (Hybrid capacitors).

The most important material parameters of the different dielectrics used and the appr. Helmholtz-layer thickness are given in the table below.

Table of Key parameters

I0(z) = Modified Bessel Function of First kind of order 0
I1(z) = Modified Bessel Function of First kind of order 1

The first mathematical discussions on the skin effect of wires were done by Maxwell, who started modern electromagnetics in 1873. This was followed by significant contributions by the creative genius Heaviside, who followed the theoretical work of Maxwell in 1884-1887, and by Poynting in 1884-1885. Experimental confirmation was first done by Hughes in 1886. Numerical engineering calculations on the applicable level was first done by Lord Rayleigh for the skin effect on infinite planes in 1886, and on cylindrical conductors by Lord Kelvin using ber-bei functions in 1899. Afterwards, there was work by Hertz and J.J. Thomson, who took over experimental work by Maxwell. The first person to use the word "skin effect" is considered to be J.Swinburne in 1891. Until recently, there were continuous challenges by many brilliant theorists and engineers, including H.B.Dwight and A.E.Kennelly.

For your information, the word "Litz wire" comes from the German word "Litzendraht" (twisted strands).

Kouichi Hirabayashi, (C) 2003

"if we increase the number of wires, we can make a cable with less skin effect. If company A has a product with 19 wires, we make one with 30 wires. If they beat us, we will increase the number of wires by one...".

As a consequence, many audio cables with Litz wire were born. Is this strategy effective?

This is the question.

Footnote 1

Obtaining the current distribution of a single cylindrical conductor under high frequency is relatively simple, and the solution is shown below. The current distribution shown in Fig. 1 has been calculated using this equation. However, only the magnitude is shown. Note that the phase of the current also depends on the position. As we get deeper into the conductor from the surface, the phase of the current delays and may even become backward!

Ir / Ia = I0(sqrt(j)*k*r) / I0(sqrt(j)*k*a)
Here,
Ir = current density (A/m^2) at radius r (m)
Ia = current density at outer surface of conductor (A/m^2)
a = conductor radius (m)
j = sqrt(-1)
k = sqrt(w*u*g)
w = angular velocity (rad/s)
= 2*PI*f
PI = 3.14159265..
f = frequency (Hz)
u = permeability (H/m)
g = dielectric constant (S/m)

Capacitor style	Dielectric	Permittivity at 1 kHz	Maximum/realized dielectric strength V/µm	Minimum thickness of the dielectric µm
Ceramic capacitors, Class 1	paraelectric	12–40	< 100(?)	1
Ceramic capacitors, Class 2	ferroelectric	200–14,000	< 25(?)	0.5
Film capacitors	Polypropylene (PP)	2.2	650/450	2.4–3.0
Film capacitors	Polyethylen terephthalate, Polyester (PET)	3.3	580/280	0.7–0.9
Film capacitors	Polyphenylene sulfide (PPS)	3	470/220	1.2
Film capacitors	Polyethylene naphthalate (PEN)	3	500/300	0.9–1.4
Film capacitors	Polytetrafluoroethylene (PTFE)	2	450(?)/250	5.5
Paper capacitors	Paper	3.5–5.5	60	5–10
Aluminium electrolytic capacitors	Aluminium oxide Al_2O_3	9,6	710	< 0.01 (6.3 V) < 0.8 (450 V)
Tantalum electrolytic capacitors	Tantalum pentoxide Ta_2O_5	26	625	< 0.01 (6.3 V) < 0.08 (40 V)
Niobium electrolytic capacitors	Niobium pentoxide, Nb_2O_5	42	455	< 0.01 (6.3 V) < 0.10 (40 V)
Supercapacitors Double-layer capacitors	Helmholtz double-layer	-	-	< 0.001 (2.7 V)
Vacuum capacitors	Vacuum	1	40	-
Air gap capacitors	Air	1	3.3	-
Glass capacitors	Glas	5–10	450	-
Mica capacitors	Mica	5–8	118	4–50

The capacitor's plate area can be adapted to the wanted capacitance value. The permittivity and the dielectric thickness are the determining parameter for capacitors. Ease of processing is also crucial. Thin, mechanically flexible sheets can be wrapped or stacked easily, yielding large designs with high capacitance values. Razor-thin metallized sintered ceramic layers covered with metallized electrodes however, offer the best conditions for the miniaturization of circuits with SMD styles.

A short view to the figures in the table above gives the explanation for some simple facts:

* Supercapacitors have the highest capacitance density because of its special charge storage principles
* Electrolytic capacitors have lesser capacitance density than supercapacitors but the the highest capacitance density of conventional capacitors because its thin dielectric.
* Ceramic capacitors class 2 have much higher capacitance values in a given case than class 1 capacitors because of their much higher permittivity.
* Film capacitors with their different plastic film material do have a small spread in the dimensions for a given capacitance/voltage value of a film capacitor because the minimum dielectric film thickness differs between the different film materials.

Ceramic capacitors

A ceramic capacitor is a non-polarized fixed capacitor made out of two or more alternating layers of ceramic and metal in which the ceramic material acts as the dielectric and the metal acts as the electrodes. The ceramic material is a mixture of finely ground granules of paraelectric or ferroelectric materials, modified by mixed oxides that are necessary to achieve the capacitor's desired characteristics. The electrical behavior of the ceramic material is divided into two stability classes:

The red, purple, green and light green lines correspond to the current density in the conductor cross-section for 10kHz, 100kHz, 1MHz and 10MHz respectively. The current density at the center of the conductor decreases from 98% of the surface for 10kHz, to 41% for 100kHz, to 0.4% for 1MHz, and to just 0.00000006 % for 10 MHz.

As a result of the skin effect, the effective cross-section of the conductor decreases and alternating current resistance increases. Therefore, in coil windings, conductor loss increases and the "Q" (quality factor), or quality, decreases with the increase in frequency.

As a countermeasure, from early times "Litz wire" was used for bar antenna coils in shortwave radios and mid-frequency transformer windings. Litz wire is many thin insulated conductor strands twisted together.

Many people considered applying the concept behind Litz wire to "audio cables", and came up with the idea as shown in Fig. 2.

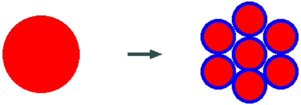

Fig. 2 Splitting a single wire conductor into many insulated conductors (conductors in red, insulators in blue)

Here, we split a single wire conductor into many insulated conductors with the same conductor cross-section area. For example, if we split into seven wires, we get the configuration shown in Fig. 2. People who think of these products think that

Skin effect and Litz wire
Kouichi Hirabayashi
http://www.mogami.com/e/puzzle/pzl-21.html

Current flowing through a cylindrical conductor distributes evenly along the conductor cross-section when the current is a direct current. However, with the increase in frequency, the effect where the current concentrates at the conductor surface and the current density decreases at the center, is known well as the skin effect (footnote 1).

For example, Fig. 1 shows the current density in the conductor cross-section of an annealed copper wire with a diameter of 1mm for frequencies 10kHz, 100kHz, 1MHz and 10MHz.

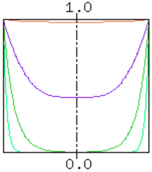

Fig. 1 Skin effect for annealed copper wire with a diameter of 1mm

The x-axis in Fig. 1 is the distance from the conductor center (center of the figure represents the center of the conductor and both ends show the conductor surface), and the y-axis is (internal current density/current density at conductor surface).

* Class 1 ceramic capacitors with high stability and low losses compensating the influence of temperature in resonant circuit application. Common EIA/IEC code abbreviations are C0G/NP0, P2G/N150, R2G/N220, U2J/N750 etc.

* Class 2 ceramic capacitors with high volumetric efficiency for buffer, by-pass and coupling applications Common EIA/IEC code abbreviations are: X7R/2XI, Z5U/E26, Y5V/2F4, X7S/2C1, etc.

The great plasticity of ceramic raw material works well for many special applications and enables an enormous diversity of styles, shapes and great dimensional spread of ceramic capacitors. The smallest discrete capacitor, for instance, is a "01005" chip capacitor with the dimension of only 0.4 mm × 0.2 mm.

The construction of ceramic multilayer capacitors with mostly alternating layers results in single capacitors connected in parallel. This configuration increases capacitance and decreases all losses and parasitic inductances. Ceramic capacitors are well-suited for high frequencies and high current pulse loads.

Because the thickness of the ceramic the dielectric layer can be easily controlled and produced by the desired application voltage, ceramic capacitors are available with rated voltages up to the 10 kV range.

Some ceramic capacitors of special shapes and styles are used as capacitors for special applications, including RFI/EMI suppression capacitors for connection to supply mains, also known as safety capacitors,[9][10] X2Y® capacitors for bypassing and decoupling applications,[11] feed-through capacitors for noise suppression by low-pass filters[12] and ceramic power capacitors for transmitters and HF applications.[13][14]

Film capacitors

Film capacitors or plastic film capacitors are non-polarized capacitors with an insulating plastic film as the dielectric. The dielectric films are drawn to a thin layer, provided with metallic electrodes and wound into a cylindrical winding. The electrodes of film capacitors may be metallized aluminum or zinc, applied on one or both sides of the plastic film, resulting in metallized film capacitors or a separate metallic foil overlying the film, called film/foil capacitors.

Metallized film capacitors offer self-healing properties. Dielectric breakdowns or shorts between the electrodes do not destroy the component. The metallized construction makes it possible to produce wound capacitors with larger capacitance values (up to 100 µF and larger) in smaller cases than within film/foil construction.

Film/foil capacitors or metal foil capacitors are use two plastic films as the dielectric. Each film is covered with a thin metal foil, mostly aluminium, to form the electrodes. The advantage of this construction is the ease of connecting the metal foil electrodes, along with an excellent current pulse strength.

A key advantage of every film capacitor's internal construction is direct contact to the electrodes on both ends of the winding. This contact keeps all current paths very short. The design behaves like a large number of individual capacitors connected in parallel, thus reducing the internal ohmic losses (ESR) and

simplicity facilitates finding those optimal diameters. Other models (such as [14] and the similar analysis in [17]) also model large strand diameters and circular bundle configurations accurately, but they fully calculate only internal, and not external, proximity effect, and so are not useful for the present purposes.

CONCLUSION

The number of strands for a minimum-loss litz-wire winding may be found by evaluating the tradeoff between proximity-effect losses and dc resistance. Of the factors leading to increased dc resistance in a litz-wire winding, only the space allocated to strand insulation varies significantly with the number of strands in a well designed construction. A power law can be used to model insulation thickness in the region of interest. Combining this with standard models for eddy-current loss results in an analytic solution for the optimal number of strands. The simplest model for loss, using only the first terms of a series expansion can be used since good designs use strands that are small compared to a skin depth. Experimental results correlate well with the simple model. Stranding for minimum loss may lead to many strands of fine wire and thus excessive expense. Minimum loss designs constrained by minimum strand size or maximum number of strands have also been derived.

where Fr is a factor relating dc resistance to an ac resistance which accounts for all winding losses, given a sinusoidal current with rms amplitude Iac. As shown in Appendix A, we can approximate Fr by

$$F_r = 1 + \frac{\pi^2 \omega^2 \mu_0^2 N^2 n^2 d_c^6 k}{768 \rho_c^2 b_c^2}$$

(2)

where ; is the radian frequency of a sinusoidal current, n is the number of strands, N is the number of turns, dc is the diameter of the copper in each strand, is the resistivity of the copper conductor, bc is the breadth of the window area of the core, and k is a factor accounting for field distribution in multi-winding transformers, normally equal to one (see Appendix A). For waveforms with a dc component, and for some non-sinusoidal waveforms, it is possible to derive a single equivalent frequency that may be used in this analysis (Appendix C). In an inductor, the field in the winding area depends on the gapping configuration, and this analysis is not directly applicable.

The analysis described here considers the strands of all litz bundles to be uniformly distributed in the window, as they would be in a single winding using N turns of wire the diameter of the litz strands. In fact, the strands are arranged in more or less circular bundles. In this sense, the analysis of [15] may be more accurate, but this difference has very little effect on the results. The most important difference between the model used here and the model in [15] is the greater accuracy of [15] for strands that are large compared to a skin depth. The simpler model is used because it is accurate for the small strand diameters that are found to be optimal, and because its

parasitic inductance (ESL). The inherent geometry of film capacitor structure results in low ohmic losses and a low parasitic inductance, which makes them suitable for applications with high surge currents (snubbers) and for AC power applications, or for applications at higher frequencies.

The plastic films used as the dielectric for film capacitors are Polypropylene (PP), Polyester (PET), Polyphenylene sulfide (PPS), Polyethylene naphthalate (PEN), and Polytetrafluoroethylene or Teflon (PTFE). Polypropylene film material with a market share of something about 50% and Polyester film with something about 40% are the most used film materials. The rest of something about 10% will be used by all other materials including PPS and paper with roughly 3%, each.[15][16]

Table Characteristics of plastic film materials for film capacitors

		Film material, abbreviated codes			
Film characteristics		PET	PEN	PPS	PP
Relative permittivity at		3.3	3	3	2.2
Minimum film thickness		0.7–0.9	0.9–1.4	1.2	2.4–3.0
Moisture absorption (%)		low	0.4	0.05	<0.1
Dielectric strength (V/µm)		580	500	470	650
Commercial realized voltage proof (V/µm)		280	300	220	400
DC voltage range (V)		50–1,000	16–250	16–100	40–2,000
Capacitance range		100 pF–22 µF	100 pF–1 µF	100 pF–0.47 µF	100 pF–10 µF
Application temperature range (°C)		−55 to +125 /+150	−55 to +150	−55 to +150	−55 to +105
ΔC/C versus temperature		±5	±5	±1.5	±2.5
Dissipation factor (•10^{-4})					
	at 1 kHz	50–200	42–80	2–15	0.5–5
	at 10 kHz	110–150	54–150	2.5–25	2–8
	at 100 kHz	170–300	120–300	12–60	2–25
	at 1 MHz	200–350	–	18–70	4–40
Time constant	at 25 °C	≥10,000	≥10,000	≥10,000	≥100,000
	at 85 °C	1,000	1,000	1,000	10,000
Dielectric absorption (%)		0.2–0.5	1–1.2	0.05–0.1	0.01–0.1
Specific capacitance		400	250	140	50

Some film capacitors of special shapes and styles are used as capacitors for special applications, including RFI/EMI suppression capacitors for connection to the supply mains, also known as safety capacitors,[17] Snubber capacitors for very high surge currents,[18] Motor run capacitors, AC capacitors for motor-run applications[19]

(the effect of current in other bundles) [14], [15]. However, the distinction is useful only as a form of bookkeeping. The actual losses in one strand of a litz bundle are simply a result of the total external field, due to the currents in all the other strands present. Another approach to calculating the loss in a litz winding is to look at it as a single winding, made up of nN turns of the strand wire, each with current i/n flowing in it, where n is the number of strands, N is the number of turns of litz wire, and i is current flowing in the overall litz bundle. The loss in the litz winding will be the same as in the equivalent single-strand winding as long as the currents flowing in all the strands are equal [16]. Other methods of calculating loss in litz wire also assume equal current in all strands [14], [12], [17]].

The objective of twisting or weaving litz wire, as opposed to just grouping fine conductors together, is to ensure that the strand currents are equal. Simple twisted bunched-conductor wire can accomplish this adequately in situations where proximity effect would be the only significant problem with solid wire. Where skin effect would also be a problem, more complex litz constructions can be used to ensure equal strand currents. Thus, in a well-designed construction, strand currents are very close to equal. However, our results remain valid even when simple twisting results in significant skin effect at the litz-bundle level. This is because the the bundle-level skin-effect loss is independent of the number of strands, and is orthogonal [15] to the strand-level eddy-current losses.

We represent winding losses by

$$P_{loss} = F_r I_{ac}^2 R_{dc}$$

(1)

All of these methods give similar results for strands that are small compared to the skin depth [12]. (See Appendix B for a discussion of one minor discrepancy.) The solutions for optimal stranding result in strand diameters much smaller than a skin depth. In this region the distinctions between the various methods evaporate, and the simplest analysis is adequate. More rigorous analysis (e.g. [12]) is important when strands are not small compared to a skin depth. In this case, losses are reduced relative to what is predicted by the analysis used here, due to the self-shielding effect of the conductor.

Previous work, such as [2] , [3], [4] has addressed optimal wire diameter for single-strand windings. The approach in [2], [3], [4] is also applicable for litz-wire windings in the case that the number of strands is fixed, and the strand diameter for lowest loss is desired. As discussed in Section V, this can be useful for cost-sensitive applications, if the number of strands is the determining factor in cost, and the maximum cost is constrained. However, this will, in general, lead to higher-loss designs than are possible using the optimal number of strands.

SKIN EFFECT, PROXIMITY EFFECT, AND LITZ WIRE

Skin effect is the tendency for high-frequency currents to flow on the surface of a conductor. Proximity effect is the tendency for current to flow in other undesirable patterns---loops or concentrated distributions---due to the presence of magnetic fields generated by nearby conductors. In transformers and inductors, proximity-effect losses typically dominate over skin-effect losses. In litz-wire windings, proximity effect may be further divided into internal proximity effect (the effect of other currents within the bundle) and external proximity effect

High pulse current load is the most important feature of film capacitors so many of the available styles do have special terminations for high currents

Electrolytic capacitors

Electrolytic capacitors have a metallic anode covered with an oxidized layer used as dielectric. The second electrode is a non-solid (wet) or solid electrolyte. Electrolytic capacitors are polarized. Three families are available, categorized according to their dielectric.

* Aluminum electrolytic capacitors with aluminum oxide as dielectric
* Tantalum electrolytic capacitors with tantalum pentoxide as dielectric
* Niobium electrolytic capacitors with niobium pentoxide as dielectric.

The anode is highly roughened to increase the surface area. This and the relatively high permittivity of the oxide layer gives these capacitors very high capacitance per unit volume compared with film- or ceramic capacitors.

The permittivity of tantalum pentoxide is approximately three times higher than aluminium dioxide, producing significantly smaller components. However, permittivity determines only the dimensions. Electrical parameters, especially conductivity, are established by the electrolyte's material and composition. Three general types of electrolytes are used:

* non solid (wet, liquid)—conductivity approximately 10 mS/cm and are the lowest cost

* solid manganese oxide—conductivity approximately 100 mS/cm offer high quality and stability
* solid conductive polymer (Polypyrrole)—conductivity approximately 10,000 mS/cm,[21] offer ESR values as low as <10 mΩ

Internal losses of electrolytic capacitors, prevailing used for decoupling and buffering applications, are determined by the kind of electrolyte.

Table Some important values of the different electrolytic capacitors

loss. This paper presents a method of finding that optimum, using standard methods of estimating the eddy-current losses.
Optimizations on magnetics design may be done to minimize volume, loss, cost, weight, temperature rise, or some combination of these factors. For example, in the design of magnetic components for a solar-powered race vehicle [1] (the original impetus for this work) an optimal compromise between loss and weight is important. Although we will explicitly minimize only winding loss, the results are compatible with and useful for any minimization of total loss (including core loss), temperature rise, volume or weight. This is because the only design change considered is a change in the degree of stranding, preserving the overall diameter per turn and overall window area usage. This does not affect core loss or volume, and has only a negligible effect on weight. However, the degree of stranding does significantly affect cost. Although we have not attempted to quantify or optimize this, additional results presented in Section V are useful for cost-constrained designs. The analysis of eddy-current losses used here does not differ substantially from previous work [2], [3], [4], [5], [6], [7], [8], [9], [10], [11], [12], [13] ([10] gives a useful review). Although different descriptions can be used, most calculations are fundamentally equivalent to one of three analyses. The most rigorous approach uses an exact calculation of losses in a cylindrical conductor with a known current, subjected to a uniform external field, combined with an expression for the field as a function of one-dimensional position in the winding area [12]. Perhaps the most commonly cited analysis [11] uses "equivalent" rectangular conductors to approximate round wires, and then proceeds with an exact one-dimensional solution. Finally, one may use only the first terms of a series expansion of these solutions, e.g. [9].

Skin Effect, Proximity Effect, and Litz Wire

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http://www.dartmouth.edu/~sullivan/litzwire/litz.html

INTRODUCTION

A salient difficulty in designing high-frequency inductors and transformers is eddy-current effects in windings. These effects include skin-effect losses and proximity-effect losses. Both effects can be controlled by the use of conductors made up of multiple, individually insulated strands, twisted or woven together. Sometimes the term litz wire is reserved for conductors constructed according to a carefully prescribed pattern, and strands simply twisted together are called bunched wire. We will use the term litz wire for any insulated grouped strands, but will discuss the effect of different constructions.

This paper addresses the choice of the degree of stranding in litz wire for a transformer winding. The number of turns and the winding cross-sectional area are assumed to be fixed. The maximum cross-sectional area of each turn is thus fixed, and as the number of strands is increased, the cross-sectional area of each strand must be decreased. This typically leads to a reduction in eddy-current losses. However, as the number of strands increases, the fraction of the window area that is filled with copper decreases and the fraction filled with insulation increases.This results in an increase in dc resistance. Eventually, the eddy-current losses are made small enough that the increasing dc resistance offsets any further improvements in eddy-current loss, and the losses start to increase. Thus, there is an optimal number of strands that results in minimum

Anode material	Electrolyte	Capacitance range (µF)	Max. rated voltage at 85 °C (V)	Upper categorie temperature (°C)	Specific ripple current (mA/mm³) [1)]
Aluminum (roughned foil)	non solid, e.g. Ethylene glycol, DMF, DMA, GBL	0.1–2,700,000	600	150	0.05–2.0
	solid, Manganese dioxide (MnO_2	0.1–1,500	40	175	0.5–2.5
	solid conductive polymere (e.g. Polypyrrole)	10–1,500	25	125	10–30
Tantalum (roughned foil)	non solid Sulfuric acid	0.1–1,000	630	125	–
Tantalum (sintered)	non solid sulfuric acid	0.1–15,000	150	200	–
	solid Manganese dioxide (MnO_2	0.1–3,300	125	150	1.5–15
	solid conductive polymere (e.g. Polypyrrole)	10–1,500	35	125	10–30
Niobium (sintered)	solid Manganese dioxide (MnO_2	1–1,500	10	125	5–20
	solid conductive polymere (e.g. Polypyrrole)	2.2–1,000	25	105	10–30

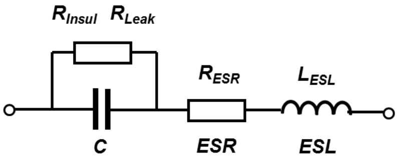
[1)] Ripple current at 100 kHz and 85 °C / volumen (nominal dimensions)

The large capacitance per unit volume of electrolytic capacitors make them valuable in relatively high-current and low-frequency electrical circuits, e.g. in power supply filters for decoupling unwanted AC components from DC power connections or as coupling capacitors in audio amplifiers, for passing or bypassing low-frequency signals and storing large amounts of energy. The relatively high capacitance value of an electrolytic capacitor combined with the very low ESR of the polymer electrolyte of polymer capacitors, especially in SMD styles, makes them a competitor to MLC chip capacitors in personal computer power supplies.

Bipolar electrolytics (also called Non-Polarized capacitors) contain two anodized aluminium foils, behaving like two capacitors connected in series opposition.

Electolytic capacitors for special applications include motor start capacitors,[22] flashlight capacitors[23] and audio frequency capacitors.[24]

Electrical characteristics



Series-equivalent circuit

Do = Diameter of the finished cable over the strands in inches = 0.056
K = Constant depending on the number of strands = 2

Rac/Rdc = 1.0003 +2(660*0.0016 / 0.056)^2 * (0.0016*1000 / 10.44)^4 = 1.393

Therefore the AC Resistance of 660/46 litz wire is: 1.39 * 7.27 = 10.13 ohms/1000ft.

A 5inch diameter basket weave coil made from 660/46 Litz wire having an inductance of 230 uH will typically require some 40 turns or about 53ft of wire. At 10.13 Ohm/kft, that comes to an AC wire resistance of 0.53 ohms for the coil alone. Were all the losses represented by just the wire, the coil would have, at 1 MHz an unloaded Q = 2700!

