MEASURING THE LOSS IN VARIABLE AIR CAPACITORS

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 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 :

$$R_{cap} = R_s + \alpha /(f\ C^2) \quad \text{ohms} \qquad 1.1$$

where f is the frequency
C is the capacitance

The first factor R_s 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 :

$$R_{cap} = 0.0045 + 0.028\ [10^6/(f\ C^2)] \quad \text{ohms} \quad \text{(Jackson)} \qquad 1.2$$

$$R_{cap} = 0.05 + 0.091\ [10^5/(f\ C^2)] \quad \text{ohms} \quad \text{(Moullin)} \qquad 1.3$$

$$R_{cap} = 0.007 + 0.012[10^5/(f\ C^