Naturally, wire resistance is not the only source of loss in the tank. There is a capacitor, metallic objects intruding into the magnetic field of the coil, diaelectric losses, eddy current and other losses. Its no wonder that the best coils just top out above Q = 1000 or so.

kjs

Litz Coil wire Resistance Calculation
Kevin Smith

From different Litz wire manufacturing sites one can find tables and data allowing the calculation of the resistance of your favorite litz wire.

The formula for the D.C. resistance of any Litz construction is:

Rdc = Rs (1.0515)^Nb * (1.025)^Nc / Ns

Where:
Rdc = Resistance in Ohms/1000 ft.
Rs = Maximum D.C. resistance of the individual strands (4544 for 46awg wire)
Nb = Number of Bunching operations (assume = 2)
Nc = Number of Cabling operations (assume = 1)
Ns = Number of individual strands (assume = 660)

Rdc = 4544 (1.015)^2 (1.025)^1 / 660 = 7.27 ohms / 1000 ft.

The ratio of AC resistance to DC resistance of any Litz construction is:

Rac/Rdc = S + K (N Di / Do)^2 * G

Where:
S = Resistance ratio of individual strands when isolated (1.0003 for 46awg wire)
G = Eddy Current basis factor = (Di * sqrt(f) / 10.44)^4
F = Operating Frequency in HZ (assume 1MHz)
N = Number of strands in the cable = 660
Di = Diameter of the individual strands over the copper in inches = 0.0016

Discrete capacitors deviate from the ideal capacitor. An ideal capacitor only stores and releases electrical energy, with no dissipation. Capacitor components have losses and parasitic inductive parts. These imperfections in material and construction can have positive implications such as linear frequency and temperature behavior in class 1 ceramic capacitors. Conversely, negative implications include the non-linear, voltage-dependent capacitance in class 2 ceramic capacitors or the insufficient dielectric insulation of capacitors leading to leakage currents.

All properties can be defined and specified by a series equivalent circuit composed out of an idealized capacitance and additional electrical components which model all losses and inductive parameters of a capacitor. In this series-equivalent circuit the electrical characteristics are defined by:

* C, the capacitance of the capacitor
* Rinsul, the insulation resistance of the dielectric, not to be confused with the insulation of the housing
* Rleak, the resistance representing the leakage current of the capacitor
* RESR, the equivalent series resistance which summarizes all ohmic losses of the capacitor, usually abbreviated as "ESR"
* LESL, the equivalent series inductance which is the effective self-inductance of the capacitor, usually abbreviated as "ESL".

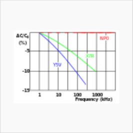
Using a series equivalent circuit instead of a parallel equivalent circuit is specified by IEC/EN 60384-1.

Frequency dependence

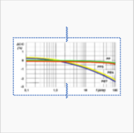
Most discrete capacitor types have more or less capacitance changes with increasing frequencies. The dielectric strength of class 2 ceramic and plastic film diminishes with rising frequency. Therefore their capacitance value decreases with increasing frequency. This phenomenon for ceramic class 2 and plastic film dielectrics is related to dielectric relaxation in which the time constant of the electrical dipoles is the reason for the frequency dependence of permittivity. The graphs below show typical frequency behavior of the capacitance for ceramic and film capacitors.

For electrolytic capacitors with non-solid electrolyte, mechanical motion of the ions occurs. Their movability is limited so that at higher frequencies not all areas of the roughened anode structure are covered with charge-carrying ions. As higher the anode structure is roughned as more the capacitance value decreases with increasing frequency. Low voltage types with highly-roughened anodes display capacitance at 100 kHz approximately 10 to 20% of the value measured at 100 Hz.

Frequency dependence of capacitance for ceramic and film capacitors



Frequency dependence of capacitance for ceramic class 2 capacitors (NP0 class 1 for comparisation)



Frequency dependence of capacitance for film capacitors with different film materials

Wire-Gauge Ampacity
http://wiki.xtronics.com/index.php/Wire-Gauge_Ampacity

Material	Melt pt °C	Resistivity (Ω-m) at 20 °C
Silver	961.78	1.59×10-8
Copper	1084.62	1.72×10-8
Aluminum	660.32	2.82×10-8
Iron	1538	1.0×10-7
Tin	231.93	1.09×10-7
Platinum	1768.3	1.1×10-7
Lead	327.46	2.2×10-7
Tin/lead solder	183	1.44x10-7

Resistivities at room temp:
http://www.epanorama.net/documents/wiring/wire_resistance.html

Element	Electrical resistivity (microohm-cm)
Aluminum	2.655
Copper	1.678
Gold	2.24
Silver	1.586
Platinum	10.5

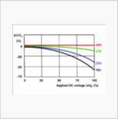
This clearly puts silver as the number one conductor and gold has higher resistance than silver or copper. It's desireable in connectors because it does not combine well with other materials so remains relatively pure at the surface. It also has the capability to adhere to itself (touch pure gold to pure gold and it sticks together) which makes for very reliable connections.

Voltage dependence

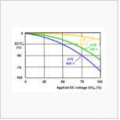
Capacitance may also change with applied voltage. This effect is more prevalent in class 2 ceramic capacitors. The permittivity of ferroelectric class 2 material depends on the applied voltage. Higher applied voltage lowers permittivity. The change of capacitance can drop to 80% of the value measured with the standardized measuring voltage of 0.5 or 1.0 V. This behavior is a small source of non-linearity in low-distortion filters and other analog applications. In audio applications this can be the reason for harmonic distortion.

Film capacitors and electrolytic capacitors have no significant voltage dependence.

Voltage dependence of capacitance for some different class 2 ceramic capacitors



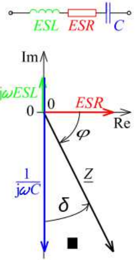
Simplified diagram of the change in capacitance as a function of the applied voltage for 25-V capacitors in different kind of ceramic grades



Simplified diagram of the change in capacitance as a function of applied voltage for X7R ceramics with different rated voltages

Impedance

Simplified series-equivalent circuit of a capacitor for higher frequencies (above); vector diagram with electrical reactances XESL and XC and resistance ESR and for illustration the impedance Z and dissipation factor tan δ



In general, a capacitor is seen as a storage component for electric energy. But this is only one capacitor function. A capacitor can also act as an AC resistor. In many cases the capacitor is used as a decoupling capacitor to filter or bypass undesired biased AC frequencies to the ground. Other applications use capacitors for capacitive coupling of AC signals; the dielectric is used only for blocking DC. For such applications the AC resistance is as important as the capacitance value.

The frequency dependent AC resistance is called impedance \scriptstyle Z and is the complex ratio of the voltage to the current in an AC circuit. Impedance extends the concept of resistance to AC circuits and possesses both magnitude and phase at a particular frequency. This is unlike resistance, which has only magnitude.

$$Z = |Z|e^{j\theta}$$

The magnitude \scriptstyle |Z| represents the ratio of the voltage difference amplitude to the current amplitude,

16	2,426.30	4.27	0.058	19/29
16	2,600.00	4.00	0.059	26/30
16	2579.85	4.02	0.059	65/34
16	2625	3.99	0.059	105/36
14	4480	2.31	0.073	7/22
14	3830.4	2.7	0.073	19/27
14	4100	2.53	0.073	41/30
14	4167.5	2.49	0.073	105/34
12	7168	1.45	0.096	7/20
12	6087.6	1.7	0.093	19/25
12	6500	1.75	0.095	65/30
12	6548.9	1.58	0.095	165/30

Stranded Wire Data Chart
http://www.wetechnologies.com/technical-info/Stranded%20Wire%20Chart.pdf

AWG	Area mil	Ohms kft	~OD	strands
36	28.00	371.00	0.006	7/44
34	43.75	237.00	0.007	7/42
32	67.27	164.00	0.008	7/40
32	76.00	136.40	0.009	19/44
30	112.00	103.20	0.012	7/38
30	118.75	87.30	0.012	19/42
28	141.75	64.90	0.015	7/36
28	182.59	56.70	0.016	19/40
27	219.52	51.47	0.018	7/35
26	277.83	37.30	0.019	7/34
26	250.00	41.48	0.021	10/36
26	304.00	34.43	0.020	19/38
24	448.00	23.30	0.024	7/32
24	396.90	26.09	0.023	10/34
24	475.00	21.08	0.024	19/36
24	384.40	25.59	0.023	41/40
22	700.00	14.74	0.030	7/30
22	754.11	13.73	0.031	19/34
22	650.00	15.94	0.030	26/36
20	1,000.00	10.32	0.035	10/30
20	1,216.00	8.63	0.037	19/32
20	1,031.94	10.05	0.036	26/34
20	1,025.00	10.02	0.036	41/36
18	1,769.60	5.86	0.048	7/26
18	1,600.00	6.48	0.047	16/30
18	1,900.00	5.46	0.049	19/30
18	1,627.29	6.37	0.047	41/34
18	1,625.00	6.39	0.047	65/36
16	2,828.00	3.67	0.060	7/24

\scriptstyle j is the imaginary unit, while the argument \scriptstyle \theta gives the phase difference between voltage and current.

In capacitor data sheets, only the impedance magnitude |Z| is specified, and simply written as "Z" so that the formula for the impedance can be written in Cartesian form

$$Z = R + jX$$

where the real part of impedance is the resistance \scriptstyle R (for capacitors \scriptstyle ESR) and the imaginary part is the reactance \scriptstyle X.

As shown in a capacitor's series-equivalent circuit, the real component includes an ideal capacitor C, an inductance L (ESL) and a resistor R (ESR). The total reactance at the angular frequency ω therefore is given by the geometric (complex) addition of a capacitive reactance (Capacitance) $X_C = -\frac{1}{\omega C}$ and an inductive reactance $X_L = \omega L_{\mathrm{ESL}}$.

To calculate the impedance \scriptstyle Z the resistance has to be added geometrically and then Z is given by

$$Z = \sqrt{ESR^2 + (X_\mathrm{C} + (-X_\mathrm{L}))^2}.$$

The impedance is a measure of the capacitor's ability to pass alternating currents. In this sense the impedance can be used like Ohms law

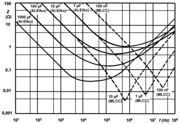
$$Z = \frac{\hat{u}}{\hat{\imath}} = \frac{U_\mathrm{eff}}{I_\mathrm{eff}}.$$

to calculate either the peak or the effective value of the current or the voltage.

In the special case of resonance, in which the both reactive resistances

$$X_C = -\frac{1}{\omega C} \text{ and } X_L = \omega L_{\mathrm{ESL}}$$

have the same value $(X_C = X_L)$ then the impedance will only be determined by {ESR}.



Typical impedance curves for different capacitance values over frequency showing the typical form with a decreasing impedance values below resonance and increasing values above resonance. As higher the capacitance as lower the resonance.

The impedance specified in the datasheets often show typical curves for the different capacitance values. With increasing frequency as the impedance decreases down to a minimum. The lower the impedance, the more easily alternating currents can be passed through the capacitor. At the apex, the point of resonance, where XC has the same value than XL, the capacitor has the lowest impedance value. Here only the ESR determines the impedance. With frequencies above the resonance the impedance increases again due to the ESL of the capacitor. The capacitor becomes to an inductance.

As shown in the graph the higher capacitance values can fit the lower frequencies better while the lower capacitance values can fit better the higher frequencies.

28	159.79	64.9	0.01264	1.4
29	126.72	81.83	0.01126	1.2
30	100.5	103.2	0.01002	0.86
31	79.7	130.1	0.00893	0.7
32	63.21	164.1	0.00795	0.53
33	50.13	206.9	0.00708	0.43
34	39.75	260.9	0.0063	0.33
35	31.52	329	0.00561	0.27
36	25	414.8	0.005	0.21
37	19.83	523.1	0.00445	0.17
38	15.72	659.6	0.00396	0.13
39	12.47	831.8	0.00353	0.11
40	9.89	1049	0.00314	0.09
46		4544	0.0016	

American Wire Gages (AWG) Sizes and Resistances
http://hyperphysics.phy-astr.gsu.edu/hbase/tables/wirega.html

AWG	Area CM	Ohms kft	Dia in	Max I
0	167810	0.0618	0.40965	328
0	133080	0.078	0.3648	283
0	105530	0.0983	0.32485	245
1	83694	0.124	0.2893	211
2	66373	0.1563	0.25763	181
3	52634	0.197	0.22942	158
4	41742	0.2485	0.20431	135
5	33102	0.3133	0.18194	118
6	26250	0.3951	0.16202	101
7	20816	0.4982	0.14428	89
8	16509	0.6282	0.12849	73
9	13094	0.7921	0.11443	64
10	10381	0.9989	0.10189	55
11	8234	1.26	0.09074	47
12	6529	1.588	0.0808	41
13	5178.4	2.003	0.07196	35
14	4106.8	2.525	0.06408	32
15	3256.7	3.184	0.05707	28
16	2582.9	4.016	0.05082	22
17	2048.2	5.064	0.04526	19
18	1624.3	6.385	0.0403	16
19	1288.1	8.051	0.03589	14
20	1021.5	10.15	0.03196	11
21	810.1	12.8	0.02846	9
22	642.4	16.14	0.02535	7
23	509.45	20.36	0.02257	4.7
24	404.01	25.67	0.0201	3.5
25	320.4	32.37	0.0179	2.7
26	254.1	40.81	0.01594	2.2
27	201.5	51.47	0.0142	1.7

Aluminum electrolytic capacitors have relatively good decoupling properties in the lower frequency range up to about 1 MHz due to their large capacitance values. This is the reason for using electrolytic capacitors in standard or switched-mode power supplies behind the rectifier for smoothing application.

Ceramic and film capacitors are already out of their smaller capacitance values suitable for higher frequencies up to several 100 MHz. They also have significantly lower parasitic inductance, making them suitable for higher frequency applications, due to their construction with end-surface contacting of the electrodes. To increase the range of frequencies, often an electrolytic capacitor is connected in parallel with a ceramic or film capacitor.[36]

Many new developments are targeted at reducing parasitic inductance (ESL). This increases the resonance frequency of the capacitor and, for example, can follow the constantly increasing switching speed of digital circuits. Miniaturization, especially in the SMD multilayer ceramic chip capacitors (MLCC), increases the resonance frequency. Parasitic inductance is further lowered by placing the electrodes on the longitudinal side of the chip instead of the lateral side. The "face-down" construction associated with multi-anode technology in tantalum electrolytic capacitors further reduced ESL. Capacitor families such as the so-called MOS capacitor or silicon capacitors offer solutions when capacitors at frequencies up to the GHz range are needed.

Inductance (ESL) and self-resonant frequency

ESL in industrial capacitors is mainly caused by the leads and internal connections used to connect the capacitor plates to the outside world. Large capacitors tend to have higher ESL than small ones because the distances to the plate are longer and every mm counts as an inductance.

For any discrete capacitor, there is a frequency above DC at which it ceases to behave as a pure capacitor. This frequency, where XC is as high as XL, is called the self-resonant frequency. The self-resonant frequency is the lowest frequency at which the impedance passes through a minimum. For any AC application the self-resonant frequency is the highest frequency at which capacitors can be used as a capacitive component.

This is critically important for decoupling high-speed logic circuits from the power supply. The decoupling capacitor supplies transient current to the chip. Without decouplers, the IC demands current faster than the connection to the power supply can supply it, as parts of the circuit rapidly switch on and off. To counter this potential problem, circuits frequently use multiple bypass capacitors—small (100 nF or less) capacitors rated for high frequencies, a large electrolytic capacitor rated for lower frequencies and occasionally, an intermediate value capacitor.

Q factor

The quality factor (or Q) of a capacitor is the ratio of its reactance to its resistance at a given frequency, and is a measure of its efficiency. The higher the Q factor of the capacitor, the closer it approaches the behavior of an ideal, lossless, capacitor.

1. Calculate the D.C. resistance of the Litz construction using Formula 3.
Rdc = 681.90 (1.015)^2 (1.025)^1 / 400 = 1.80 ohms / 1000 ft.

2. Calculate the A.C. to D.C. Resistance Ratio using Formula 2.

$$\frac{R_{AC}}{R_{DC}} = 1.000 + 2\left(\frac{400 \times 0.0045}{0.1040}\right)^2 \times \left(\frac{0.0045 \quad 50000}{10.44}\right)^4 = 1.2068$$

3. The A.C. resistance is, therefore, 1.2068 x or 1.80 ohms/1000ft.

*****

Following is an example of the calculations required to evaluate a Type 2 Litz construction consisting of 650 strands of 46 AWG single-film polyurethane-coated wire operating at 1.0 MHZ.

1. Calculate the D.C. resistance of the Litz construction using Formula 3.
Rdc = 4544 (1.015)^2 (1.025)^1 / 660 = 7.27 ohms / 1000 ft.

2. Calculate the A.C. to D.C. Resistance Ratio using Formula 2.

Rac/Rdc = 1.0003 +2(660*0.0016 / 0.056)^2 * (0.0016*1000 / 10.44)^4 = 1.393

3. The A.C. resistance is, therefore, 1.39*7.27 = **10.13** ohms/1000ft.

The D.C. Resistance of a Litz conductor is related to the following parameters:

1. Individual strands AWG.
2. Number of strands in the cable.
3. Factors relating to the increased length of the individual strands per unit length of cable (take - up). For standard Litz constructions a 1.5% increase in D.C. resistance for every bunching operation and a 2.5% increase in D.C. resistance for every cabling operation are roughly correct.

The formula derived from these parameters for the D.C. resistance of any Litz construction is:

$$R_{DC} = \frac{R_S\,(1.015)^{N_B}\,(1.025)^{N_C}}{N_S}$$

Where: Rdc = Resistance in Ohms/1000 ft.
Rs = Maximum D.C. resistance of the individual strands (taken from Table 2).
Nb = Number of Bunching operations
Nc = Number of Cabling operations
Ns = Number of individual strands

Following is an example of the calculations required to evaluate a Type 2 Litz construction consisting of 400 strands of 38 AWG single-film polyurethane-coated wire operating at 500 KHZ. This construction, designed with two bunching operations and one cabling operation, would be written 5 x 5 / 16 / 40 AWG (HM Wire uses x to indicate a cabling operation and / to indicate a bunching operation.)

The Q factor of a capacitor can be found through the following formula:

$$Q = \frac{X_C}{R_C} = \frac{1}{\omega C R_C}$$

Where:

ω is frequency in radians per second,
C is the capacitance,
X_C is the capacitive reactance, and
R_C is the series resistance of the capacitor.

Ohmic losses, ESR, dissipation factor, and quality factor

The summarized losses in discrete capacitors are ohmic AC losses. DC losses are specified as "leakage current" or "insulating resistance" and are negligible for an AC specification. AC losses are non-linear, possibly depending on frequency, temperature, age or humidity. The losses result from two physical conditions:

* line losses including internal supply line resistances, the contact resistance of the electrode contact, line resistance of the electrodes, and in "wet" aluminum electrolytic capacitors and especially supercapacitors, the limited conductivity of liquid electrolytes and
* dielectric losses from dielectric polarization.

The largest share of these losses in larger capacitors is usually the frequency dependent ohmic dielectric losses. For smaller components, especially for wet electrolytic capacitors, conductivity of liquid electrolytes may exceed dielectric losses. To measure these losses, the measurement frequency must be set. Since commercially available components offer capacitance values cover 15 orders of magnitude, ranging from pF (10–12 F) to some 1000 F in supercapacitors, it is not possible to capture the entire range with only one frequency. IEC 60384-1 states that ohmic losses should be measured at the same frequency used to measure capacitance. These are:

* 100 kHz, 1 MHz (preferred) or 10 MHz for non-electrolytic capacitors with CR ≤ 1 nF:
* 1 kHz or 10 kHz for non-electrolytic capacitors with 1 nF < CR ≤ 10 μF
* 100/120 Hz for electrolytic capacitors
* 50/60 Hz or 100/120 Hz for non-electrolytic capacitors with CR > 10 μF

A capacitor's summarized resistive losses may be specified either as ESR, as a dissipation factor(DF, tan δ), or as quality factor (Q), depending on application requirements.

Capacitors with higher ripple current I_R loads, such as electrolytic capacitors, are specified with equivalent series resistance ESR. ESR can be shown as an ohmic part in the above vector diagram. ESR values are specified in datasheets per individual type.

The losses of film capacitors and some class 2 ceramic capacitors are mostly specified with the dissipation factor tan δ. These capacitors have smaller losses than electrolytic capacitors and mostly are used at higher frequencies up to some hundred MHz. However the numeric value of the dissipation factor, measured at the same frequency, is independent on the capacitance value and can be specified for a capacitor series with a range of capacitance. The dissipation

Table 2

Frequency	Recommended Wire Gauge	Nominal Diameter over Copper	D.C. Resistance Ohms/M' (Max)	Single Strand R_{AC} / R_{DC} "S"
60 HZ - 1 KHZ	28 AWG	0.0126	66.37	1.0000
1 KHZ – 10 KHZ	30 AWG	0.0100	105.82	1.0000
10 KHZ – 20 KHZ	33 AWG	0.0071	211.70	1.0000
20 KHZ – 50 KHZ	36 AWG	0.0050	431.90	1.0000
50 KHZ – 100 KHZ	38 AWG	0.0040	681.90	1.0000
100 KHZ – 200 KHZ	40 AWG	0.0031	1152.3	1.0000
200 KHZ – 350 KHZ	42 AWG	0.0025	1801.0	1.0000
350 KHZ – 850 KHZ	44 AWG	0.0020	2873.0	1.0003
850 KHZ – 1.4 MHZ	46 AWG	0.0016	4544.0	1.0003
1.4 MHZ – 2.8 MHZ	48 AWG	0.0012	7285.0	1.0003

After the individual wire gauge has been decided and assuming that the Litz construction has been designed such that each strand tends to inhabit all possible positions in the cable to just about the same extent, the ratio of A.C to D/C resistance of an remote Litz conductor can be determined from the following formula.

$$\frac{\text{Resistance to Alternating Current}}{\text{Resistance to Direct Current}} = S + K\left(\frac{N\,D_1}{D_O}\right)^2 G$$

Where:
S =Resistance ratio of individual strands when isolated (taken from Table 1 or 2)

G =Eddy Current basis factor = $\left(\frac{D_1\sqrt{F}}{10.44}\right)^4$
F =Operating Frequency in HZ
N =Number of strands in the cable
Dl =Diameter of the individual strands over the copper in inches
Do =Diameter of the finished cable over the strands in inches
K =Constant depending on N, given in the following table:

N	3	9	27	Infinity
K	1.55	1.84	1.92	2

Litz Designing Calculations
HM Wire International, Inc.
http://www.litz-wire.com/New%20PDFs/Litz_Designing_Calculations.pdf

Typically, the designing engineer that uses Litz knows the operating frequency and RMS current required for his application. Since the main advantage of a Litz conductor is the reduction of AC losses, the first thought in any Litz design is the operating frequency. The operating frequency not only influences the authentic Litz construction, it also determines the individual wire gauge.

Ratios of alternating-current resistance to direct-current resistance for an isolated solid round wire (S) in terms of a value (X) are shown in Table 1.

Table 1

X	0	0.5	0.6	0.7	0.8	0.9	1.0
S	1.0000	1.0003	1.0007	1.0012	1.0021	1.0034	1.005

The Value of X for copper wire is determined by Formula 1.

$$X = 0.271\, D_M \sqrt{F_{MHZ}}$$

Where: Dm = Wire Diameter in mils
Fmhz = Frequency in Megahertz

From Table 1 and other practical data, the following table of recommended wire gauges vs. frequency for most Litz constructions has been prepared.

Table 2

factor is determined as the tangent of the reactance (X_C - X_L) and the ESR, and can be shown as the angle δ between imaginary and the impedance axis.

If the inductance ESL is small, the dissipation factor can be approximated as:

$$\tan\delta = ESR \cdot \omega C$$

Capacitors with very low losses, such as ceramic Class 1 and Class 2 capacitors, specify resistive losses with a quality factor (Q). Ceramic Class 1 capacitors are especially suitable for LC resonant circuits with frequencies up to the GHz range, and precise high and low pass filters. For an electrically resonant system, Q represents the effect of electrical resistance and characterizes a resonator's bandwidth B relative to its center or resonant frequency f_0. Q is defined as the reciprocal value of the dissipation factor.

$$Q = \frac{1}{tan\delta} = \frac{f_0}{B}$$

A high Q value is for resonant circuits a mark of the quality of the resonance.

Comparization of ohmic losses for different capacitor types for resonant circuits (Reference frequency 1 MHz)

Capacitor type	Capacitance (pF)	ESR at 100 kHz (mΩ)	ESR at 1 MHz (mΩ)	tan δ at 1 MHz (10^{-4})	Quality factor
Silicon capacitor[37]	560	400	—	2,5	4000
Mica capacitor[38]	1000	650	65	4	2500
Class 1 ceramic capacitor (NP0)[39]	1000	1600	160	10	1000

Equivalent series resistance

In a non-electrolytic capacitor and electrolytic capacitors with solid electrolyte the metallic resistance of the leads and electrodes and losses in the dielectric cause the ESR. Typically quoted values of ESR for ceramic capacitors are between 0.01 and 0.1 ohms. ESR of non-electrolytic capacitors tends to be fairly stable over time; for most purposes real non-electrolytic capacitors can be treated as ideal components.

Aluminium and tantalum electrolytic capacitors with non solid electrolyte have much higher ESR values, up to several ohms, and ESR tends to increase with frequency due to effects of the electrolyte. A very serious problem, particularly with aluminium electrolytics, is that ESR increases over time with use; ESR can increase enough to cause circuit malfunction and even component damage,[1] although measured capacitance may remain within tolerance. While this happens with normal aging, high temperatures and large ripple current exacerbate the problem. In a circuit with significant ripple current, an increase in ESR will increase heat dissipation, thus accelerating aging.

Electrolytic capacitors rated for high-temperature operation and of higher quality than basic consumer-grade parts are less susceptible to become prematurely unusable due to ESR increase. A cheap electrolytic capacitor may be rated for a life of less than 1000 hours at 85°C (a year is about 9000 hours). Higher-grade parts are typically rated at a few thousand hours at maximum rated temperature, as can be seen from manufacturers' datasheets. Electrolytics of higher capacitance have lower ESR; if ESR is critical, specification of a part of larger capacitance than is otherwise required may be advantageous.

References:

AVX Basin Capacitor Formulas.
http://www.avx.com/docs/Catalogs/cbasic.pdf

Ben Tongue 1999, Sensitivity and selectivity issues in crystal radio sets including diode problems; measurements of the Q of variable and fixed capacitors, RF loss in slide switches and loss tangent of various dielectrics.
http://www.bentongue.com/xtalset/24Cmnts/24Cmnts.html

Wikipedia, Types of Capacitors, Ohmic Losses, Dissipation Factor and Q.
http://en.wikipedia.org/wiki/Types_of_capacitor#Ohmic_losses.2C_ESR.2C_dissipation_factor.2C_and_quality_factor

Dr. Gary L. Johnson, 2001. Lossy Capacitors
http://www.eece.ksu.edu/~gjohnson/tcchap3.pdf

Polymer capacitors usually have lower ESR than wet-electrolytic of same value, and stable under varying temperature. Therefore polymer capacitors can handle higher ripple current. From about 2007 it became common for better-quality computer motherboards to use only polymer capacitors where wet electrolytics had been used previously.[2]

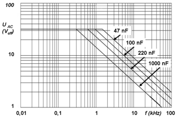
The ESR of capacitors of relatively high capacity (from about 1 µF), which are the ones likely to cause trouble, is easily measured in-circuit with an ESR meter.

Typical values of ESR for capacitors

Type	22 µF	100 µF
Standard aluminum	0.1 - 3.0 Ω	0.05 - 0.5 Ω
Ceramic	<0.015 Ω	

AC current

An AC load only can be applied to a non-polarized capacitor. Capacitors for AC applications are primarily film capacitors, metallized paper capacitors, ceramic capacitors and bipolar electrolytic capacitors.



The rated AC load for an AC capacitor is the maximum sinusoidal effective AC current (rms) which may be applied continuously to a capacitor within the specified temperature range. In the datasheets the AC load may be expressed as

* rated AC voltage at low frequencies,
* rated reactive power at intermediate frequencies,
* reduced AC voltage or rated AC current at high frequencies.

Typical rms AC voltage curves as a function of frequency, for 4 different capacitance values of a 63 V DC film capacitor series

The rated AC voltage for film capacitors is generally calculated so that an internal temperature rise of 8 to 10 °K is the allowed limit for safe operation. Because dielectric losses increase with increasing frequency, the specified AC voltage has to be derated at higher frequencies. Datasheets for film capacitors specify special curves for derating AC voltages at higher frequencies.

If film capacitors or ceramic capacitors only have a DC specification, the peak value of the AC voltage applied has to be lower than the specified DC voltage.

AC loads can occur in AC Motor run capacitors, for voltage doubling, in snubbers, lighting ballast and for power factor correction PFC for phase shifting to improve transmission network stability and efficiency, which is one of the most important applications for large power capacitors. These mostly large PP film or metallized paper capacitors are limited by the rated reactive power VAr.

Bipolar electrolytic capacitors, to which an AC voltage may be applicable, are specified with a rated ripple current.

Insulation resistance and self-discharge constant

The resistance of the dielectric is finite, leading to some level of DC "leakage current" that causes a charged capacitor to lose charge over time. For ceramic and film capacitors, this

134	1025	0.0006250	1600	0.852
48	1712	0.0019341	517	6.425

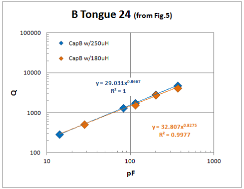
On the main plot then I have included for Ben's Cap B a second Q vs. ESR curve to illustrate the estimated benefit of using a higher tuning capacitance range. The new curve slides mostly to lower ESR (especially at the high end of the tuning range) and also to slightly lower Q. The improvements made appear modest from the modeled data and, for the case of a high-Q capacitor the possible improvements may be slight indeed. It is comforting to know that at least there should be no negative impact. Lower ESR certainly improves tank overall efficiency. This is a win-win situation where the set designer wins twice!

In seeking the "Holy Grail" tank capacitor I would add to the list of valuable features. We have ceramic stator supports, silver plating on the vanes, and good wiper contacts as features desired. I would add to this an extended tuning range if the capacitor is truly to be considered "Holy". With a standard 230 uH coil the tuning for the BCB ranges from 40 to 380 pF or so. With a 180 uH coil the capacitance needed ranges from 50 to 490 pF, (stray capacitance not accounted for).

Two questions immediately come to mind with this analysis:

1) To how low a coil inductance can you go?
2) If you find a few of these super-caps can you send me one?

kjs 6/2013 (in France)



provides the estimated Q, and calculated D, and ESR for the capacitances from Fig 5 (with 20pF stray added). The values of Q do not improve by moving to higher capacitances but the ESR of the cap is slightly (at low tuning freq) and well improved (at the high tuning freq). This is the goal. It is worthy to note that the tuning frequencies do not line up with the original data from Ben's Fig 3 as seen on my main plot. While Ben specifically noted that his figure 3 included 20pF strays, in his figure 5 he speaks in absolute capacitances. I did my best to correct for this but there remains some ambiguity in the presentation. The point remains though; moving to higher capacitances in the tank will lower the tank losses due to ESR.

250 inductor (w/20pF stray)

tank pF	KHz	D	Q	ESR ohm
385	513	0.0002083	4800	0.177
220	679	0.0003448	2900	0.404
134	870	0.0005714	1750	0.917
101	1000	0.0007638	1309	1.502
34	1726	0.0034973	286	23.046

180 inductor (w/20pF stray)

tank pF	KHz	D	Q	ESR ohm
385	605	0.0002381	4200	0.172
220	800	0.0003571	2800	0.355

resistance is called "insulation resistance Rins". This resistance is represented by the resistor Rins in parallel with the capacitor in the series-equivalent circuit of capacitors. Insulation resistance must not be confused with the outer isolation of the component with respect to the environment.

The time curve of self-discharge over insulation resistance with decreasing capacitor voltage follows the formula

$$u(t) = U_0 \cdot \mathrm{e}^{-t/\tau_s},$$

With stored DC voltage U_0 and self-discharge constant

$$\tau_s = R_{ins} \cdot C$$

Thus, after τ_s voltage U_0 drops to 37 % of the initial value.

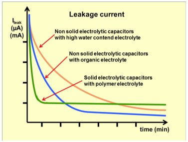
The self-discharge constant is an important parameter for the insulation of the dielectric between the electrodes of ceramic and film capacitors. For example, a capacitor can be used as the time-determining component for time relays or for storing a voltage value as in a sample and hold circuits or operational amplifiers.

Class 1 ceramic capacitors have an insulation resistance of at least 10 GΩ, while class 2 capacitors have at least 4 GΩ or a self-discharge constant of at least 100 s. Plastic film capacitors typically have an insulation resistance of 6 to 12 GΩ. This corresponds to capacitors in the uF range of a self-discarge constant of about 2000–4000 s.[42]

Insulation resistance respectively the self-discharge constant can be reduced if humidity penetrates into the winding. It is partially strongly temperature dependent and decreases with increasing temperature. Both decrease with increasing temperature.

In electrolytic capacitors, the insulation resistance is defined as leakage current.

Leakage current



The general leakage current behavior of electrolytic capacitors depend on the kind of electrolyte

For electrolytic capacitors the insulation resistance of the dielectric is termed "leakage current". This DC current is represented by the resistor Rleak in parallel with the capacitor in the series-equivalent circuit of electrolytic capacitors. This resistance between the terminals of a capacitor is also finite. Rleak is lower for electrolytics than for ceramic or film capacitors.

The leakage current includes all weak imperfections of the dielectric caused by unwanted chemical processes and mechanical damage. It is also the DC current that can pass through the dielectric after applying a voltage. It depends on the interval without voltage applied (storage time), the thermic stress from soldering, on voltage applied, on temperature of the capacitor, and on measuring time.

The leakage current drops in the first minutes after applying DC voltage. In this period the dielectric oxide layer can self-repair weaknesses by building up new layers. The time required depends generally on the electrolyte. Solid

pF		f	Q	
tank	cap	MHz	Cap A	Cap B
375	355	520	19000	4750
190	170	730	14000	3230
114	94	943	14400	1610
101	81	1000	*13148*	*1828*
60	40	1300	11600	870
35	15	1710	9800	460

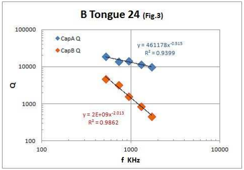
for I = 250 uH and 20pF stray in tank
(Q data in italics is computed)

The quest for high Q is indeed a quest for low ESR as long as you work in the capacitance range needed. One interesting piece of advice I have occasionally seen is that, to increase your tank Q, it is better to use lower-value coils and higher-value tuning caps. As much of the losses in the tank derive from losses in the coil this makes excellent sense. I wish to ask the question what about those higher-value caps? The plot immediately suggests that moving to higher capacitance drives down the ESR, this is a good thing. It may or may not improve the cap Q but I am not sure it matters as long as resistive losses are minimized. To better set my expectations I have again turned to Ben's Cap B data.

The small plot below takes data presented in his Figure 5. Data presented include Q at various capacitances and frequencies. I cross-plotted his data for pF and Q for the appropriate frequencies using both 250 uH and 180 uH inductances. Reading the actual Q off the tiny plot was an estimate at best, but it is interesting. This allows one to estimate his cap Q at any capacitance setting. The table below the likely capacitor data will be specific to caps in the uF range, or 10^6 higher that what we need in a crystal radio tank. Be careful not to get enamored with the high Q values occasionally quoted, check the capacitance. From the plot it should be apparent that not all Q's are created equal. Larger-value caps have lower ESR and so the Q's are often agreeably high. I have plotted a few specific cap measurements from various sources.

Two measurements should stand out; these are indicated by the larger blue diamonds. These are measurements by Bill Hebbert and presented by Ben Tongue in section B of his Article 24. I reproduce his Figure three below, with grateful apologies for the liberties I have taken,

By replotting Ben's figure I can calculate the relation between Q and frequency (table below) and so compute the estimated Q at any frequency. On my main plot I give Ben's Capacitor A and B for 1.0 MHz given the 250 uH inductor. Ben states that the plot assumes



20pF stray added. Since the capacitance needed to resonate at 1 MHz and 250 uH is 101 pF, I make the assumption his caps were set to 81 pF. This accounts for why the diamonds are set slightly off the 100 pF line of the plot. Given his Q measurements it is easy to compute the ESR for each cap. This is the world of the crystal radio builder. I include the ESR at the BCB endpoints for each cap as well to give the full picture.

electrolytes drop faster than non-solid electrolytes but remain at a slightly higher level.

The leakage current in non-solid electrolytic capacitors as well as in manganese oxide solid tantalum capacitors decreases with voltage-connected time due to self-healing effects. Although electrolytics leakage current is higher than current flow over insulation resistance in ceramic or film capacitors, the self-discharge of modern non solid electrolytic capacitors takes several weeks.

A particular problem with electrolytic capacitors is storage time. Higher leakage current can be the result of longer storage times. These behaviors are limited to electrolytes with a high percentage of water. Organic solvents such as GBL do not have high leakage with longer storage times.

Leakage current is normally measured 2 or 5 minutes after applying rated voltage.

Dielectric absorption (soakage)

Dielectric absorption occurs when a capacitor that has remained charged for a long time discharges only incompletely when briefly discharged. Although an ideal capacitor would reach zero volts after discharge, real capacitors develop a small voltage from time-delayed dipole discharging, a phenomenon that is also called dielectric relaxation, "soakage" or "battery action".

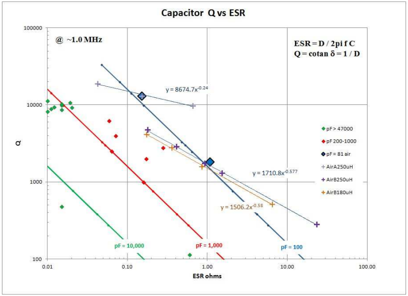
Values of dielectric absorption for some often used capacitors

Type of capacitor	Dielectric Absorption
Air and vacuum capacitors	Not measurable
Class-1 ceramic capacitors, NP0	0.6%
Class-2 ceramic capacitors, X7R	2.5%
Polypropylene film capacitors (PP)	0.05 to 0.1%
Polyester film capacitors (PET)	0.2 to 0.5%
Polyphenylene sulfide film capacitors (PPS)	0.05 to 0.1%
Polyethylene naphthalate film capacitors (PEN)	1.0 to 1.2%
Tantalum electrolytic capacitors with solid electrolyte	2 to 3%,[45] 10%[46]
Aluminium electrolytic capacitor with non solid electrolyte	10 to 15%
Double-layer capacitor or super capacitors	data not available

In many applications of capacitors dielectric absorption is not a problem but in some applications, such as long-time-constant integrators, sample-and-hold circuits, switched-capacitor analog-to-digital converters, and very low-distortion filters, it is important that the capacitor does not recover a residual charge after full discharge, and capacitors with low absorption are specified.[47] The voltage at the terminals generated by the dielectric absorption may in some cases possibly cause problems in the function of an electronic circuit or can be a safety risk to personnel. In order to prevent shocks most very large capacitors are shipped with shorting wires that need to be removed before they are used.[48]

dependant. It is important to remember that not all Q's are created equal. What we want is to keep the ESR to a minimum.

On the plot there are a number of data trends presented. Note first that both plot axes are logarithmic, it makes a difference on how one thinks of the data. Additionally, note that the plot is specific (mostly) for data at 1.0 MHz. I make the assumption that a cap that is superior at that frequency is likely to be superior across the BCB.



The first things to discuss are the three slanting lines. These are computed from a table of Dissipation Factors at 1.0MHz by Dr. Johnson. I have supplied the capacitances and calculated the resulting Q and ESR values. When one goes looking at capacitor data sheets or summary tables one is likely to find data computed at 100 KHz and 1.0 MHz (mercifully) but that

Crystal Radio Capacitors and ESR
Kevin Smith

In my ongoing effort to understand crystal radios and the theory that makes them work I have recently turned to setting my expectations concerning capacitance. The tuning capacitor is, next the coil itself, the principal component of the set. Few builders fabricate their own tuning caps so I suppose it is possible to take them a bit for granted. I have been reading on the radio forums some threads concerning radio Q and the subject of the variable cap Q seems to be getting more attention. So, what is the deal?

I set about my study by gathering what I could find on the web and at the close of this note I include a short reference list of useful pages from which I extracted data the ideas. I wish to extend special thanks to Ben Tongue who, as usual, has provided the essential measurements and background. The plot below of capacitor Q versus Equivalent Series Resistance (ESR) summarizes my findings. It is a bit busy as there are a number of interesting and important ideas represented in the relationships presented.

Any discussion of component quality needs first to acknowledge that the fight to increase the component Q is in reality a fight to eliminate (not possible) or reduce as much as feasible the resistive losses associated with that component. It helps then to look not just at Q, but at how Q relates to the component ESR. There are two fundamental equations that describe this relation: 1) Q = cotan theta = 1/D. D is the dissipation factor and is an attribute of the dielectric material used in the construction of the capacitor. Equation 2) ESR = D / 2pi f C. This equation tells us that capacitor losses expected due to resistance are both angular frequency and capacitance

ARTICLE 24

Sensitivity and selectivity issues in crystal radio sets including diode problems; measurements of the Q of variable and fixed capacitors, RF loss in slide switches and loss tangent of various dielectrics

By Ben H. Tongue
http://www.bentongue.com/xtalset/20MeaAGs/20MeaAGs.html

Part A: The Issues.

The sensitivity and selectivity in a crystal radio set can be impacted by many factors, including:

* A1: Ineffective coupling of the antenna to the tank circuit.
* A2: Resistive RF losses in the tank circuit.
* A3: Inappropriate diode.
* A4: Unintentional diode damage from exposure to electrostatic discharge (ESD) or voltages higher than the minimum reverse voltage rating. This is one cause of the loss of weak-signal sensitivity over time.
* A5: Relations between antenna-ground system, diode RF input and tank loss resistance, as effecting selectivity and loaded Q.
* A6: Audio transformer loss.

A1. There is a practical minimum limit to the possible impedance transformation ratio of the series resistance of the antenna-ground circuit to the shunt value desired across the tank circuit, when the transformation means is just a series capacitor between the antenna and the top of the tank. This problem occurs when the capacitor required for the desired

Do you really want a linear ohm meter scale? How about a LOG scale? Analog or digital?
What are the reasonable good and bad limits of ESR? What range in ohms? How sensitive does it need to be?
Input - To protect or not protect? What about charged caps?
Power - Battery or A/C adaptor? Battery life?
Physical Size
Cost

All of the above choices were carefully considered in the design of the Capacitor Wizard®. The end product was to be easy to use, durable, small & lightweight, accurate, measure ESR in circuit and inexpensive. I think I have met these design goals with the Capacitor Wizard®!

impedance transformation becomes so large that it causes the tank inductance to resonate below the desired frequency (See Article #22, Part 2). The solution here is to use a lower value inductance for the tank or to tap the antenna down on the tank (towards ground). This is why some experimenters find that low inductance tank circuits seem to work better than those of higher inductance. If one does not use a series capacitor for impedance transformation, the antenna may be just tapped down on the tank. Another alternative is to connect the antenna-ground system to a low value unturned inductance that is coupled to the tank.

The main advantage of using a series capacitor connected to the top of the tank (for impedance matching) is that it moves the undesired short wave resonance of the antenna-ground circuit (present in every single tuned crystal radio set) to the highest possible frequency and reduces its strength. A disadvantage is that unless a high enough Q variable capacitor is used, insertion power loss is increased, especially at the high frequency end of the band. See Part B, Section 1 of this Article and Article #22, Part 7.

A2. Resistive RF losses in the tank circuit are affected by: 1) Losses in any capacitor used for tuning or RF coupling. 2) Physical size of the coil and such items as length/diameter ratio, cross-section size and shape, and turns spacing (to reduce coil proximity losses). 3) Loss tangent in the coil form material, wire insulation and all dielectric material penetrated by the electric field of the coil. 4) Wire size and plating, if any. Silver plating is good but tin plating is bad, especially at the high end of the band. 5) Wire construction such as litz, solid or un-insulated stranded. The latter should be avoided. 6) Switches (if used). 7) Magnetic coupling from the coil to nearby lossy metallic objects. 9) Capacity coupling from "hot" high impedance RF points through a lossy RF return path to ground. See Article #22, point 7 of Part 10. Some comments: The loss from the loss tangent of the dielectric material used for the mounting base of detector stands can be nontrivial. Loss tangent of the material used for a front panel can cause dissipative loss if terminals provided for the connection of an external diode are too close together. See Part B, section 3 of this Article.

A3. A diode with too low an axis-crossing resistance (too high a saturation current) will resistively load the tank too heavily, causing a low loaded Q that results in loss of selectivity and sensitivity. A diode with too high an axis-crossing resistance (too low a saturation current) will increase selectivity because it only lightly resistively loads the tank circuit. The disadvantage is that sensitivity is reduced (many people liken this to the diode having too high a "turn-on" voltage.) and considerable audio distortion is generated. A little reverse or forward DC voltage bias will usually fix up these performance problems. See Article #9.

A way to check whether weak signal performance would be improved if one used a diode with a different saturation current (Is), but without experimenting with DC bias, is as follows: 1) Give the diode a one second or so spray with an aerosol "component cooler" . The reduction in temperature will temporarily substantially reduce the Is of the diode. If performance improves during the subsequent warm-back-up period, but before reaching room temperature, the diode has too high a room temperature Is. 2) Heat the diode by holding a hot soldering iron next to it for 5 seconds or so or give it a quick blow from a hot hair dryer. If performance improves during the subsequent cool-back-down period, but before reaching room temperature, the diode has too low a room

voltage across the test terminals to maintain its constant current. This voltage would turn on solid state devices when measuring ESR in circuit. For this reason the constant voltage method was chosen for the Capacitor Wizard.

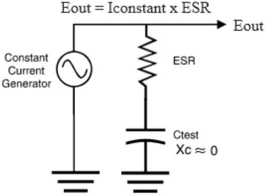
For convenience we don't want a linear scale because it would require a Range Switch. I modified the basic constant voltage diagram above to include a resistor in parallel with the constant voltage generator that has the desirable effect of logging the ohm scale. It also serves to lower the output voltage to a level far below that required to turn on solid state devices.

Conclusion:

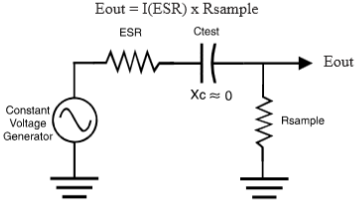
It should now be obvious that a DC ohm meter cannot measure ESR! A Capacitance meter is equally unsuitable for ESR measurements. ESR doesn't exist under the STATIC measurement condition of a DC ohm meter. Capacitance tells one nothing about a capacitors ESR. To measure ESR a controlled DYNAMIC test condition must be created. To create this test condition the engineer needs to consider a number of "Design Judgement Calls" based upon his experience and the desired end result. Electronic design is an exercise in compromise. There are many ways to design an ESR meter and each has its advantages and disadvantages. The quality of the end product is the result of wise choices made by the design engineer. Some of the important design considerations for an ESR meter are:

Which to use - constant voltage or constant current? At what magnitudes?
Test Signal - Sine wave or square wave? At what frequency?

with a linear meter scale. We could use either digital or analog meter displays.



Constant Current Method



Constant Voltage Method

Constant current has the advantage of conveniently measuring low values of ESR down in the milli ohm range. It however has the disadvantage of applying a theoretical infinitely large

temperature Is. Is, for the usual Schottky diode, changes by about two times for each 10° C. temperature change. Germanium diodes probably act the same. Aerosol component coolers are available from most Electronics Distributors, including Radio Shack.

A4. Semiconductor diodes are subject to damage from exposure to electrostatic discharge (ESD) and voltages higher than their reverse breakdown values. Easy-to-check-for damage shows up as an increased reverse current when tested with a reverse DC voltage close to or less than the specified minimum breakdown voltage, as compared to an undamaged diode. Other types of damage can also occur. The effect on detector performance can be anywhere from a mild to a very great reduction of weak-signal sensitivity. This type of damage has the effect of placing a resistor across the diode, which reduces tank Q and adds a parasitic, unneeded audio load resistance. If one has an old VOM having a d'Arsonval moving-coil meter movement (not a digital type) such as the Triplett 360, Weston 980 or Simpson 260, one can do a quick check of the back leakage of a Schottky diode by measuring its back resistance on the X1,000 range. If the needle does not move, the diode has probably not been damaged from ESD. This test does not apply to so-called zero-bias Schottky detector diodes because of their usual low reverse breakdown voltages. Undamaged germanium diodes normally have greater reverse leakage (lower back-resistance) than the usual Schottky detector diodes. This does not effect performance in crystal set applications. The germaniums seem less subject to the electrostatic damage problem precisely because their lower back-resistance (and high reverse breakdown voltages) tends to bleed off any electrostatic charge that might accumulate, maybe from handling. Diodes with the lowest leakage can be selected with a simple test as follows: Connect a 3-4.5 volt

DC source, a 4.7k to 10k resistor, the diode and a DVM set to read DC current in series. Polarize the battery so that the diode is back biased. If the current is 2 uA or less, the parasitic back leakage resistance is greater than 2 Megohms or so and all should be well, as far as weak signal loss is concerned. The resistor is used to prevent damage to the test diode if it is accidentally connected in the forward direction. This test may be used to sort out the best weak-signal sensitivity diodes and exclude damaged ones. It is probable that for a selection of diodes having the same part number, ones with the lowest reverse leakage current will deliver the best weak-signal performance.

I believe poor weak-signal results some people have reported with the Avago (formerly Agilent) 5082-2835 and HSMS-2820 diodes can be laid at the doorstep of ESD damage. I experienced my first problem with these diodes by storing a few new ones in a propylene vitamin pill bottle for a while. When I wanted a new one I picked up the bottle, shook it (I think,) to see if something was inside and withdrew a diode. It performed badly, so I checked its back resistance with my Triplett 360 VOM as suggested above and found a low back resistance reading. Checking all the others, I found they were all bad. Static electricity did them in... beware! It was pretty stupid of me to not use anti-static bags. BTW, I've never had problems with those diodes if stored in paper bags. The bags must have had enough moisture content to provide sufficient conductivity to bleed off any static charge that might accumulate.

It is difficult to properly impedance match these diodes (see Part 4 of Article #5) because of their low saturation current (high axis-crossing resistance). They work best if 3 or or so are connected in parallel to raise the effective saturation current. If

Disadvantages: Not suitable for in circuit use, will not find open caps, keying must be highly accurate and short duration or ESR readings will be wrong, moderately complex, moderate cost, requires periodic calibration

We now have a method to correct for capacitive reactance by using a high test frequency and a valid formula to compute ESR. Let's investigate our formula E=IR.

The formula E = IR has three unknown quantities and it is not usable until we get the formula into the form of two known and one unknown quantities . The quantity we are trying to find is "R" so let's make it the unknown. We are left with "E" (voltage) and "I" (current) as the two variable quantities to be measured. Using conventional test equipment we can drive a capacitor with an ac 100khz signal, measure both "E" and "I" and, using the formula E = I x R, compute "R" (ESR). That method will work but it is clumsy, slow, and requires two pieces of test equipment - an AC volt meter, and an AC ammeter. If we were to hold one of the two variables in the test signal at a known CONSTANT VALUE we could measure the other variable with only one piece of test equipment! We could then calibrate its scale to represent the COMPUTED VALUE of "R"!! We could hold the test signals current(I) constant and measure the voltage(E) and then use the result to compute the value of "R". Or we could hold the test signals voltage(E) constant (the Capacitor Wizards method) and measure the test signals current(I). The result could be displayed on an analog or digital meter calibrated in ohms. These two methods of measuring resistance are well known by electronic measurement engineers as the "CONSTANT CURRENT" and "CONSTANT VOLTAGE" methods. Since the formula E = IR is a linear expression, both methods will yield an ohm meter

(1) Use a high test frequency: This is the technique the Capacitor Wizard® uses to "Zero Out" capacitive reactance.

Advantages: Will find open capacitors within the specified range of capacitance, Simple circuit, Reasonable Cost, no calibration required

Disadvantages : Only valid above a certain capacitance value, dependent upon test frequency

Xc = capacitive reactance

Zc = Total AC resistance (Vector Sum of Xc and ESR)

ESR = Equivalent Series Resistance

C = capacitance in Farads

F = Frequency in Hertz

Proof:

$$\text{If } Xc = 1/(6.28 * F * C) = 0 \quad \text{AND} \quad Zc = \sqrt{ESR^2 + Xc^2}$$

$$\text{THEN } Zc = \sqrt{ESR^2 + 0^2} = \sqrt{ESR^2} = ESR$$

(2) Electronically subtract capacitive reactance from measured AC ohms:

Uses: Suitable for capacitor manufacturer

Advantages: Measure ESR of any capacitor, can also be adapted to measure capacitance

Disadvantages: Requires operator skills, not suitable for in circuit use, circuit complex, cost more, easily damaged, will not find open capacitors, requires periodic calibration

(3) Sample/Hold Keying:

Uses: Inexpensive method to measure ESR of any capacitor

Advantages: Measure ESR of any capacitor

one wishes to use only one of these diodes, it could have a low forward bias applied to get the same result (see Article #9). Keep in mind that these diodes do not do very well on strong signals when optimally impedance matched for weak signal reception because of their low reverse voltage breakdown rating. Avago specifies that rating as 15 volts for these diodes. If a rectified voltage approaching 7.5 volts appears across the DC load (now operating in 'peak-detector' mode), a maximum reverse voltage of about 15 volts will appear across the diode for every cycle of the RF carrier. When the diode is operated at this signal level or greater, a 'bucking' rectified current flows through the diode reducing audio output and causing audio distortion. This problem is eliminated by using diodes having a greater reverse breakdown voltage such as most ITT FO 215 and most 1N34A units.

A5. A question asked by many designers is this: What is the best approach when deciding how to 'impedance match' the transformed antenna-ground-system resistance to the RF input resistance of the diode detector? This would be a no-brainer if the resonant resistance of the tank was infinite, but it is not.

Table 1 - Definitions of Terms

IPL	Insertion power loss
Lo	Inductance of the tank
Ql	Loaded Q of the tank
Qo	Unloaded Q of the tank
Rag	Actual antenna-ground-system resistance
Ragt	A antenna-ground-system resistance transformed its value at the top of the tank
Rd	Actual input RF resistance of the diode.
Rdt	Input RF resistance of the diode as transformed to its value at the top of the tank.
Ro	Resistor representing all losses in the tank
Xa	Antenna-ground-system impedance transformation.
Xd	Diode input RF resistance transformation.

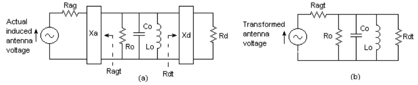


Fig. 1 - Simplified schematics of a single-tuned crystal set.

Simplified Schematic

The presence of finite tank loss increases IPL from the theoretical zero of the matched impedance case. It also reduces selectivity. Fig. 1a shows a simplified schematic of a single-tuned crystal radio set. The input impedance transformation Xa might take the form of capacitor in series with the 'Rag' and the 'induced voltage source' series combination or a tap on the tank. The diode input RF resistance transformation might take the form of a tap on the tank or a capacitor in series with the diode, along with other components. See Son of Hobbydyne and Hobbydyne II at http://www.hobbytech.com/crystalradio/crystalradio.htm. Table 2 shows calculated data for the simplified schematic shown in Fig.1b.

Table 2 - Calculated Insertion power loss (IPL), ratio of loaded to unloaded Q and impedance match (return loss*) in a single-tuned circuit.

does not really exist as a physical entity so direct measurements across the ESR resistor are no t possible! However, if a method of correcting for the effects of capacitive reactance is provided, and considering all ESR resistances are INPHASE, ESR can be calculated and measured by using the basic electronics formula E = I x R! This is the Basic Electronics foundation used to design the Capacitor Wizard®!

Before the formula E = I x R can be used to compute ESR, we must find a way to "ZERO OUT" or otherwise negate the effects of capacitive reactance. Here's why: Consider a series AC circuit that has an equal capacitive reactance (Xc) and resistance (ESR) of 10 ohms. The total value of these resistances when measured with an AC ohm meter will be the "VECTOR SUM" of the CAPACITIVE REACTANCE and the INPHASE resistance measured in ohms. In this example the total resistance measured by the AC ohm meter will be 14.14 ohms! This vector sum is always larger than the INPHASE resistance(ESR). For this reason an AC ohm meter is not a valid device to measure ESR unless a means of negating the effects of capacitive reactance is applied. Three equally valid basic techniques come to mind: (1) Raise the AC ohm meters test frequency so high that the capacitive reactance is so low compared to the expected ESR that the effects of capacitive reactance can be ignored. (2) Electronically subtract the capacitive reactance portion of the vector sum from the total measured AC resistance. (3) Use a sample/hold keying circuit and only "key on" the AC ohm meter when the PHASE ANGLE between the measured AC resistance and the test signal is at 0 or 180 degrees (INPHASE). Each technique has its advantages and disadvantages which will be explained below.

Capacitor ESR - Methods of Measurement
by Doug Jones (Capacitor Wizard)
http://www.midwestdevices.com/pdfs/Tnote3.pdf

E = I x ESR

ESR would be easy to measure if it really existed as the single physical resistor represented by the drawing above! Unfortunately you can't just hang two test leads across the ESR resistor in the drawing above and measure the ESR! The drawing is just a symbolic representation of the total effect of the many physical quantities that make up ESR! This article is intended to give the reader an understanding of ESR and an introduction into the complexities of measuring ESR.

ESR is a DYNAMIC quantity of a capacitor. ESR does not exist as a STATIC quantity therefore it cannot be measured by a conventional DC ohm meter. ESR exists only when alternating current is applied to a capacitor or when a capacitors dielectric charge is changing states. ESR can be considered to be the TOTAL INPHASE AC resistance of a capacitor. INPHASE means the current(I) and the voltage(E) if measured across the resistive component ESR(R) of a capacitor are INPHASE with each other, just as if ESR(R) were pure resistance (no capacitive or inductive reactance). ESR includes the DC resistance of the leads, DC resistance of the connections to the dielectric, capacitor plate resistance, and the INPHASE AC resistance of the dielectric material at a particular frequency and temperature. The combination of components that make up ESR are symbolized by a resistor in series with a capacitor as shown above. This symbolic resistor

Line #	Ragt/Ro	Rdt/Ro	IPL in dB	Ql/Qo	Input match: S11 in dB	Output match: S22 in dB
1	0.250	0.250	1.023	0.1111	-19.85	-19.85
2	0.333	0.500	1.761	0.1667	-infinity	-9.53
3	0.667	0.667	2.50	0.250	-12.04	-12.04
4	1.000	0.500	3.01	0.250	-6.02	-infinity
5	0.500	1.000	3.01	0.250	-infinity	-6.02
6	1.000	1.000	3.52	0.333	-9.54	-9.54
7	2.000	0.667	4.77	0.333	-3.52	-infinity
8	0.667	2.000	4.77	0.333	-infinity	-3.52
9	2.000	2.000	6.02	0.500	-6.02	-6.02
10	3.000	3.000	7.96	0.600	-4.34	-4.34

*Return loss is a measure of the "goodness" of an impedance match. A value of minus infinity indicates a 'perfect impedance match' (all the available input power is delivered to the load). A value of zero indicates a 'perfect impedance mismatch' (all of the available input power is reflected back to the source, and none is delivered to the load). An intermediate value indicates the amount of available power that is reflected back to the source by the load. The usual way of referring to mismatch of a two-port network is by using S parameters. S11 is the input reflection coefficient and S22, the output reflection coefficient. The magnitude of a voltage reflection coefficient is 20*log|{(Rload-Rsource)/(Rload+Rsource)}|. One reference is Radiotron Designer's Handbook, Fourth Edition, pp 891-892. In our case, consider out little circuit (at resonance) to be a zero length transmission line having attenuation. Note that the Handbook was written before 'S" parameters were widely used.

Table 1 shows the tradeoff between IPL and selectivity. The lower Rag and Rd become, the lower IPL becomes, but Ql/Qo drops (poorer selectivity). Higher values of Rag and Rd result in greater IPL and greater Ql/Qo (greater selectivity). Ramon

Vargas has suggested that many people consider the parameters on line 4 to be close to a practical optimum, and they are. Line 3 shows alternate parameters for achieving the same selectivity at an IPL 0.51 dB less. A general rule may be stated that for any given value of Ql/Qo, equal values for Rag and Rd will result in the least possible IPL. In this case, input and output return losses will be equal. It appears to me that the parameters in lines 3 or 6 are probably the ones to shoot for in most design calculations.

An effect I have observed is that one cannot simultaneously attain a perfect impedance match at both input and output ports in systems of the type shown in Fig. 1a (simultaneous conjugate match). As shown on lines 4 and 5 in Table 2, Ragt and Rdt can be arranged to provide a perfect match at one port, but then the other port will be mismatched. Ro could be replaced by a series resistor (not a real-world crystal radio set anymore) and one would still not be able to arrange a simultaneous impedance match. If circuit losses were represented by proper values of both series and shunt resistances, it would then be possible to attain a simultaneous perfect match at both input and output.

An audio transformer is one passive device that has series and shunt loss components. If it is so designed, it can provide a simultaneous perfect impedance match at both input and output when loaded by its designed-for load resistances (ignoring reactance effects). If operated at any other impedance level, such as doubling the source and load resistances, simultaneous perfect input and output impedance matching cannot be attained. This info is of mainly theoretical interest for most crystal radio set applications except when one tries to operate an audio transformer at considerably higher or lower source and load resistance values than it was designed

5. WELSBY V G : 'The Theory and Design of Inductance Coils', Macdonald & Co, Second edition 1960.

6. The Plumbers Handbook : http://www.kembla.com/assets/Uploads/general-PDFs/The-Plumbers-Handbook.pdf

Issue 1 : August 2013
Issue 2 : September 2013 : the addition of Annex 1 Broadcast Capacitor

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for. If this is done, IPL is considerable increased compared to using it the impedances for which it was designed.

A6. Audio transformer loss. See Articles #1 and #5.

Part B: The Measurements.

Loss measurements at various frequencies on some components, with the loss expressed as a parallel resistance (Rp) in parallel with the capacitance of the component or the loss tangent, or equivalent Q of the component:

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B1: Q and equivalent parallel loss resistance (Rp) of two variable capacitors vs frequency when resonated with a 250 uH inductor, in parallel with a total circuit stray circuit capacitance of 20 pF. Rp and Q vs. frequency at four different capacitance settings is also shown.
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B2: Equivalent parallel loss resistance of some DPDT slide switches.
*
B3: Loss tangent of some coil form and sheet plastic (front panel) materials.
*
B4: Q of inductors and L/C resonators.
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B5: High Q fixed value ceramic capacitors.

B1: Measurements at 520, 730, 943, 1300 and 1710 kHz, made on two different variable capacitors used in crystal radio sets, are shown in the two graphs below. Capacitor A is a 485 pF variable capacitor that was purchased from Fair Radio Sales. It has a ceramic insulated stator and silver plated brass plates with silver plated wiper contacts. Capacitor B is a small 365 pF air-variable capacitor purchased from the Xtal Set Society. Its plates are made of aluminum and the stator support insulators are made of a phenolic plastic. This capacitor is similar to those sold by Antique Electronic Supply and others. See note at the end of this section B1.

Fig. 2 shows Rp plotted against frequency, with the capacitor adjusted at each frequency to a value 20 pf lower than that required to resonate a 250 uH inductor. This allows for a stray capacitance of 20 pF, in an actual circuit. Any losses that may be in the stray capacitance are assumed assigned to the inductor. The plot shows how, in actual practice, the Rp of variable capacitors A and B vary when tuned across the broadcast band. Fig. 3 shows how the Q of the total capacitance (including the 20 pF stray capacitance) varies across the BC band for each capacitor. Do not make the mistake, when looking at the two graphs below, of thinking that they represent Rp and Q of capacitors A or B vs. frequency, with the capacitor set to a fixed capacitance. The capacitance is varied as a function of frequency, along the horizontal axis of the graphs, to a value that would resonate with a 250 uH inductor. Figs. 4 and 5 show Rp and Q of capacitor B as a function of frequency, at four different fixed capacity settings (the frequency at each capacitance setting is always a value that brings about resonance with the 250 uH inductor).

The higher loss of this capacitor compared to that of the variable capacitor can be partially attributed to the difference in conductivity – copper versus silver, but this would give a difference of only about 6-7%. The additional loss is probably due to the longer conducting path in the fixed capacitor, because the connection to the outer tube was half way down its *outside* surface, so current to the inside had about twice the distance to travel. It became clear that even if the conductors were silver plated and the construction changed to minimise the conduction path it was unlikely that a loss significantly lower than the variable capacitor would be achieved, which was the aim.

11. References

1. MOULLIN E B : ‘A Method of Measuring the Effective Resistance of a Condenser at Radio Frequencies, and of Measuring the Resistance of Long Straight Wires’. Proc. R. Soc. Lond. A. 1932 137 116-133 doi:10.1098/rspa.1932.0125 (published 1 July 1932).

2. JACKSON W : ‘The Analysis of Air Condenser Loss Resistance’ Proc IRE, Vol22 Number 8, Aug 1934.

3. FIELD R F & SINCLAIR D B : ‘A Method for Determining the Residual Inductance and Resistance of a Variable Air Condenser at Radio Frequencies’. Proc IRE, Vol 24, Number 2, Feb 1936.

4. BOCK E M : ‘Low-Level Contact Resistance Characterisation’, AMP Journal of Technology, Vol 3, November 1993.

If the plate is silver (as here) , then p=1.629 10-8, and u = 4pi10-7, so Zwall = 8 10-4 at 10 Mhz. Evaluating ***l*** and w is somewhat difficult, but if we take semicircular plates of radius r, then when the capacitance is maximum the rotor and stator plates will fully overlap. The average length from the centre of the rotor (where current enters) will be r/2 and the average width (taken as the average circumference) will be $2\pi(r/2)/2$, so the ratio ***l*** / w will be $1/\pi$. As the spindle is rotated the overlap between the rotor and stator will reduce in direct proportion to the capacitance, so ***l*** / w will vary as:

$$l \,/\, w \approx 1/(\pi \; C/Cmax) \qquad \text{A1.2}$$

So at maximum capacitance we can expect the loss resistance in the plates to be, Rv = 2 Zwall / (πN) = 0.034 mΩ at 10Mhz, for N=15. This is a factor of around 1000 less than that measured, and so any variation with spindle rotation would be insignificant, and as experiment confirms.

Appendix 2 : Standard Capacitor

An attempt was made to produce a standard capacitor, having a very low loss. This was made using two coaxial copper pipes, 15mm and 22mm diameter, about 100mm long. The insulation between the two pipes was two bands of PTFE tape, located at the ends of the tubes. Capacitance was around 25pf.

However this capacitor had a lower Q than the variable capacitor by about 12%. It was thought the PTFE tape may be lossy (it was plumber's tape), and it was replaced with thin expanded polystyrene foam, but with the same results.

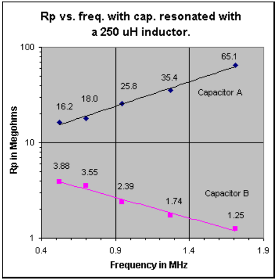


Fig. 2

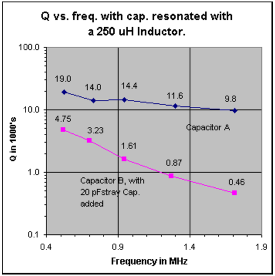


Fig. 3

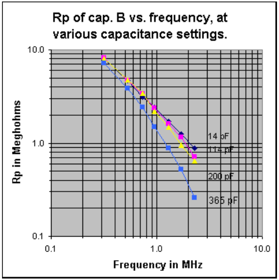


Fig.4

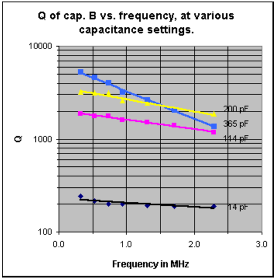
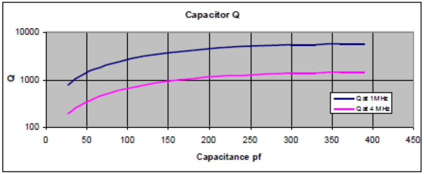


Fig. 5

capacitor uses a relatively inexpensive insulator, such as porcelain with a loss tangent of 0.0075, then the capacitance of its insulators is 4.8 pf (see Section 8.5), around half the minimum capacitance.

The contact resistance is also seen to be much larger at 32 $m\Omega$ compared with 10 $m\Omega$. However, the metallic loss is about 1/3rd that of the Photo 1 capacitor (0.0039 cf 0.01), despite the metalwork being aluminium rather than silver. This lower metallic loss is because of the very much shorter path which the current takes. Its Q at 1 MHz and 4 MHz is given below :



Appendix 1 : Estimated Loss in the Plates

The resistance of two plates will be twice that of one plate, and if we have N pairs of plates the resistance Rv will be :

Rv = 2 Zwall ***l*** / (N w) A1.1

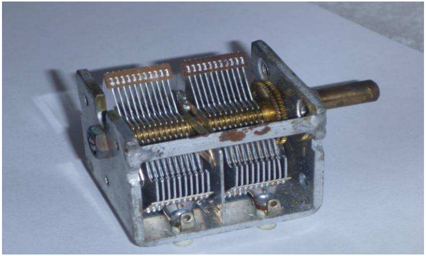
where Zwall = p /d = (p d u f)0.5
p is the resistivity of the metal
u is the permeability of the metal
l is the length of the current path through the plate
w is the width of the current path through the plate

Subsequent to the above work, the loss of the broadcast-band tuning capacitor shown below was measured, and gave the following equation for its series resistance:

Rcap = 0.032+ 5800 /(f C2) + 0.0039 (f 0.5) ohms   10.1

where f is in MHz and C in pf

The accuracy of the above equation is estimated at ± 13%.



Photograph 3 : Broadcast-band Capacitor

This capacitor has 2 sections each with a range of 10 pf to 380 pf, and the above equation is for one section.

It is interesting to compare the losses of the capacitor shown in Photo 1 with that of this broadcast-band capacitor. In common with most capacitors of this type it has a very short insulation path and this accounts for the high dielectric loss factor of 5800 in the above equation, over 7 times the loss of the capacitor shown in Photo 1. Assuming this broadcast-band

Discussion:

There are two main sources of loss in an air-dielectric variable cap: 1) Loss in the dielectric of the stator insulators, and 2) Resistive losses in the metal parts. Of course, there is also the important very low-loss capacitor made from the air dielectric between the plates. The losses in 2) include resistive loss in the plates and in the wiper that connects the rotor to the frame. Resistive loss in the plates is very small at low frequencies, but increases with increasing frequency because of skin effect.

Notice, that for capacitor B, Rp (vertical scale) in Fig. 4 is about the same for all capacitance settings at the lowest frequency plotted (310 kHz). Rp is approximately constant over the full 14-365 pF capacitance range. This is because we are varying a high Q air dielectric capacitor of very low loss, in parallel with a low Q fixed capacitance made up of the lossy phenolic stator supports. Most of Rp comes from the loss in the phenolic. In Fig. 5, again at the low frequency end of the horizontal axis, observe that Q is a direct linear function of the overall capacitance, as it should be. The Q of the air-variable part, taken alone, is much higher than that of the capacitor made up of the phenolic insulators. The main loss, here, comes from the approximately fixed shunt Rp provided by the phenolic stator supports. At 14 pF the C from the air cap is relatively low compared to that from the phenolic insulators. At 365 pF the C from the air cap is much higher than that from the phenolic supports.

Things change at higher frequencies. The reactance of the capacitor drops. This, combined with the series resistance of the plates and wiper now come into play as an additional factor reducing the Q. If this series resistance (Rs) were the only resistance affecting Q, the equation for Q would be: Q=(reactance of the capacitor)/Rs. One can see from this equation that the introduction of Rs makes the Q drop when frequency increases. Up to now, the main loss came from the parallel loss resistance of the phenolic supports. The Q from a setting of 365 pF drops quite rapidly with increase of frequency because of this series resistance. Skin effect makes the effect worse by increasing Rs, the higher one goes in frequency. Notice that at the low frequency end of Fig. 5, Q is approximately proportional to the capacitance, as it should be if the main loss is the fixed shunt resistance Rp, coming from the phenolic stator insulators.

Fig. 5 shows an approximately constant Q (vs. frequency) when the capacitor is set to 14 pF. The loss causing this low, constant Q, comes from the loss tangent of the capacitor formed from the phenolic stator insulators. At low frequencies, Q increases when the capacitor is successively set to 114, 200 or 365 pF because engaging the plates adds a high Q air dielectric capacitor component in parallel with the low Q capacitor formed from the phenolic dielectric supports. As frequency increases, with the capacitor set to 365 pF, one can see that the Q drops at a faster rate than it does when set to 200 or 114 pF. This is because Rs (being in series with the air capacitor, that dominates at the 365 pF setting), acting with its lower capacitive reactance results in a capacitor of lower Q (Q=reactance of air capacitor at 365 pF/Rs).

In the Figs. 2 and 3, the capacitor is always set for a circuit capacity value that resonates with 250 uH. This means that at 520 kHz, the varicap is set so the circuit capacity is 375 pF. At 1710 kHz, the circuit capacity is set to 34.7 pF. Even though the capacitor Q goes from 19,000 to 9800 as frequency goes from 520 to 1710 kHz, Rp increases as frequency increases

The series resistance of a variable air capacitor at radio frequencies is shown to be given by :

$Rcap = Rs + \alpha /(f\ C2) + \beta\ (f\ 0.5)$ ohms 9.1

where Rs , α and β are constants for any particular capacitor.

f is the frequency and C the capacitance

The first factor Rs is due to the contact resistance, and this is largely due to the wiping contacts. The second factor is due to the dielectric loss in the supporting insulator, and the third factor is due to the resistivity of the metalwork, particularly that supporting the plates.

Measurement of this resistance requires the capacitive reactance to be tuned-out with a low loss inductor and it is shown that this can be achieved with straight copper pipes arranged as a shorted two-wire transmission line.
For the capacitor shown in Photo 1, the series resistance is given by:

$Rcap = 0.01 + 800 /(f\ C2) + 0.01\ (f\ 0.5)$ ohms 9.2

where f is in MHz and C in pf

The accuracy of the above equation is estimated at ± 13%.

10. ANNEX 1 : BROADCAST CAPACITOR

In respect of dielectric loss, Jackson says 'The overall capacitance C of the capacitor may be regarded as composed of a fixed and imperfect portion C1 due to the supporting insulation, in parallel with a variable air portion devoid of energy loss. The power factor tan δ of the portion C1 seems usually to be substantially constant over a wide frequency range'.

It is interesting, although not particularly useful, to calculate capacitance of the supporting insulation, C1. This can be determined from Equation 6.1.2, by setting C = C1, since then the loss is not diluted by the air capacitor and the series resistance due to the dielectric loss will then be :

$$Rc = 800/(f\ C1^2) \qquad 8.5.1$$

Also from the definition of tan δ :

$Rc = \tan\delta /(2\pi f\ C1)$
for f in Hz and C in farads

$= 10^6 \tan\delta /(2\pi f\ C1) \qquad 8.5.2$
for f in MHz and C in pf (as equation 8.5.1)

Equating 8.5.1 and 8.5.2 :

$$C1 = 2\pi\ 800 / (10^6 \tan\delta)\ \text{pf} \qquad 8.5.3$$

The material of the insulator is unknown but is possibly Steatite, for which tan δ = 0.003, and then the capacitance C1 due to the supporting insulation is 1.7 pf.

9. SUMMARY

because the circuit capacity must be reduced from 375 to 34.7 pF to tune the tank from 520 to 1710 kHz (Rp=Q/(2*pi*f*C).

If the question is posed: 'Is it more important to have ceramic insulated stators or silver plated plates on a variable capacitor used in a BC band crystal radio set?', the answer is that ceramic insulated stators are the way to go.

Capacitor A: The main practical conclusion that can be taken from Fig .2 above is that Rp of capacitor A is very high over the whole band and varies roughly proportionally to frequency, over the frequency range of interest: 520-1710 kHz In fact, Rp is so high that it will not contribute any appreciable loss even when used in high performance crystal radio sets using a high Q tank inductor. Another plus is that its Rp increases with increasing frequency, further reducing any effect on loaded Q, sensitivity or selectivity at the high end of the band. The silver plating on the brass plates is beneficial because the resistivity of silver is about 25% of that of brass. Practically speaking, the silver should have little effect on the operation of a crystal radio set in the BC band, but short wave is another matter. The Q of the capacitor drops at higher frequencies, especially when set to a high capacitance value. Silver plating can materially improve performance at higher frequencies by providing a needed higher Q. Be aware that some capacitors are made with cadmium plated brass plates. Silvery-whitish colored cadmium has a resistivity 4.6 times that of silver. This higher resistivity will somewhat reduce Q at high frequencies and at high capacity settings. Some people mistakenly assume, because of the silvery-whitish color, that cadmium plated plates are really silver plated.

Capacitor B: One can see from Fig. 2 that Rp of capacitor B varies approximately inversely with frequency. The DLF (dielectric loss factor) of the phenolic stator support insulators is the main cause of this loss, over the whole frequency band. Towards the high end of the band, some loss is contributed by the series resistance in the capacitor plates, the rotor shaft wiper contact and skin effect. This loss effect is greater than that in capacitor A because the resistivity of the aluminum plates is 1.7 times that of silver. Practically speaking, this effect is minimized at the high end of the band because the plates are mostly disengaged. The Rp of the capacitor will have its greatest effect in reducing sensitivity and selectivity at the high frequency end of the band because that is where its value is the the lowest. See note at the end of this Article.

The usual crystal radio set uses shunt capacitor tuning with a fixed tank inductance. This configuration causes the tank reactance to be highest at the high end of the band, thus further reducing loaded Q, sensitivity and selectivity, for a given value of Rp. At the low and medium frequency parts of the band, Rp of the capacitor is so high that its effect is small in many crystal radio sets. Highest performance crystal radio sets made with high Q inductors, with careful attention to impedance matching may experience a noticeable reduction in sensitivity and selectivity at the high end of the band when using this or other capacitors using phenolic stator insulation. This is because the Rp of the capacitor becomes comparable to the higher equivalent Rp of the high Q tank inductor at the high end of the band as compared to its value at the low end. To clarify this, an ideal condition would exist if Rp of the capacitor and inductor were infinite. If this were to be the case, and good impedance matching of antenna-ground system to tank, diode to tank and headphones to diode existed, all of the power intercepted by the antenna-ground system would be delivered to the headphones and maximum sensitivity would occur. Any loss present in capacitors, inductors or

remaining 30% is grooved. So we might anticipate the loss increasing considerably, and if it is assumed by a factor of 4, to 12 mΩ, this agrees well with the measurements as the following shows : a similar situation exists for the moveable vanes, and they are supported by the central spindle and this has very deep grooves. So by the same argument its resistance is estimated to be 15 mΩ, giving a total due to metallic loss of 27 mΩ, close to the measurements of total metallic loss of 28 mΩ at 10 MHz.

So it seems that the current flow in the support pillars is seriously interrupted by the presence of the plates, and this is a major source of loss.

8.4. Conductor Loss Budget

Bringing together the estimated conductor losses discussed above gives the following for 10 MHz (all figures are very approximate) :

Loss in vanes	0.03 mΩ i.e. negligible
Pillar supporting static vanes	12 mΩ
Shaft supporting moving vanes	15 mΩ.
Slip ring contact resistance	10mΩ.
Two solder tags	1 mΩ
Total	38 mΩ

The measurements indicate a total of 42 mΩ, so the above seems a reasonable estimate.

8.5. Dielectric Loss and Capacitance of Plate Supports leaving about 4.5 mΩ attributable to each of the sliding contacts. This is somewhat larger than Bock's measurements (ref 4) of contact resistance of silver contacts of 1 mΩ, but the contact force here may be lower than his 100 gm, and also the contacts could be worn.

8.2. Loss in Plates

There is no indication in the measurements that the loss varies with capacitance setting, and so the loss in the plates is presumably very small, and this is supported by the analysis given in Appendix 1. Field & Sinclair (ref 3) found the same and stated 'the change in series resistance with rotor displacement is therefore small despite the change in current distribution in the plates'.

8.3. Loss in Plate Support Pillars

The static plates are supported by two silver-plated cylindrical bars, which are bolted to the end ceramic. If only one pillar is connected to the external circuit (the normal configuration), then the whole of the current flows through this pillar. Its diameter is 6.5 mm, and it has an effective length of 70 mm, and at 10 MHz such a bar would have a resistance of around 3 mΩ, assuming the whole circumference carries current. However part of the circumference is interrupted by the plates, and these will seriously affect the current flow. To see this, imagine a cylindrical conductor with circular washers along its length, forming a structure similar to a thermal radiator. Current flowing from one end to the other will now have to flow outwards across one face of each washer, across its thickness, and down the other face, and this will increase the resistance considerably. In the capacitor here, around 70% of the circumference is interrupted by plates and in addition the

transformers reduces sensitivity. In this discussion we are dealing primarily with losses in the resonating capacitor and tank inductor, both referred to as their respective values of Rp. The higher the value of these Rps, compared to the transformed antenna-ground source resistance across which they appear, the lesser the loss they cause. One approach to counter the effect the drop in Rp as frequency increases is to change to two-step inductive tuning by dividing the band into two sections as described in Article #22.

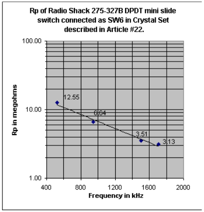
Technical Note: At any frequency, a real world capacitor of value C1 having a Q of Q1 (Q1>10), can be quite accurately modeled as a series combination of an ideal no-loss capacitor of value C1 and a resistor (Rs) equal to: (reactance of C1)/Q1. Alternatively, at any frequency, a capacitor of value C1, having a Q of Q1 (Q1>10) can be modeled as a parallel combination of an ideal capacitor of value C1 and a resistor (Rp). The resistor, Rp in this case, has a value of:: (reactance of C1)*Q1. Note that since capacitor reactance is a function of frequency, the value of the resistor will, in general, vary with frequency. In crystal radio set design it is sometimes convenient to model the tuning capacitor loss as a parallel resistor, other times as a series resistor.

Credit must go to Bill Hebbert for making the time consuming, difficult, precision measurements required for Figs. 2-5.

B2: Slide switches used in the Crystal Radio Set described in Article #22 have dielectric losses, as do all switches. To get a handle on this loss, samples of several different types of DPDT switch were measured at 1500 kHz. Each switch except the last three, below, were wired as a SPDT unit by paralleling the two sections. The Q of the capacitance appearing across the open contacts was then measured and Rp was calculated. Rp usually varies approximately inversely with frequency and therefore causes more loss at the high end of the band than at the low end. Contact resistance of all switches was found to be very low. It is unknown how well that characteristic will hold up over time. Note the extremely low loss of the Switchcraft 56206L1. The loss is so low that this switch is overkill in most crystal radio set applications. For applications in which the crystal radio set builder wants to use the lowest loss DPDT slide-switch available, this switch is the best I have found.

Table 3 - Equivalent parallel Loss Resistance (Rp) caused by the loss tangent of the dielectric of some DPDT Slide Switches at 1.5 MHz.

Brand name, model number and Size of switch	Rp in Megohms
ARK-LES (std. size)	2.3
Radio Shack 275-403A (std. size)	3.5
Radio Shack 275-407A (sub mini)	4.1
Stackpole S7022X (std. size)	4.9
Switchcraft 46206EE-6 (std. size)	5.1
Radio Shack 275-327B (mini)	5.7
CW (mini)	6.0
Switchcraft 46206LR (std. size)	6.5
C&K L202-1 (mini)	6.8
Switchcraft 56206L1 (mini)	13.3
Radio Shack 275-327B connected as SW3 in Article #22 (HI band)	4.1
Switchcraft 56206L1 connected as SW6 in Article #22	9.4
Radio Shack 275-327B connected as SW6 in Article #22, **See Graph--->>**	3.5



Rotary selector switches using ceramic insulation should have very low loss, even lower than the Switchcraft 56206L1. Quality switches using brown phenolic insulation probably have losses similar to slide switches using similar material. I

random uncertainty in bandwidth of around ± 1%. These random errors and quantisation errors were reduced by using the analyser's averaging of 16 measurements, to gave an overall precision (narrowness of scatter) in the measurement of Q of less than ± 1%.

The *accuracy* of the bandwidth measurements is less easy to estimate, but the errors in the AIM 4170 are probably less significant than other practical matters, such as small changes to the experimental set-up, such as the size of the coupling loop with respect to the tuned circuit. These practical aspects probably account for ±5%. In addition there is the ±1.5% error in matching the measurements (para 6.1) to the model.

So the overall measurement accuracy is assessed at ±9%, for the measurement of the overall Q and its comparison with theory. Since the capacitor resistance is about half the total this translates to an uncertainty in the capacitor resistance of ±13% (assuming the errors add in quadrature).

8. ANALYSIS OF LOSS FACTORS

8.1. Contact Resistance

There are a number of contacts contributing to the contact resistance. There is a sliding contact to the rotor, with two contacts in series, and for connection to the external circuit there is a solder tags bolted to each of the rods supporting the stator plates, and another to the static part of the sliding contact. The measurements show a combined contact resistance of 10 mΩ. The pressures on the solder tags are likely to be much greater than the slider pressure, and so their contact resistance will probably be of the order of 1 mΩ each or less, The proximity multiplier and the radiation resistance contribute a very small resistance and so uncertainties here will have a negligible effect on the results.

The major contributor is the resistance of the copper pipes, and the main uncertainties here are the resistivity and permeability of the copper, and each have an uncertainty of around ±6 %, so the uncertainty on Rwall is ±6 %. Also there was an uncertainty in the room temperature of ± 5o during the experiments, giving a further uncertainty in resistance of ±1%.

The SRF multiplier increases the resistance by only 5% and so the uncertainty here is say ±0.5% of total resistance.

The leads to the capacitor contribute a surprisingly large 15% to the total resistance, but it is very difficult to calculate this accurately and the uncertainty to the total resistance is assumed to be ±1.5 %

These uncertainties are uncorrelated and so they can be added in quadrature, giving an overall uncertainty in the *calculations* of ±7%.

The measurements also have uncertainty and error. Measurements were of Q and carried-out on an AIM 4170 analyser. This measures Q by taking the two frequencies either side of resonance where the phase is 45o, and so the error here is related to the uncertainties in measuring frequency and phase. The analyser gives frequency to 6 places, for example 24.3652 MHz, but with a Q of say 1000, the bandwidth will be only 0.0244 MHz. Assuming the last figure is uncertain (i.e. ±0.0001) this will give an random uncertainty on the measurement of bandwidth of around ± 0.5%. The uncertainty on phase is not known, but if we assume ±1o, this will give a

would expect that the slope of the Rp vs. frequency graph of the other slide switches to be the similar that shown above.

B3: The loss tangent of an insulating material is the reciprocal of the Q of a capacitor made of that material. Some of the insulating materials listed below are used as front panels, detector stand bases, wire insulation and coil forms in crystal radio sets. A capacitor formed by the use of one of these materials, connected across a high impedance point and ground, will contribute a loss proportional to the loss tangent and capacitance.

Table 4 - Loss Tangent of some Insulating Materials used in Crystal Radio Sets, Measured at 1.5 MHz.

Dielectric Material	Loss tangent	Q of a capacitor using the material
PVC as used in coil form	0.017	59
PVC wire insulation	0.03	33
High-impact styrene coupler from Genova Mfg. Co. (opaque white)	0.0017	590
Polypropylene 1.5" diameter drain pipe from Genova Mfg. Co.	0.0022	452
High impact styrene sheet, 0.1" thick (opaque white and flexible)	0.0023	430
Plexiglas 0.115" thick	0.016	55
FR-4 PCB material 1/16' thick	0.027	37
Black 3/16" Condensite panel, brand name "Celoron" (Bakelite, new old-stock radio panel from the '20's)	0.035 (@ 0.8 MHz)	29
ABS styrene (black)	0.010	100
Garolite (black)	0.033	30
ABS styrene (light beige)	0.020	50
HDPE (milky white)	0.0009	1120
Polystyrene (light brown opaque)	0.0032	320

Note: GE's version of polycarbonate is called Lexan. A review GE's spec. sheets of various grades of Lexan show loss tangent values at 1.0 MHz ranging between 0.006 and 0.026 . Many grades are specified 0.01.

B4: Please see Parts 10 and 11 of Article #26 as well as Table 4. Also see Table 2 of Article #22 and Article #29.

B5: Sometimes, when working with high Q tank circuits, a need pops up for a fixed capacitor with a value between say, 100 and 1000 pF that will not degrade overall circuit Q. Generic NPO disc capacitors in that range usually have a Q of around 2000-3000 at 1 MHz. Table 4 shows some caps having higher Q values. The only downside to the high Q caps is that they are SMD types and require some skill when soldering pigtail leads to them to easy connecting to one's circuit. The capacitors were measured singly, in parallel or in series, aiming for values approximating 500 pF. This was for convenience in measurement. I used solid tinned copper wire having a diameter of about 0.010" for my pig-tails. The source for the strands was a piece of stranded hook-up wire.

Table 4 - Q of easily available capacitors in the 100 - 1000 pF range

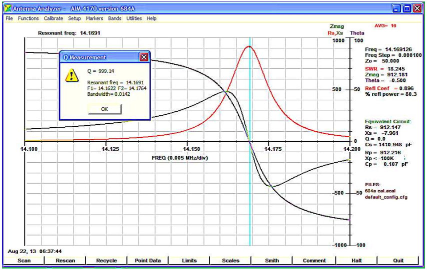
	Type	Value in pF	Voltage	Q at about 920 kHz	Mfg	Mfg. part number
1	Polypropylene	Two 1k in series=500	630	2,000	Xicon (Mouser)	1431-6102K
2	Polystyrene	One 470	50	6,200	Xicon (Mouser)	23PS147
3	Generic NPO disc	One 220	500	2,800	3/8" dia.	NA
4	Multilayer, hi Q SMD	Two 1k in series=500	50	8,000	Murata	ERB32Q5C1H102JDX1L
5	Multilayer, hi Q SMD	One 470	100	18,000	Murata	ERB32Q5C2A471JDX1L
6	Multilayer, hi Q SMD	Two 220 in parallel=440	200	20,000	Murata	ERB32Q5C2D221JDX1L
7	Multilayer, hi Q SMD	Two100 in parallel=200	500	40,000	Murata	ERB32Q5C2H101JDX1L

Note: Solder flux contamination on a dielectric is the enemy of high Q because it usually provides a resistive leakage path. If one gets solder flux on the insulation of a variable cap or switch, remove it with a commercial flux remover. This is important when a DX crystal radio set is involved.

#24 Published: 03/25/2002; Revised: 06/10/2008

6.4. Typical Resonance

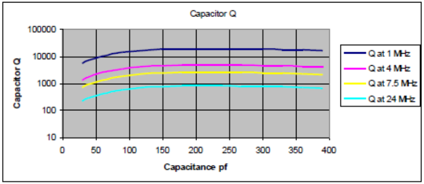
A typical resonance curve is shown below, from which the vector analyser calculates the Q.



7. ACCURACY

To evaluate the likely uncertainty, it is useful to consider the individual contributions to the loss of the measurement circuit. At 10 MHz these are :

Copper Pipes (ignoring proximity and SRF) :	0.04 Ω
Proximity multiplier on above :	1.009
SRF multiplier on above :	1.048
Radiation resistance :	0.0005 Ω
Lead resistance :	0.0061 Ω



6.2. Test of Ground Effects

The measured Q was as large as 1000, and so it is possible that the floor of the room and other surfaces could be introducing loss. To evaluate this an aluminium sheet, approx 0.8mx 0.5m, was placed on the floor below the tuned circuit, a distance of about 0.8m. A measured Q of 953 was unaffected, to within the repeatability of the measurement.

6.3. Standard Capacitor

The measurement uncertainty could be reduced if a fixed capacitor could be made with a very low loss since this could be used as a standard to calibrate the system. However attempts to make such a capacitor were not successful in that it had a higher loss than the variable capacitor to be measured (see Appendix 2).

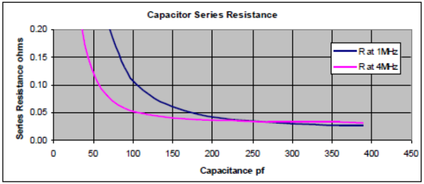
Capacitor Losses

Conrad Hoffman, 2012
http://conradhoffman.com/cap_losses.htm

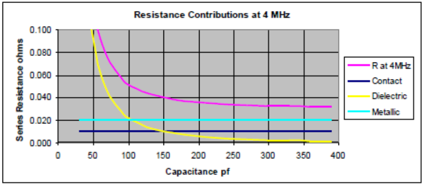
Dielectrics

Material	Dielectric Constant	Dissipation Factor	Dielectric Absorption	Temp Coefficient /°C	Notes
Vacuum	1	0	zero	0	high power RF
Air	1.0006	0	zero	0	RF & variables
Teflon	2	0.0001	low	-120 ppm	high rel mil, high end audio
Polystyrene	2.5	0.0001	low	-150 ppm	max temp 85°C/very low DF, very low DA
Polypropylene	2.3	0.0002	low	-120 to -200 ppm	higher temp sub for styrene
Polyester	3.2	0.016	medium	600 to 900 ppm	popular general purpose
Polycarbonate	3	0.01		65 to 100 ppm	
Paper/wax	2.5				obsolete
Paper/oil	4				AC/HV, NOS replacement
Mica	5-9	.0001-.0007	medium	0 to 200 ppm	stable RF and GP, but unexpected dielectric absorption for its DF
Porcelain	6			0 to 90 ppm	HV insulators
Bakelite	4.5-7.5				obsolete, rare, may char
Glass	4.5-7.0	0.002		140 ppm	rad hard, high stability
AlO_2		.15 varies	high	poor	aluminum electrolytic
AlO_2, pure	9.8	0.0002			insulators
MnO_2		.15 varies	high		tantalum electrolytic
Ceramic, C0G	75 typ.	0.001		±30 ppm	low tempco RF/GP
Ceramic, X7R	3000 typ.	0.025		±15% over temp range	RF/GP
Ceramic, Z5U	8000 typ.	0.03		+22 to -56% over temp range	high value RF/GP bypass only
Ceramic, PZT-8	1000	0.004			hard low sensitivity piezoelectric
Ceramic, PZT-4	1300	0.004			med. sensitivity piezoelectric
Ceramic, PZT-5H	3400	0.02			soft high sensitivity piezoelectric

Capacitors are constructed of two or more electrodes, separated by a dielectric. The dielectric is commonly ceramic, plastic film, oiled paper, mica, or air. Each one has advantages and disadvantages in regards to dielectric constant, losses,

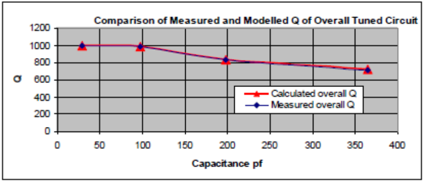


It is useful to know the contributions which each factor in Equation 6.2.1 makes to the overall resistance, and for f = 4 MHz this is shown below:



This shows that at low capacitance the dielectric loss dominates and at high capacitance the metallic loss dominates.

The capacitor Q is given by 1/ (ω C Rcap), and is given below:



The agreement is within ±1.5%.

It should be noted that the frequency is not constant in the above graph, and is 24.5 MHz at 27pf, 10.2 MHz at 198 pf, 14 MHz at 98 pf and 7.56 MHz at 365 pf, all frequencies corresponding to resonance with the inductor. These combinations of capacitance and frequency give relatively high losses, so that the overall Q to be measured is not excessive (below 1000). Also, when the capacitance is set to its minimum of 27 pf the dielectric loss is dominant, even at 24.5 MHz, and so allows the constant α to be determined fairly accurately. Similarly at 365 pf and 7.56 MHz the metallic loss and contact loss dominate permitting these to be determined. So the series resistance of this capacitor is given by:

Rcap = 0.01+ 800 / (f C2) + 0.01(f 0.5)   6.1.2

This equation is plotted below, for f = 1 MHz and 4 MHz:

temperature coefficient, and, of course, cost. High dielectric constants result in smaller capacitors, but usually with poorer properties than the lower constant materials. Some properties of various dielectric materials are shown above:

The data in this table comes from many sources, not all of which are in agreement. If you have better information, want to fill in any blanks or have additional entries, please email me with the data and its source.

Low Frequency Losses

How do we specify loss? If you ask most engineers about capacitor loss, they will mumble something about "loss tangent", then disappear for an emergency coffee refill. There are several different ways of expressing capacitor losses, and this often leads to confusion. They are all very simply related, as shown below.

If you drive a perfect capacitor with a sine wave, the current will lead the voltage by exactly 90°. The capacitor gives back all the energy put into it on each cycle. In a real capacitor, the current will lead the voltage by a bit less than 90°. The capacitor will dissipate a small fraction of the energy put into it as heat. Real capacitors can be modeled, at least to a first order, as a perfect capacitor in series with a resistance. This resistance is referred to as the effective series resistance or ESR, and is only valid for a single frequency. For the examples below, assume a 0.47µF capacitor, driven at 5000 Hz, 35 volts RMS, and showing a phase angle of -89.5°. Note that the relationships shown only apply to sine waves. Capacitors can also be modeled using parallel elements, but we'll limit our discussion to the series model. You'll find a collection of "handy formulas" on this site that includes conversions between series and parallel models, plus other useful data.

Dissipation factor, or "D" as it is usually marked on test bridges, is the tangent of the difference between the phase angle of a perfect capacitor, and the capacitor in question. In our example, -90°- -89.5°=-0.5° The tangent of -0.5° is -0.00873. We take the absolute value so D=0.00873. Since this number is directly read from most test bridges, other parameters are often calculated from it. It is also known as the loss tangent and is sometimes expressed as a percentage. D=.873%

This is probably a good time to mention that by general agreement capacitive reactances are negative and inductive reactances are positive. A vector impedance meter would display the phase angle of the capacitor in question as "-89.5". Xc, the capacitive reactance would be -1/(2*pi*f*c) or -67.7255 ohms. The sign is often omitted.

Power factor, or "PF", is less common than it used to be, at least as applied to capacitors. It is the cosine of the phase angle itself. In our example, COS(-89.5)=0.00873. Note that for small angles, PF is essentially equal to D, and both are approximately equal to the phase angle expressed in radians. For large angles the situation is quite different. A power factor of 1 is 100% resistive loss whereas D can exceed 1 and approaches infinity for 100% loss.

Q is the quality factor, a dimensionless figure of merit. It is the reciprocal of D. In our example, 1/.0087=114.58, so Q=114.58.

6.1. Procedure and Initial Measurements

The Q of the total circuit was measured with an AIM 4170 network analyser. This analyser could have been connected across the variable capacitor, but the resistance of the tuned circuit at resonance is very high at this point (around 60kΩ) and the analyser is not very accurate at this level. In addition the analyser will introduce stray capacitance which will upset the measurements. The analyser was therefore connected to a 'tap', on the coil. For this the ground terminal of the analyser was connected at a point which was assumed to be at zero potential wrt ground i.e. half-way across the end pipe forming the transmission-line short. The active port of the analyser was connected to one of the copper pipes at a distance of 150 mm from the end of the pipe, and the connection lead taken down the centre of the transmission line, forming a small loop (see Photograph 2). This gave a resonant resistance at the analyser of around 700 Ω (at 14 MHz). The measurement of Q was not particularly sensitive to the exact dimensions of this coupling loop (≈ ± 5%), but the dimensions above gave the highest Q.

The measured Q was compared with the theoretical Q given by:

$$Q_{theory} = \omega L / R_s \qquad 6.1.1$$

Where L is as measured, or as given by Equation 4.4.1
Rs is given by Equation 5.1

The capacitor resistance Rc was assumed to conform to Equation 2.1, and the constants Rs, α and β were adjusted in the model to give the best correlation between the measured Q and the theoretical Q, and this was achieved with Rs = 0.01, α = 800, and β = 0.01, to give the following :

Rr is the radiation resistance = 31200 $(A/\lambda2)2$
Rleads is the resistance of the leads connecting the capacitor
Rcap is series resistance of the capacitor being measured, see below.

It was assumed that the solder joints connecting the pipes together had negligible resistance, as did those connecting the metal tape to the capacitor.

Plumbing pipes were used here and the copper in these has a higher resistivity than normal copper wire, and lies between 1.92 10-8 and 2.30 10-8 Ωm (ref 6). The average of these is 2.11 10-8 and this was used here. The relative permeability μr of pure copper is unity, and it is assumed to be the same for plumbing copper.

The frequency fr was derived from the measurement of inductance as 60 MHz (para 4.3).

The capacitor was connected to the copper pipes via copper straps, having a combined length of 80 mm and a width of 11 mm and thickness 0.25 mm. The resistance of these was calculated from :

Rstrip = Rwall l / [2(w+ t)] Kfringing   5.2

where Kfringing = 1.06 +0.22 Ln w/t +0.28 (t/w)2

The fringing factor allows for the increased resistance due to the current crowding towards the edges of the strip.

6. MEASUREMENTS

Effective series resistance, or "ESR" is the value of resistance in series with a perfect capacitor that produces the phase angle error. It can be calculated by dividing D by ωC (2 pi F C). In our example, .0087/(6.28*5000*.00000047)=0.589, so ESR=0.589 ohms.

Capacitive reactance is the negative reciprocal of ωC. -1/(6.28*5000*.00000047)=-67.725, so X=-67.725 ohms.

Total impedance of a capacitor is obtained by taking the absolute value of the root sum of the squares of capacitive reactance and ESR. (67.72522+0.58922)1/2=67.727, so Z=67.727 ohms.

Capacitor current is the RMS voltage divided by the total impedance. 35/67.7=0.52 amps.

Power dissipation in the ESR component is calculated from the RMS voltage times current times the ratio of ESR to total impedance. 35*.52*(.589/67.727)=0.16 watts. Or, use I^2 times ESR. The resulting temperature rise depends on the size and heat sinking of the capacitor.

Verifying capacitor ESR on the bench requires both care and good instrumentation. Because ESR is usually small, test lead resistance and poor connections can easily contribute more resistance than the capacitor. Noise pickup from long leads or a hostile environment can make readings unstable. Note that neither D nor ESR is constant with frequency. ESR will decrease rapidly as frequency increases, then drop at a much slower rate above about 1 kHz, until it starts rising again at self resonance. Be sure to make measurements in the same frequency and voltage range that the parts will be used in- test it the way you use it. Be aware that in simulations ESR is a constant unless you take special measures to make it frequency dependent.

Dielectric Absorption is another imperfection. Briefly, the dielectric refuses to give up its full charge, and a previously discharged capacitor will self charge. This can be modeled with additional C-R pairs in parallel with the main capacitor. Dielectric absorption is a particular problem in capacitors used in integrators. There is some debate as to its importance in audio applications. Much that has been written about dielectric absorption remains obscure; at least two standard tests exist, but there is very little published data for specific parts.

Other Effects Capacitors are not 100% linear and may contribute a small amount of distortion to signals, particularly the electrolytic types. Capacitance may vary with DC voltage. Each dielectric type will have some characteristic temperature coefficient. Some capacitors are microphonic, and may also "sing" when driven at high voltages.

This relationship is useful later in determining the change of *resistance* with frequency, which Welsby gives as $1/[1-(f/fr)^2]^2$.

4.4. Calculated Inductance

In the initial design of the experiments it is useful to be able to calculate the approximate inductance, and the following has an accuracy of around 5% :

$L = \mu o\ p/(2\pi)\ [\ \mathrm{Ln}\ (2p/r) + 0.25 - \mathrm{Ln}\ (p^2/A)]\ SL$ 4.4.1

where p is the perimeter and A the area
SL is the inductance SRF factor $1/[1-(f/fr)^2]$
fr = 0.85(150/ length)

5. CALCULATED TOTAL LOSS OF RESONANT CIRCUIT

The series resistance of the resonant circuit, including the capacitor being evaluated, is given by :

$Rs = Rwall\ [l pipes / (\pi\ dw)]\ SR\ P + Rr + Rleads + Rcap$ 5.1

where dw is the diameter of the copper pipes
$Rwall = (\rho\ \pi\ \mu o\ \mu r\ f)^{0.5}$
ρ is the resistivity of the pipes (see below)
$\mu o = 0.4\pi\ \mu H/m$
μr is the relative permeability of the copper pipes (see below)
SR is the resistance SRF factor $=1/[1-(f/fr)^2]^2$
P is the proximity factor $1/ [1- (dw/D)^2]^{0.5}$ (see Moullin ref1)
D is the spacing of the pipes (centre to centre)

4.3. Inductance changes with Frequency

The inductance of the coil changes with frequency and it was therefore measured over the frequency range of interest, and also at some lower frequencies to establish the low frequency inductance, with the following results (f is in MHz and L in μH) :

f	L	Equation 4.3.1	Error
0.5	1.1947	1.1948	0.0%
5	1.1988	1.2031	0.4%
7.459	1.2096	1.2135	0.3%
10.21	1.2464	1.2303	-1.3%
14.098	1.3147	1.2645	-3.8%
24.426	1.4322	1.4320	0.0%

It is known that the change of inductance with frequency follows approximately an equation of the form (Welsby ref 5 p37):

$$L = Lo / [1-(f/fr)2] \qquad 4.3.1$$

where Lo is the low frequency inductance
f is the frequency
fr is the Self Resonant Frequency

This equation becomes less accurate as the Self Resonant Frequency is approached, but its accuracy can be improved if a value for fr is chosen which is slightly lower than the actual SRF. A good fit with the measured inductance above is given by fr = 60 MHz (about 85% of the actual SRF), and the results of the equation with this value are shown in the table above.

Lossy Capacitors
Dr. Gary L. Johnson December 10, 2001
Manhattan,Kansas

http://www.eece.ksu.edu/~gjohnson/tcchap3.pdf

1 Dielectric Loss

Capacitors are used for a wide variety of purposes and are made of many different materials in many different styles. For purposes of discussion we will consider three broad types, that is, capacitors made for ac, dc, and pulse applications. The ac case is the most general since ac capacitors will work (or at least survive) in dc and pulse applications, where the reverse may not be true.

It is important to consider the losses in ac capacitors. All dielectrics (except vacuum) have two types of losses. One is a conduction loss, representing the flow of actual charge through the dielectric. The other is a dielectric loss due to movement or rotation of the atoms or molecules in an alternating electric field. Dielectric losses in water are the reason for food and drink getting hot in a microwave oven.

One way of describing dielectric losses is to consider the permittivity as a complex number,
defined as

$$\epsilon = \epsilon' - j\epsilon'' = |\epsilon| e^{-j\delta} \qquad (1)$$

where
e’ = ac capacitivity
e” = dielectric loss factor

(Photograph 2). The parallel pipes and the connection strips will subsequently be referred-to as the coil.

Photograph 2 : The Measurement Apparatus

4.2. Inductance of Capacitor

The total inductance of the tuned circuit will include the inductance of the capacitor. The value for this is not known and is difficult to measure, however Field & Sinclair (ref 3) report that their measurements of capacitor inductance ‘compare well with that calculated from the dimensions of the leads, stator supports, and rotor shaft’. So for the capacitor measured here, its inductance is assumed to be represented by a rod of 7.75 mm diameter and 90 mm long, and a rod of these dimensions was inserted during the inductance measurements. Assuming the inductance of this rod is given by L = u ***l*** /2pi [ln (2 ***l*** /a) - 1] henrys, then this will have a free space inductance of 0.05 μH, about 3.5% of the total inductance.

require a large number of measurements. For instance, if there are 5 capacitance settings and each is measured at 5 frequencies this gives a total of 25 separate measurements, each having to be done with high precision. In addition more than one coil would be necessary to cover the frequency range and, for instance, Moullin used 5 coils. However, with the benefit of Equation 2.1 the task is considerably reduced, since it is now only necessary to carry-out sufficient measurements to determine the factors Rs, α and β. In this respect notice in Equation 2.1 that if C is small the dielectric loss will tend to dominate, and if C is large the metallic loss and contact resistance will tend to dominate. So measurements should be made at the minimum and maximum capacitance at the least, and also with other values for added confidence.

Only one coil was used for the measurements, and this gave resonance between 7.5 and 24 MHz with the variable capacitor. On reflection some measurements at a lower frequency may have improved the accuracy, particularly of the dielectric loss, but this would have required a second larger coil.

4. INDUCTANCE

4.1. The Coil

The inductance coil was made from two copper pipes, each 15.03 mm outside diameter and 0.985 meter long. These were spaced at 115 mm (centre to centre), and shorted at one end with a section of copper pipe.

The capacitor was connected across the open ends with two short flexible copper strips, the whole forming a tuned circuit

δ = dielectric loss angle

Capacitance is a complex number $C*$ in this definition, becoming the expected real number C as the losses go to zero. That is, we define

$$C^* = C - jC'' \quad (2)$$

One reason for defining a complex capacitance is that we can use the complex value in any equation derived for a real capacitance in a sinusoidal application, and get the correct phase shifts and power losses by applying the usual rules of circuit theory. This means that most of our analyses are already done, and we do not need to start over just because we now have a lossy capacitor.

Equation 1 expresses the complex permittivity in two ways, as real and imaginary or as magnitude and phase. The magnitude and phase notation is rarely used. Instead, people usually express the complex permittivity by e and tan δ,where

$$\tan\delta = \frac{\epsilon''}{\epsilon'} \quad (3)$$

where tan δ is called either the loss tangent or the dissipation factor DF.

The real part of the permittivity is defined as

$$\epsilon' = \epsilon_r \epsilon_o \quad (4)$$

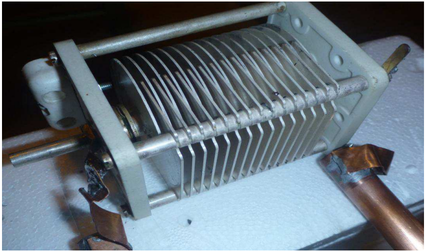
where er is the dielectric constant and eo is the permittivity of free space.

Dielectric properties of several different materials are given in Table 1 [4, 5]. Some of these materials are used for capacitors, while others may be present in oscillators or other devices where dielectric losses may affect circuit performance. The dielectric constant and the dissipation factor are given at two frequencies, 60 Hz and 1 MHz. The righthand column of Table 1 gives the approximate breakdown voltage of the material in V/mil, where 1 mil = 0.001 inch. This would be for thin layers where voids and impurities in the dielectrics are not a factor. Breakdown usually destroys a capacitor, so capacitors must be designed with a substantial safety factor.

It can be seen that most materials have dielectric constants between one and ten. One exception is barium titanate with a dielectric constant greater than 1000. It also has relatively high losses which keep it from being more widely used than it is.

We see that polyethylene, polypropylene, and polystyrene all have small dissipation factors. They also have other desirable properties and are widely used for capacitors. For high power, high voltage, and high frequency applications, such as an antenna capacitor in an AM broadcast station, the ruby mica seems to be the best.

Each of the materials in Table 1 has its own advantages and disadvantages when used in a capacitor. The ideal dielectric would have a high dielectric constant, like barium titanate, a low dissipation factor, like polystyrene, a high breakdown voltage, like mylar, a low cost, like aluminum oxide, and be easily fabricated into capacitors. It would also be perfectly stable, so the capacitance would not vary with temperature or


Photograph 1 : Variable Capacitor

3. MEASUREMENT PRINCIPLES

The measurement principle is that the reactance of the capacitor is tuned-out with a low loss coil, measurements are then made of the series resistance of this tuned circuit, and the calculated resistance of the coil subtracted to give that of the capacitor.

In practice the resistance values are too low to be measured accurately ($\approx 0.04\Omega$), and so the Q (i.e. bandwidth) of the tuned circuit is measured at resonance and compared with the calculations of the Q, assuming a given loss in the capacitor. This assumed capacitor loss is then adjusted manually in the theoretical model to make the overall calculated Q agree with the measured Q.

To fully characterise the capacitor over its full capacitance range and over its useful frequency range would normally

So the measured value of Rs lay between 4.5 mΩ and 50 mΩ. The low value in this range is consistent with the value of contact resistance measured by Bock (ref 4) who gives 1 mΩ for silver plated contacts, but the high values are too large to be due to contact resistance unless the contacts were particularly dirty or worn and this seems unlikely. It is significant then that Field & Sinclair (ref 3) say that this resistance is 'joulean loss in the metal structure'. Such metallic loss will be frequency dependent due to skin effect (i.e. proportional to $f^{-0.5}$), whereas contact resistance seems to be independent of frequency. So for capacitors with a long metallic path Equation 1.1 an additional term needs to be added:

$$R_{cap} = R_s + \alpha /(f C^2) + \beta (f^{0.5}) \qquad 2.1$$

The factors Rs, α and β are assumed to be constant for any capacitor, and it is the objective here to derive them for the capacitor shown in Photograph 1 (also the capacitor shown in Annex 1). This capacitor has square ceramic ends with sides of 65mm, and an overall length of 90mm, not including the protruding shaft. Contact to the rotating shaft is at one end only, via a spring loaded washer which is stationary and makes sliding contact with a lip on the shaft. All metalwork is silver plated. The capacitor has a range of 25 – 390 pf.

voltage. No such dielectric has been discovered so we must apply engineering judgment in each situation, and select the capacitor type that will meet all the requirements and at least cost.

Capacitors used for ac must be unpolarized so they can handle full voltage reversals. They also need to have a lower dissipation factor than capacitors used as dc filter capacitors, for example. One important application of ac capacitors is in tuning electronic equipment. These capacitors must have high stability with time and temperature, so the tuned frequency does not drift beyond some specified amount.

Another category of ac capacitor is the motor run or power factor correcting capacitor. These are used on motors and other devices operating at 60 Hz and at voltages up to 480 V or more. They are usually much larger than capacitors used for tuning electronic circuits, and are not sold by electronics supply houses. One has to ask for motor run capacitors at an electrical supply house like Graingers. These also work nicely as dc filter capacitors if voltages higher than allowed by conventional dc filter capacitors are required.

The term power factor PF may also be defined for ac capacitors. It is given by the expression

$$\mathrm{PF} = \cos\theta \qquad (5)$$

where θ is the angle between the current flowing through the capacitor and the voltage across it.
The capacitive reactance for the sinusoidal case can be defined as

$$X_C = \frac{1}{\omega C} \quad (6)$$

where ω =2 πf rad/sec, and f is in Hz.

In a lossless capacitor, e"= 0, and the current leads the voltage by exactly 90o. If e" is greater than zero, then the current has a component in phase with the voltage.

$$\cos\theta = \frac{\epsilon''}{\sqrt{(\epsilon'')^2 + (\epsilon')^2}} \quad (7)$$

For a good dielectric, e'>>e", so

$$\cos\theta \approx \frac{\epsilon''}{\epsilon'} = \tan\delta \quad (8)$$

Therefore, the term power factor is often used interchangeably with the terms loss tangent or dissipation factor, even though they are only approximately equal to each other.

We can define the apparent power flow into a parallel plate capacitor as

$$S = VI = \frac{V^2}{-jX_C} = jV^2\omega C^* = jV^2\frac{\omega A}{d}(\epsilon' - j\epsilon'') = V^2\frac{\omega A}{d}\epsilon_r\epsilon_o(j + \mathrm{DF})$$

(9)

By analogy, the apparent power flow into any arbitrary capacitor is

that proximity effect was minimal, and shorted at the far end. It therefore resembled a shorted two-wire transmission line, and the capacitor to be measured was connected to the open ends. This type of inductor is used here and is shown in Photograph 2.

A number of authors have measured the loss of variable capacitors (refs 1, 2, 3), and found that the series resistance conformed to the following equation :

Rcap = Rs + α /(f C2) ohms 1.1

where f is the frequency
C is the capacitance

The first factor Rs is a constant resistance, often accredited to the contact resistance of the slip rings, or equivalent. The second factor is due to the loss in the insulators, where again factor α is a constant. Jackson (ref 2) summarises his own measurements and those of previous workers as follows :

Rcap = 0.0045 + 0.028 [106/(f C2)] ohms (Jackson) 1.2
Rcap= 0.05 + 0.091 [105/(f C2)] ohms (Moullin) 1.3
Rcap= 0.007 + 0.012[105/(f C2)] +10/(f 2C2) ohms (Dye) 1.4
Rcap= 0.02 + 0.022 [103/(f C2)] ohms (Wilmotte) 1.5

where f is in MHZ and C in pf

Notice that Dye has an additional term related to a constant resistance across the dielectric, but this term is only significant at very low frequencies, and so is not considered here.

2. EVALUATION OF LOSS COMPONENTS

Measuring the Loss in Variable Air Capacitors
Alan Payne 2013

http://g3rbj.co.uk/wp-content/uploads/2013/10/Measurements_of_Loss_in_Variable_Capacitors_issue_2.pdf

The resistive loss of variable air capacitors is difficult to measure because they have a very high Q. The method described here uses a twin-wire transmission line made from copper pipe as the inductor to tune-out the reactance of the capacitor.

Previous authors have shown that the capacitor loss comes partly from resistive loss and partly from dielectric loss in the insulating supports. They have attributed the resistive loss to that in the rotating contacts, but it is shown that there is also a major contribution from the metallic part of the structure, although the plates themselves have negligible resistance.

1. INTRODUCTION

The loss resistance of high quality variable capacitors is difficult to measure because of the presence of the very high reactance. This must be tuned-out with an inductor, which must be of very high Q if its resistance is not to dominate the measurements, and its own resistance must be calculable to a high degree of accuracy, because this cannot be independently measured. Conventional helical coils cannot be used therefore because they do not have a sufficiently high Q, and, more importantly their loss resistance cannot be calculated with sufficient accuracy. Moullin (ref 1) solved this problem by making his inductor from two parallel conductors, spaced so

$$S = P + jQ = V^2\omega C(j + \mathrm{DF}) \qquad (10)$$

Table 1: Dielectric Constant er, Dissipation Factor DF and Breakdown Strength Vb of selected materials.

Material	ϵ_r 60 Hz	ϵ_r 10^6 Hz	DF 60 Hz	DF 10^6 Hz	V_b V/mil
Air	1.000585	1.000585	-	-	75
Aluminum oxide	-	8.80	-	0.00033	300
Barium titanate	1250	1143	0.056	0.0105	50
Carbon tetrachloride	2.17	2.17	0.007	<0.00004	-
Castor oil	3.7	3.7	-	-	300
Glass, soda-borosilicate	-	4.84	-	0.0036	-
Heavy Soderon	3.39	3.39	0.0168	0.0283	-
Lucite	3.3	3.3	-	-	500
Mica, glass bonded	-	7.39	-	0.0013	1600
Mica, glass, titanium dioxide	-	9.0	-	0.0026	-
Mica, ruby	5.4	5.4	0.005	0.0003	-
Mylar	2.5	2.5	-	-	5000
Nylon	3.88	3.33	0.014	0.026	-
Paraffin	2.25	2.25	-	-	250
Plexiglas	3.4	2.76	0.06	0.014	-
Polycarbonate	2.7	2.7	-	-	7000
Polyethylene	2.26	2.26	<0.0002	<0.0002	4500
Polypropylene	2.25	2.25	<0.0005	<0.0005	9600
Polystyrene	2.56	2.56	<0.00005	0.00007	500
Polysulfone	3.1	3.1	-	-	8000
Polytetrafluoroethylene(teflon)	2.1	2.1	<0.0005	<0.0002	1500
Polyvinyl chloride (PVC)	3.2	2.88	0.0115	0.016	
Quartz	3.78	3.78	0.0009	0.0001	500
Tantalum oxide	2.0	-	-	-	100
Transformer oil	2.2	-	-	-	250
Vaseline	2.16	2.16	0.0004	<0.0001	-

The power dissipated in the capacitor is

$$P = V^2\omega C'' = V^2\omega C(\mathrm{DF}) \qquad (11)$$

Example

Find the real and reactive power into a ruby mica capacitor with area A = 0.03m2, and a dielectric thickness d = 0.001 m, if the voltage is 2000 V (rms) at a frequency f = 1 MHz.

$$S = V^2 \frac{\omega A}{d} \epsilon_r \epsilon_o (j + \mathrm{DF})$$

$$= (2000)^2 \frac{2\pi(10^6)(0.03)}{0.001}(5.4)(8.854 \times 10^{-12})(j + 0.0003)$$

$$= j36040 + 10.8$$

The capacitor is absorbing 36040 capacitive VARs (Volt Amperes Reactive) and 10.8 Watts. The real power of 10.8 W appears as heat and must be removed by appropriate heat sinks.

The real power dissipation in a capacitor varies directly with frequency if the dissipation factor remains constant, and also with the square of the voltage. At low frequencies, the voltage limit is determined by the dielectric strength. At high frequencies, however, the voltage limit may be determined by the ability of the capacitor to dissipate heat. If the ruby mica capacitor in the previous example could safely dissipate only 10 W, but was to be used at 5 MHz, the operating voltage must be reduced from 2000 V to keep the losses in an acceptable range.

The dissipation factor varies significantly with frequency for some materials in Table 1. The actual variation must be determined experimentally. If interpolation or extrapolation seems necessary to find the loss at some frequency not given in

Test System Safety

Many electrical test systems or instruments are capable of measuring or sourcing hazardous voltage and power levels. It is also possible, under single fault conditions (e.g., a programming error or an instrument failure), to output hazardous levels, even when the system indicates no hazard is present. These high-voltage and power levels make it essential to protect operators from any of these hazards at all times. It is the responsibility of the test system designers, integrators and installers to make sure operator and maintenance personnel protection is in place and effective. Protection methods include:

* Design test fixtures to prevent operator contact with any hazardous circuit.
* Make sure the device under test is fully enclosed to protect the operator from any flying debris.
* Double insulate all electrical connections that an operator could touch. Double insulation ensures the operator is still protected, even if one insulation layer fails.
* Use high-reliability, fail-safe interlock switches to disconnect power sources when a test fixture cover is opened.
* Where possible, use automated handlers so operators do not require access to the inside of the test fixture.
* Provide proper training to all users of the system so they understand all potential hazards and know how to protect themselves from injury.

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suitable soak period. After the capacitors have been tested, the voltage source should be set to zero, and then some time allowed so the capacitors can discharge before they are removed from the test fixture.

Note that in Figure 4 the capacitors have a discharge path through the normally closed contacts of the relays. To prevent electric shock, test connections must be configured in such a way that the user cannot come in contact with the conductors, connections or DUT. Safe installation requires proper shielding, barriers and grounding to prevent contact with conductors.

More complex test systems that combine leakage measurement with capacitance measurements, dielectric absorption and other tests, if desired, are possible. A simplified schematic of such a test system using an LCZ bridge and a picoammeter with a voltage source is shown in Figure 5.

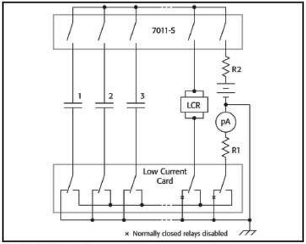


FIGURE 5. Capacitance and IR measurement system.

the Table, the best assumption would be that DF varies linearly with log f. That is, if DF = 0.02 at f = 102 and 0.01 at f = 106, a reasonable assumption at f = 104 would be that DF = 0.015.

The basic circuit model for a capacitor is shown in Fig. 1. Any conductor, whether straight or wound in a coil, has inductance, and the capacitor inductance is represented by a series inductance Ls. The effects of conductor resistance and dielectric losses are represented by a series resistance Rs. Leakage current through the capacitor at dc flows through a parallel resistance Rp. In some manufacturer's databooks, these are called ESL, ESR, and EPR, where
Ls = ESL = equivalent series inductance
Rs = ESR = equivalent series resistance
Rp = EPR = equivalent parallel resistance

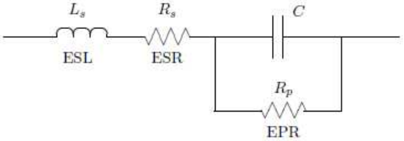


Figure 1: Equivalent Circuit for a Capacitor

This circuit indicates that every capacitor has a self-resonant frequency, above which it becomes an inductor. This is certain to puzzle a student making measurements on a capacitor above this frequency if the student is not aware of this fact. Rs is readily measured by applying this frequency to a capacitor, measuring the voltage and current, and calculating the ratio. The capacitive and inductive reactances cancel at the resonant frequency, leaving only Rs to limit the current. The resistance

Rp will always be much larger than the capacitive reactance at the resonant frequency, so this resistance can be neglected for this computation.

The self-resonant frequency of a high capacitance unit is lower than that for a low capacitance unit. Hence, in some circuits we will see two capacitors in parallel, say a 10μF in parallel with a 0.001μF capacitor as shown in Fig. 2. At first glance, this seems totally unnecessary. However, the larger capacitor is used to filter low frequencies, say in the audio range, while the small capacitor filters the high frequencies which are above the self-resonant frequency of the large capacitor.

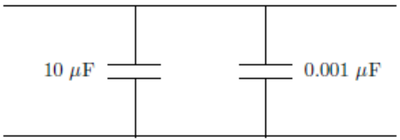


Figure 2: Capacitors in Parallel to Filter Two Different Frequencies

An example of the variation of Rs and self-resonant frequency fres with capacitance value and rated voltage is given in Table 2. These are metallized polypropylene capacitors designed for switch-mode power supplies by the company Electronic Concepts, Inc. This application requires a low Rs and a high fres so capacitors designed for other applications will tend to have higher values of Rs and lower values of fres.

that can integrate easily with a switching system capable of higher channel counts.

Configuration examples.

Producing enough useful data for statistical analysis requires testing a large quantity of capacitors quickly. Obviously, performing these tests manually is impractical, so an automated test system is required. Figure 4 illustrates such a system, which employs an electrometer with built-in voltage source, as well as a switching mainframe that houses a low current scanner card and a Form C switching card. In this test setup, a single instrument provides both the voltage sourcing and low current measurement functions. A computer controls the instruments to perform the tests automatically.

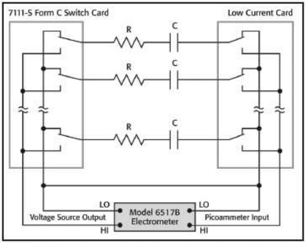


FIGURE 4. Example capacitor leakage test system configuration.

One set of switches is used to apply the test voltage to each capacitor in turn; a second set of switches connects each capacitor to the electrometer's picoammeter input after a

Test Hardware Considerations

A variety of considerations go into the selection of the instrumentation used when measuring capacitor leakage:

* Although it is certainly possible to set up a system with a separate voltage source, an integrated one simplifies the configuration and programming process significantly, so look for an electrometer or picoammeter with a built-in variable voltage source. A continuously variable voltage source calculates voltage coefficients easily. For making high-resistance measurements on capacitors with high voltage ratings, consider a 1000 V source with built-in current limiting. For a given capacitor, a larger applied voltage within the voltage rating of the capacitor will produce a larger leakage current. Measuring a larger current with the same intrinsic noise floor produces a greater signal-to-noise ratio and, therefore, a more accurate reading.
* Temperature and humidity can have a significant effect on high-resistance measurements, so monitoring, regulating and recording these conditions can be critical to ensuring measurement accuracy. Some electrometers monitor temperature and humidity simultaneously. This provides a record of conditions, and permits easier determination of temperature coefficients. Automatic time-stamping of readings provides a further record for time-resolved measurements.
* Incorporating switching hardware into the test setup allows automating the testing process. For small batch testing in a lab with a benchtop test setup, consider an electrometer that offers the convenience of a plug-in switching card. For testing larger batches of capacitors, look for an instrument

The operating frequency range will obviously be less than the self-resonant frequency. This could easily be as low as a few kHz for large motor run capacitors.

Table 2: Capacitor Resistance and Self-resonant Frequency for Electronic Concepts Type 5MP Metallized Polypropylene capacitors.

VDC volts	C μF	R_s Ω	f_{res} kHz
100	1	0.015	1065
100	2	0.012	703
100	5	0.010	385
100	10	0.009	248
200	1	0.020	861
200	2	0.015	609
200	5	0.011	323
200	10	0.009	200
400	1	0.019	784
400	2	0.015	511
400	5	0.010	283
400	10	0.006	200

Capacitors will always be rated for a working voltage as well as a specific value of capacitance. This voltage will always be well under the breakdown voltage of the dielectric. It may be specified either as a dc voltage or an ac voltage, depending on the application. Motor run capacitors are always operated on ac, so the voltage is specified as, say, 370 or 480 VAC. They can also be operated on dc, with a dc rating of at least $\sqrt{2}$ times the ac rating. Electrolytic capacitors that can only be operated on dc will have their working voltage expressed as WVDC. This is the maximum dc voltage, plus the peak of the ac ripple voltage, that should be continuously applied to a capacitor to prevent excessive deterioration and aging. Capacitors that are sometimes used on ac, and sometimes on dc, usually have working voltages expressed as WVDC, as was the case for the polypropylene capacitors in Table 2.

We will now present some more detailed background information on several of the dielectrics shown in Table 1.

Mica is a natural material that can be easily split into thin layers. It is very stable and does not deteriorate with age. The maximum capacitance is on the order of 0.03 μF. Mica capacitors tend to be quite expensive and the larger sizes are used only in critical applications like radio transmitters.

Glass capacitors were first developed during World War II as a replacement for mica capacitors when supplies of mica were threatened. Glass capacitors exhibit excellent long-term parametric stability, low losses, and can be used at high frequencies. They are used in aerospace applications where capacitance must not vary. The maximum size is limited to about 0.01μF

Paper capacitors use kraft paper impregnated with non-ionized liquid electrolyte as the dielectric between aluminum foils. These capacitors can provide large capacitances and voltage ratings but tend to be physically large. The thickness of the paper layers and the particular electrolyte used can be varied to produce a wide range of electrical characteristics, which is the reason for not listing paper in the dielectrics of Table 1.

Ceramic capacitors are made with one of a large number of ceramic materials, which include aluminum oxide, barium titanate, and porcelain. These are very widely used as bypass capacitors in electronic circuits. The older style is a single

The series resistor also adds Johnson noise – the thermal noise created by any resistor – to the measurement. At room temperature, this is roughly 6.5×1010 amps, p-p. The current noise in a 1 TΩ feedback resistor at a typical 3 Hz bandwidth would be ~8×10-16 A. When measuring an insulation resistance of 1016Ω at 10 V, the noise current will be 80% of the measured current.

Alternate test circuit.

Greater measurement accuracy can be achieved by including a forward-biased diode (D) in the circuit (Figure 3). The diode acts like a variable resistance, low when the charging current to the capacitor is high, then increases in value as the current decreases with time. This allows the series resistor used to be much smaller because it is only needed to prevent overload of the voltage source and damage to the diode if the capacitor is short-circuited. The diode used should be a small signal diode, such as a 1N914 or a 1N3595, but it must be housed in a light-tight enclosure to eliminate photoelectric and electrostatic interference. For dual-polarity tests, two diodes should be used back to back in parallel.

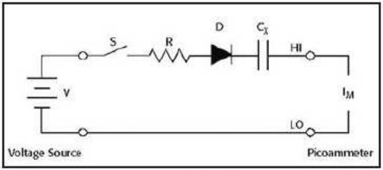


FIGURE 3. Alternative test circuit that incorporates a small-signal diode.

1MΩ·µF to 100 MΩ· µF. For example, a 4.7 µF aluminum cap specified as 50 MΩ· µF is guaranteed to have at least 10.6 MΩ insulation resistance.

Capacitor leakage test method.

Figure 2 illustrates a general circuit for testing capacitor leakage. Here, the voltage is placed across the capacitor (CX) for the soak period; then the ammeter measures the current after this period has elapsed. The resistor (R), which is in series with the capacitor, serves two important functions. First, it limits the current, in case the capacitor becomes shorted. Second, the decreasing reactance of the capacitor with increasing frequency will increase the gain of the feedback ammeter. The resistor limits this increase in gain to a finite value. A reasonable value is one that results in an RC product from 0.5 to 2 sec. The switch (S), while not strictly necessary, is included in the circuit to allow control over the voltage to be applied to the capacitor.

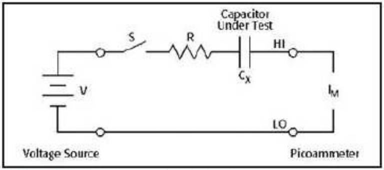


FIGURE 2. A simple capacitor leakage test circuit.

layer of dielectric separating two conducting plates and packaged in a small disc. The newer style is the monolithic, which appears in a rectangular package. It consists of alternating layers of ceramic material and printed electrodes which are sintered together to form the final package. The self resonant frequency is on the order of 15 MHz for the 0.01µF size, for a total lead length of 0.5 inch, and on the order of 165 MHz for the 0.0001µF = 100 pF size. Ceramic capacitors are classified into Class 1 (r<600) and Class 2 (r>600) by the Electronics Industries Association (EIA). Class 1 ceramic capacitors tend to be larger than their Class 2 counterparts, and have better stability of values with changes in temperature, voltage, or frequency. The best Class 1 capacitor, with nearly constant characteristic, will be labeled 'C0G' using EIA designators, but is often referred to as 'NP0' which stands for 'negative-positive-zero' (temperature coefficient).

Plastic-film capacitors use extremely thin sheets of plastic film as the dielectric between capacitor plates, usually in a coil construction. They have low losses and good resistance to humidity. Four common materials are:
1. Polycarbonate
2. Polypropylene
3. Polystyrene
4. Teflon

Generally speaking, as one moves down in this list from polycarbonate to Teflon, the capacitors get larger, better, and more expensive.

Historically, low-budget Tesla coilers have used either saltwater capacitors (as Tesla himself did) or homemade rolled polyethylene and foil capacitors. The saltwater capacitors are lossy and heavy. They are one of the first components to be

replaced as a new coiler grows with his hobby. The homemade foil capacitors are limited by corona onset to about 7000 V. They are oil filled, so if the container leaks or is tipped over, one has a real mess. Serious coilers would look for commercial high voltage, high current, pulse rated capacitors made by specialty companies like Maxwell. These worked fine, but were expensive if purchased new, and difficult
to find on the used market.

This all changed in 1999 when the Tesla coil community moved en masse to multi mini capacitors (MMC). These are small commercial capacitors that one buys by the sackful from Digi-Key and connects in series and parallel strings to get the required ratings.

There are several different capacitor series, some of which are more suitable for Tesla coil work than others. One should look for capacitor types that are rated for "high voltage, high frequency, and high pulses". A dissipation factor of 0.1% at 1 kHz is good, as is polypropylene for the dielectric. The ECWH(V) and ECQP(U) series of capacitors are good. The Panasonic ECQ-E capacitors utilize metallized polyester (Mylar) which has a higher dissipation factor (1% at 1 kHz) than the polypropylene capacitors.

Capacitor ratings have received a great deal of attention by Tesla coilers. A given capacitor might be rated at 1600 VDC and 500 VAC. The AC rating is always for rms values, so the peak voltage would be about 700 V for a rating of 500 VAC. A capacitor in a Tesla coil primary experiences full voltage reversal, so it would seem wise to find the AC rating, multiply by $\sqrt{2}$, and divide that into the peak voltage available from the iron core transformer to find the number of series capacitors to use.

source and the capacitance. This is especially true for capacitance values >1 nF. A typical series resistor would be 10 kΩ to 1 MΩ.

Capacitor Leakage

Leakage is one of the less-than-ideal properties of a capacitor, expressed in terms of its insulation resistance (IR). For a given dielectric material, the effective parallel resistance is inversely proportional to the capacitance. This is because the resistance is proportional to the thickness of the dielectric, and inverse to the capacitive area. The capacitance is proportional to the area and inverse to the separation. Therefore, a common unit for quantifying capacitor leakage is the product of its capacitance and its leakage resistance, usually expressed in megohms-microfarads (MΩ·µF). Capacitor leakage is measured by applying a fixed voltage to the capacitor under test and measuring the resulting current. The leakage current will decay exponentially with time, so it's usually necessary to apply the voltage for a known period (the soak time) before measuring the current.

In theory, a capacitor's dielectric could be made of any non-conductive substance. However, for practical applications, specific materials are used that best suit the capacitor's function. The insulation resistance of polymer dielectrics, such as polystyrene, polycarbonate or Teflon, can range from 104 MΩ·µF to 108 MΩ·µF, depending on the specific materials used and their purity. For example, a 1000 pF Teflon cap with an insulation resistance higher than 1017 Ω is specified as >108 MΩ·µF. The insulation resistance of ceramics such as X7R or NPO can be anywhere from 103 MΩ·µF to 106 MΩ·µF. Electrolytic capacitors such as tantalum or aluminum have much lower leakage resistances, typically ranging from

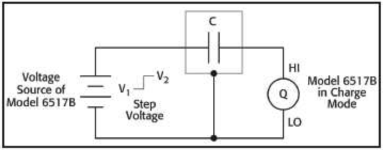


FIGURE 1. Capacitance measurement using an electrometer with an integrated voltage source.

Then, the voltage source should be turned on and the charge reading noted immediately. The capacitance can then be calculated from:

$$C = \frac{Q_2 - Q_1}{V_2 - V_1}$$

where

Q2 = final charge
Q1 = initial charge (assumed to be zero)
V2 = step voltage
V1 = initial voltage (assumed to be zero)

After the reading has been recorded, reset the voltage source to 0 V to dissipate the charge from the device. Before handling the device, verify the capacitance has been discharged to a safe level. The unknown capacitance should be in a shielded test fixture. The shield is connected to the LO input terminal of the electrometer. The HI input terminal should be connected to the highest impedance terminal of the unknown capacitance. If the rate of charge is too great, the resulting measurement will be in error because the input stage becomes temporarily saturated. To limit the rate of charge transfer at the input of the electrometer, add a resistor in series between the voltage

Experience of coilers, however, has been that there are sufficient safety factors built into the capacitors that using the DC rating is quite acceptable. For example, a 15 kV transformer has a peak voltage of 15√2 = 21. 2 kV. Divide 21.2 by 1.6 for 1600 VDC capacitors to get 13.25. Use either 13 or 14 capacitors in series.

The voltage rating is not determined by the dielectric breakdown voltage as much as it is by the Ionization or Corona Inception Level . This refers to the level at which ionization or partial discharges can begin to occur inside a bubble of entrapped air or within an air-filled void within the solid dielectric system. If one had perfect
dielectrics and could always exclude any entrapped air, derating for this phenomenon would not be necessary. Corona inside a capacitor will chemically degrade the dielectric material over a period of time until the capacitor ultimately fails. Manufacturers of capacitors might select a voltage rating whereby their capacitors will last for one million hours before this occurs. In Tesla coil use, 100 hours of actual operation is a long time. This helps explain why coilers can exceed the manufacturer ratings without immediate problems.

There is evidence that operating a capacitor above its ac voltage rating will shorten its life by a factor of the overvoltage ratio raised to the 15th power. Suppose we have a capacitor rated at one million hours at 500 VAC. For economic reasons we are thinking about operating it at 700 VAC. The life reduction factor is (700/500)15 = 155. 57. Dividing one million by 155.57 gives an expected life of 6428 hours or almost one year of continuous operation. This is more than adequate for most Tesla coil applications.

One brand of capacitor is the WIMA. Terry Fritz comments about them: “The switching power supplies we build have WIMAs in pulsed duty similar to that seen in Tesla coil use.

The real indicator of how long they live is how warm they get. If they heat to about 5 degrees C above the ambient, then they last forever. At about 8 degrees above ambient, we see occasional failures. At 10 degrees, they get to be a problem. In many situations, we go beyond the WIMA chart derating in pulsed applications with no problem at all, just being sure they don’t get hot.”

WIMA experts say that partial discharges occur only above a certain voltage level and only on ac. Frequency does not seem to be a factor. They agree that the ac rating can be exceeded in Tesla coil applications, but without a serious investigation, were mentioning factors like 1.25 to 1.5.

Some capacitors that are candidates for a MMC use actual metal foil for electrodes while others use metallized plastic. The metal foil devices are physically larger for a given rating and are much more robust in Tesla coil service.

Dielectric material, electrode thickness, and lead size all limit the rate at which charge can be added to or removed from a capacitor. This limit is expressed as the dV/dt limit. The peak current I peak can be determined from the dV/dt value by the equation

$$I_{peak} = \frac{dV}{dt}C \quad (12)$$

The coulomb’s function of an electrometer can be used with a step voltage source to measure capacitance levels ranging from <10pF to hundreds of nanofarads.

The unknown capacitance is connected in series with the electrometer input and the step voltage source.

The calculation of the capacitance is based on this equation:

$$C = \frac{Q}{V}$$

Figure 1 illustrates a basic configuration for measuring capacitance with an electrometer with an internal voltage source. The instrument is used in the charge (or coulombs) mode, and its voltage source provides the step voltage. Just before the voltage source is turned on, the meter’s zero check function should be disabled and the charge reading suppressed by using the REL function to zero the display. (The purpose of zero check is to protect the input FET from overload and to zero the instrument. When zero check is enabled, the input of the electrometer is a resistance from roughly 10 MΩ to 100 MΩ, depending on the electrometer used. Zero check should be enabled when changing conditions on the input circuit, such as changing functions and connections. The REL function subtracts a reference value from actual readings. When REL is enabled, the instrument uses the present reading as a relative value. Subsequent readings will be the difference between the actual input value and the relative value.)

Capacitor Testing Challenges and Solutions
by Dale Cigoy
http://www.circuitsassembly.com/cms/component/content/article/159/10164-capacitor-testing-challenges-and-solutions

Chip caps are prone to leakage, so consider these test methods for minimizing electrical failures.

Capacitors are widely used for bypassing, coupling, filtering and tunneling electronic circuits. However, to be useful, their capacitance value, voltage rating, temperature coefficient and leakage resistance must be characterized. Although capacitor manufacturers perform these tests, many electronics assemblers also perform some of these tests as quality checks. This article looks at some of the challenges associated with capacitor testing, as well as some of the test techniques used.

A capacitor is somewhat like a battery in that they both store electrical energy. Inside a battery, chemical reactions produce electrons on one terminal and store electrons on the other. However, a capacitor is much simpler than a battery because it can't produce new electrons – it only stores them. Inside the capacitor, the terminals connect to two metal plates separated by a non-conducting substance known as a dielectric.

A capacitor's storage potential, or capacitance, is measured in farads. A one-farad (1 F) capacitor can store one coulomb (1 C) of charge at one volt (1 V). A coulomb is 6.25´1018 electronics. One amp represents a rate of electron flow of 1C of electrons per second, so a 1 F capacitor can hold one amp-second (1 A/s) of electrons at 1 V.

where appropriate multipliers should be used to get everything in volts, seconds, and farads. By way of reference, a 1000 pF capacitor rated for 10,000 V/µs will withstand 10 A surges. A capacitor with a dV/dt rating less than 1000 V per 5µs probably does not have metal foil endplates and should not be used.

The peak current that a Tesla coil primary capacitor must provide is determined by the capacitor voltage and the surge impedance. If the secondary were removed, the surge impedance would be

$$Z_s = \sqrt{\frac{L_1}{C_1}} \quad (13)$$

where C1 is the capacitance in the Tesla coil primary and L1 is the inductance of the primary. When the secondary is in place, the surge impedance will increase from the above value, so the peak current required from the capacitor will be less. Using the above expression for Zs should be a worst case calculation. If it tells you that there might be 100 A flowing after the gap shorts, and your individual capacitors are rated at 10 A each, then you would need ten parallel strings. Since this figure includes some factor of safety, you might be able to get by with somewhat fewer strings. One WIMA expert said that the maximum amperage rating is not as critical as the Corona Inception Level mentioned above.

The polypropylene capacitors useful for Tesla coil work have a self-healing mode. A partial discharge will eventually cause a local failure. A burst of energy through this short will vaporize everything around the short, effectively removing the short.

The capacitor continues to operate, but with a slightly lower capacitance due to some of the electrodes being removed. It is a good idea to carefully measure the capacitance of the MMC periodically. A decrease of even 1% could indicate that the capacitors are being stressed, and that total failure is possible.

Terry Fritz reports testing some WIMA MKS 4 (K4) capacitors rated at 1μF and 400VDC on a large DC supply. These are metallized polypropylene dielectric capacitors that are encapsulated in epoxy filled 3×1 cm rectangles. Starting at about 900 V there are a few little snaps, indicating that the dielectric had punched through and the arc blasted the thin metal layer on the other side. Going to 1200 V gave a few more snaps. Around 1500 V there were many snaps and the capacitor shorted out.

When a capacitor cleanly blows apart, that is a sign that they were destroyed by over voltage. When a capacitor puffs up and looks melted and burned, that is from too much current. It is always a good idea to think about possible failure modes, and put the capacitors in appropriate enclosures to protect bystanders.

I had a student build a small design project once, that illustrates this point. He used an electrolytic capacitor rated at about 25 VDC in some circuit for 120 VAC usage, and I did not notice it before testing. He was proudly demonstrating performance, using a variac to bring up the voltage. At about 90 V, the electrolytic capacitor exploded with a noise somewhere between a pistol and a small shotgun. The contents of the capacitor, about a cubic centimeter of plastic fibers, hit the student in the middle of the chest. He was not injured, but for a brief moment he thought he had been killed by this exploding capacitor. I suspect he still pays

Value	Working Voltage	ESR
1µF	50V	2Ω
1µF	63V	1.3Ω
4.7µF	62V	2Ω
22µF	16V	1Ω
22µF	63V	0.8Ω
100µF	16V	0.6Ω
100µF	35V	0.5Ω
100µF	50V	0.2Ω
330µF	50V	0.3Ω
1000µF	16V	0.1Ω

Conclusion

All the capacitors had an ESR of 2 ohms or less and most of the values that you would use in a power supply (22µF and up) had an ESR of less than 1 ohm. As a rule of thumb, higher valued capacitors with higher working voltages had a lower ESR, although there were some exceptions so this cannot be guaranteed.

Keep in mind that the ESR for standard electrolytic capacitors increases with lower temperatures. For example, it will typically increase by 20 to 30 times when the temperature drops from 25°C to -40°C.

more attention to capacitor voltage ratings than most people.

Many coilers recommend placing resistors across the MMC to bleed off the charge after power is removed. This can eliminate some very unpleasant surprises when making adjustments between runs. One can get true high voltage resistors but they are expensive. Most use a long chain of ordinary resistors therefore. Digi-Key sells 0.5 W carbon film resistors rated at 350 volts each for about two cents each in quantity. At this price, there is little point in not using adequate safety factors. Suppose we feel comfortable in operating a 0.5 W resistor at 0.1 W and at the rated peak voltage of 350 V each. We solve for the resistance as

$$R = \frac{V^2}{P} = \frac{(350/\sqrt{2})^2}{0.1} = 612 \ \text{k}\Omega$$

where we round off to the nearest standard value of 620 kΩ.

The number of resistors needed for a string across say a 15 kV transformer secondary would be

$$N = \frac{15000\sqrt{2}}{350} = 60 \ \text{resistors}$$

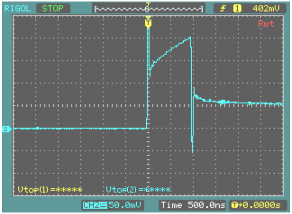
We then check the time constant to get a measure of how long it takes to discharge a

Table 3: Dielectric Constant r and Dissipation Factor DF of Water.

		Frequency in Hz			
Temperature		10^5	10^6	10^7	10^{10}
1.5°C	ϵ_r'	87.0	87.0	87	38
	DF	0.1897	0.01897	0.00195	1.026
25^o	ϵ_r'	78.2	78.2	78.2	55
	DF	0.3964	0.03964	0.00460	0.545
85°C	ϵ_r'	58	58	58	54
	DF	1.2413	0.12413	0.01259	0.259

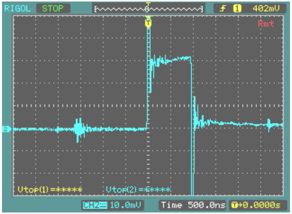
capacitor chain. If the capacitance happened to be 27 nF, the time constant would be
$\tau = RC = (60)(620000)(27\times10-9) = 1$ second

A capacitor will be mostly discharged after five time constants or 5 seconds in this example. There are obviously many other design possibilities. If we assumed the resistors would accept a higher voltage before arcing over, then we would want to use higher ohm values to keep heating within bounds. Terry Fritz tested some Yageo 10 MΩ 0.5 W carbon film resistors to see where they would actually fail. At 4000 V the resistor started to turn brown and smoke a little. This should be expected since it is now drawing 2 W, four times its power rating. At 5300 V, it gave a very satisfying crack and arced along its outside surface. After that, the 2 MΩ resistor measured 11.1 MΩ. So a resistor specified at 350 V actually failed at 5300 V, a safety factor of 15. Terry sees no problem with using these 0.5 W resistors up to 1000 V if the resulting wattage is not excessive.



Results

The following table lists the results for a representative collection of aluminium electrolytic capacitors in my parts collection. The manufacturers were Elete and Lelon and they were bought from Futurlec and Jaycar (in Australia). They represent the typical components that a hobbyist would have access to.



This method will work for values of ESR from 0.1Ω to greater than 10Ω. Alas, it is not very accurate - but then the ESR of a capacitor is not a precise value anyway and I was only looking for an approximation.

The main problem with this technique is that it is not very effective with capacitors below 10µF. With these small values the charge on the capacitor will rapidly rise making it difficult to estimate the height of the pulse before the effect of charging takes over – although, with some guesswork it can be still used for values down to 1 µF.

This is illustrated in the screenshot on the left. The capacitor on test was a 2.2µF 63VW electrolytic and it is possible to estimate that the start of the pulse is about 130mV (vertical scale is 50mV/division). This gives an ESR of 1.3Ω.

Equivalent Series Resistance (ESR) of Capacitors
QuadTech, Incorporated
http://www.lowesr.com/QT_LowESR.pdf

Questions continually arise concerning the correct definition of the ESR (Equivalent Series Resistance) of a capacitor and, more particularly, the difference between ESR and the actual physical series resistance (which we'll call Ras), the ohmic resistance of the leads and plates or foils. The definition and application of the term ESR has often been misconstrued. We hope this note will answer any questions and clarify any confusion that might exist. Very briefly, ESR is a measure of the total lossiness of a capacitor. It is larger than Ras because the actual series resistance is only one source of the total loss (usually a small part).

The Series Equivalent Circuit
At one frequency, a measurement of complex impedance gives two numbers, the real part and the imaginary part: $Z = Rs + jXs$. At that frequency, the impedance behaves like a series combination of an ideal resistance Rs and an ideal reactance Xs (Figure 1).

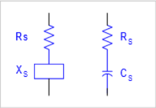
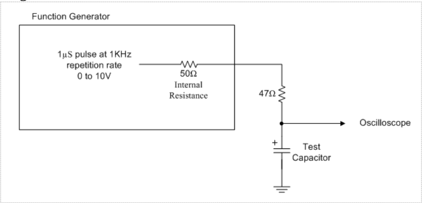


Figure 1: Equivalent Series Circuit Representation

If Xs is negative, the impedance is capacitive and the general reactance can be replaced with a capacitance of:

Test Setup

The method that I used is quite simple and is shown in the diagram below.



I used a function generator to generate narrow pulses (1µS wide) with a slow repetition rate (1KHz). This drove a voltage divider with the test capacitor in the bottom leg and 100Ω in the top leg. Because of the narrow pulse, the capacitor did not have enough time to build up a charge, so the voltage across it represented the voltage drop caused by its ESR.

With the voltage of each pulse set at 10V this arrangement meant that the height of the pulse (in mV) across the test capacitor divided by 100 was equivalent to the capacitor's ESR in ohms. For example, a pulse height of 80mV represented an ESR of 0.8Ω.

This is illustrated on the right. The vertical sensitivity of the oscilloscope was 10mV per division and the height of the pulse was about 30mV. This means that the capacitor's ESR was approximately 0.3Ω (the capacitor under test was a 330µF 50VW aluminium electrolytic).

Measuring Capacitor ESR

by Geoff
http://geoffg.net/Measuring_ESR.html

Recently, while selecting components for a power supply, I came across the following statement in the datasheet for the voltage regulator... "The output capacitor must have an ESR of less than 3 ohms."

When I looked up the retailers who supplied the capacitors in my parts bin I found that none of them listed the ESR for the capacitors that they sold. So, somehow, I needed to measure this value.

After some research I came up with the simple approach described below. It relies on some good test equipment and, while you may not want to repeat the test, the results should be informative.

ESR

Equivalent Series Resistance (ESR) is the internal resistance that appears in series with the device's capacitance. No capacitor is perfect and the ESR comes from the resistance of the leads, the aluminium foil and the electrolyte. It is often an important parameter in power supply design where the ESR of an output capacitor can affect the stability of the regulator (ie, causing it to oscillate or over react to transients in the load).

You can buy ESR meters but I just wanted to know if any of the capacitors in my parts bin had an ESR of less than 3 ohms. So, how to test them?

$$Cs = \frac{-1}{\omega Xs}$$

We now have an equivalent circuit that is correct only at the measurement frequency. The resistance of this equivalent circuit is the equivalent series resistance ESR = Rs = Real part of Z

Add the Dissipation Factor
If we define the dissipation factor D as

$$D = \frac{\text{energy lost}}{\text{energy stored}} = \frac{\text{Real part of Z}}{(-\text{Imaginary part of Z})}$$

$$\text{Then } D = \frac{Rs}{(-)Xs} = Rs\omega C = (ESR)\omega C$$

If one took a pure resistance and a pure capacitance and connected them in series, then one could say that the ESR of the combination was indeed equal to the actual series resistance. However, if one put a pure resistance in parallel with a pure capacitance (Figure 2a), the ESR of the combination is

Real part of Z = Real part of

$$\frac{1}{\frac{1}{Rp} + j\omega Cp} = \frac{Rp}{1 + \omega^2 Cp^2 Rp^2}$$

as illustrated in Figure 2b. From Figure 2a, however, it is obvious that there is no actual series resistance in series with the capacitor.

Ras = 0

But ESR > 0
Therefore ESR > Ras

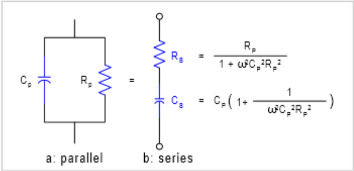


Figure 2: Series and Parallel Equivalent Circuits

QuadTech manufactures a wide variety of LCR meters designed for accurate measurements of C, Df and ESR. The design of the QuadTech 1930 LCR Meter, has been specifically optimized for high speed, low ESR measurements on tantalum and other types of capacitors. Fast settling time and quick open/short recovery significantly reduce test times especially on capacitors greater than 100µ F at 120Hz. The 1930 Low ESR Capacitance Meter performs most impedance measurements (C, Df, ESR, Z, R, X, L Q, Y, G, B, θ and DCR) over a frequency range of 100Hz – 100kHz. Actual voltage and current is measurable across the DUT.

The ESR of a Real Capacitor

Actual capacitors have three main sources of loss:
1. Actual series resistance: There is some resistance in the leads and plates or foils. This is the resistance of conductors

higher inductance in this test setup. I believe this was due to the fact that the smaller capacitor did not fill the package and internal lead inductance caused the effect. In this case, a 1 uF capacitor was a better choice than a 0.1 uF!

One of the advantages of this test is that the output waveform is the transient response of the capacitor. The voltages developed across the capacitor in this test are directly related to what will happen in a real circuit if the current risetime from the generator is similar to what the capacitor will see in its intended application.

was repetitive and the slight slope on the left half of the waveform was the end of the exponential fall from 5 volts. If a single pulse on a digital scope was used, the slope to the left of the Ldi/dt spike would be zero.

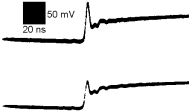


Figure 5. 1uF Capacitor

Figure 6 shows the result for a 1 uF radial ceramic capacitor (square case). Note the low inductance and undetectable ESR. Note also that the slope of the 1/C exponential rise is flatter indicating more capacitance than the 1 uF capacitor of Figure 5. This may be due to the fact that the electrolytic capacitor used for Figure 5 may have lower capacitance near zero voltage than at its operating voltage whereas the ceramic capacitor has a more constant capacitance with voltage. The inductance corresponding to the lower trace is estimated to be 4.4 nH.

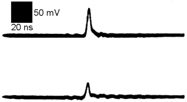


Figure 6. 1uF Ceramic Capacitor

It is interesting to note that a 0.1 uF ceramic capacitor in the same size package as the 1 uF of Figure 6 showed a slightly

and is always low. It causes a power loss I2Ras where I is the current flowing in the capacitor. This causes $D_1 = \omega R_{as} C$

2. Leakage resistance: There is some actual parallel resistance due to leakage current in the capacitor. We'll call this RL. It is the resistance of the capacitor at DC and it is a high resistance. For plastic capacitors it can be 10^12ohms (GΩ) or higher. It causes a loss of E^2/R_L where E is the applied (rms) voltage and $D_2 = \frac{1}{\omega R_L C}$. This loss is usually negligible except at very low frequencies.

3. Dielectric loss: This is due to two phenomena molecular polarization and interfacial polarization (dielectric absorption). It is not the purpose of this note to discuss its cause, but only its effect. It causes a D that can change with frequency in most any manner that is not abrupt. It acts neither as a series resistance, nor as a parallel resistance that would pass DC. One model is a parallel resistance, RD, which is variable with frequency, with DC blocked by a large series capacitance (CB) as illustrated in Figure 3.

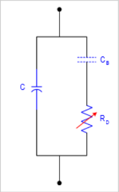


Figure 3: Parallel Resistance

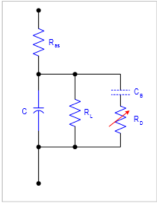


Figure 4: All 3 Capacitor Losses

If the series C were infinite, the D would be $D_3 = \dfrac{1}{\omega R_D C}$ but it must be remembered that it is not inversely proportional to ω because RD changes with frequency.

All three sources of loss are illustrated in Figure 4. This circuit would have a total dissipation factor equal to D1+ D2+ D3.

Plot of Total Dissipation

For the circuit in Figure 4, the total dissipation factor is

$$D = D_1 + D_2 + D_3 = \omega RasC + \frac{1}{\omega R_L C} + \frac{1}{\omega R_D C}$$

Figure 5 illustrates a general plot of this combination series and parallel resistance in a capacitance measurement.

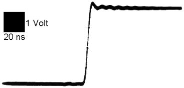


Figure 3. Input From Pulse Generator

Figure 4 shows data from a 4 uF electrolytic capacitor. The ESR offset is about 50 mV yielding an estimate of the ESR of just over one Ohm. Notice that there appears to be some oscillations on the 1/C part of the slope. This could be scope probe resonance or a resonance in the capacitor. The data was taken with a standard 10X Hi-Z probe, so the probe is suspect. I have seen capacitors with pronounced oscillation from internal resonance. If you are planning to put a large capacitor in parallel with a smaller one, especially if they are constructed from different technologies, it would a good idea to check out the impulse response of the combination using this method. It is possible for the smaller capacitor to resonate with the inductance of the larger one, causing an unexpected result.

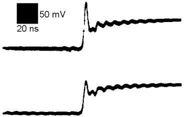


Figure 4. 4uF Capacitor

Figure 5 shows the result for a 1 uF capacitor of the same construction as the 4 uF capacitor tested in Figure 4. Note that the inductance is similar to the 4 uF capacitor, but the ESR is slightly lower. Since an analog scope was used, the waveform circuit to not significantly affect the initial current very much. For this frequency range a generator with a risetime of one to two nanoseconds will do.

If you need to check the capacitor using faster risetimes, it would be best to build the test setup on a small circuit board with a ground plane and controlled impedances. At this point, the parasitic capacitance of the 50 Ohm resistor would also be an issue to be taken into account. Fortunately, such accuracy is often not needed. Especially, if one is just comparing the relative performance of several capacitors.

Data

Figure 3 below shows the initial rise from the generator. The black square indicates the vertical voltage and horizontal time scales. The open circuit voltage was a little over 4 Volts with about a 5 nanosecond risetime. The data in Figures 3 through 6 were taken with an analog scope some years ago. Figures 4 through 6 show data obtained from several leaded capacitors (as opposed to surface mount). Two traces were taken for each capacitor. The lower trace was measured at the capacitor body where the leads entered and the upper trace included the minimum amount of lead to practically connect the capacitor to a printed wiring board. The upper trace would not be needed for modern surface mount capacitors unless one wanted to model the connection inductance from the capacitor to the point of interest on a printed wiring board.

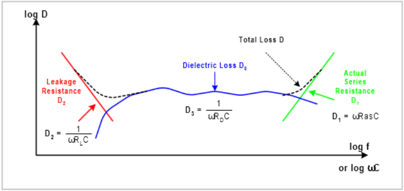


Figure 5: Total Dissipation Factor Plot

At any given frequency we could measure the real part of the impedance, Rs, or measure C and D and calculate it

$$Rs = ESR = \frac{D}{\omega C} = \frac{D_1 + D_2 + D_3}{\omega C}$$

Therefore ESR = $Ras + \frac{1}{\omega^2 C^2 R_L} + \frac{1}{\omega^2 C^2 R_D}$ and thus ESR > Ras, always.

Usually, ESR is very much larger than Ras. However, when ω C is large at high frequencies, high capacitances or some combination, the actual series resistance can cause the largest part of the total D. (See plot.) For very large capacitors (like 0.1F), ESR can be very nearly equal to the actual series resistance even at low frequencies (such as 120 Hz). For most capacitors at low frequencies, the actual series resistance is only a small part of ESR.

For complete product specifications on the 1900 Series Impedance Meters, or any of QuadTech's products, please visit us at http://www.quadtech.com/products. Call us at 1-800-253-1230 or email your questions to info@quadtech.com.

1) $dv/dt = i/C = 50$ mA/C
where C is the value of the capacitor at this low voltage and the risetime of the current << RC.

The offset between the baseline and the beginning of the exponential rise is just the voltage that the current, 50 mA for this case, develops across the ESR of the capacitor. The ESR can be easily estimated in this case by dividing the voltage offset (labeled ESR in Figure 2) by 50 mA.

Parasitic inductance in the capacitor will cause the spike in the waveform shown in Figure 2 exceeding the value of the dotted line along its length. If the current rise were in fact a ramp with constant slope and very sharp corners (high di2/dt) then the spike would be a square pulse of value:

2) $E = L*di/dt$
where L is the parasitic inductance of the capacitor.

The current rise from the generator used for the data in this article was not a ramp with very sharp corners and constant slope (the case for most generators I have used). That characteristic of the generator combined with probe effects led to a peaked shape to the Ldi/dt spike as shown in Figure 2. Using equation 2 the inductance of the capacitor can be calculated. Often, one does not need to calculate the inductance or ESR but just choose a capacitor from several available ones that has the lowest inductance and/or ESR.

Soldering the components onto a BNC connector as in Figure 1, works up to 300 MHz. I estimate the inductive reactance of the loop formed by the capacitor and resistor to be about 20 Ohms at 300 MHz (estimating the inductance at 10 nH). This is small enough relative to the 100 Ohms of resistance in the

For a pulse length that is long with respect to the RC time constant, one will see an exponential rise to the open circuit voltage of the pulse source. For the purposes of this discussion, we will be looking at the first couple of hundred millivolts of a 5 volt exponential rise. An example of this appears in Figure 2.

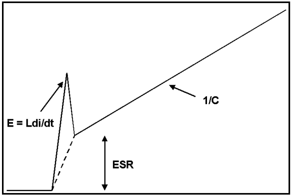


Figure 2. Initial Rise

Figure 2 shows the beginning of the exponential voltage rise across the capacitor when the generator pulse starts. The vertical scale is about 200 mV and the horizontal time is a small fraction of the RC time constant of 100 Ohms and the capacitor being measured. Since the capacitor voltage is still very small compared to the 5 Volt open circuit output of the generator, the current through the capacitor may be presumed to be constant and equal to the generator open circuit voltage divided by 100 Ohms, 50 mA in this case.

The risetime of the current will be the same as the generator voltage. If the rise is a ramp with a constant slope and the capacitor had no inductance, the initial rise shown in Figure 2 would follow the dotted line and then the slope would change to the initial slope of the exponential rise determined by:

High Frequency Measurements Web Page
Douglas C. Smith
http://www.emcesd.com/tt020100.htm

Measuring Capacitor Self-inductance and ESR

Figure 1. Test Circuit for Measurement of Capacitor Self-inductance and ESR

Technical Background

The parasitic parameters of a capacitor, that is its equivalent series resistance (ESR) and its inductance, affect the way the capacitor performs in circuits. Some applications are very sensitive to these parameters. For instance, a bypass capacitor used between power and ground in a digital circuit must be able to supply current quickly to nearby active devices. If it has too much inductance it will not be able to do this.

Similarly, the transient response of a capacitor used to divert a current pulse due to electrostatic discharge is very important to the ability of the capacitor to do its job.

So how can the parasitic parameters of a capacitor be measured? One could certainly connect the capacitor to a network analyzer and get a very good characterization. Such an instrument can be quite expensive though. Even the less expensive capacitance measuring instruments may not be available when needed. Both instruments may not provide the information in an easily usable form. If you have a pulse generator (preferably with a 50 Ohm output impedance) and an oscilloscope, you can easily measure the transient response of a capacitor. From this data the capacitor's ESR and inductance can be determined.

First, construct the simple network shown in Figure 1 at the end of a 50 Ohm coaxial cable fed from a 50 Ohm pulse generator. A 50 Ohm resistor is used in Figure 1 to terminate the coax during the rising edge and provide a total of 100 Ohms of source impedance. The resistor shown is a 51 Ohm 1/2 watt carbon composition resistor with one lead trimmed so that the resistor just seats with the trimmed lead fully inserted into the BNC connector. It may be necessary to put a little solder bump on the resistor lead so that it stays firmly in the BNC connector. The capacitor to be tested is connected between the end of the resistor and the shell of the BNC connector. An oscilloscope is connected directly across the capacitor using leads as short as possible to connect the probe. Probes with a resistive input impedance of 500 to 1000 Ohms are recommended. Standard 10X "Hi-Z" probes often have rising edge effects that will distort the part of the waveform used for the calculations.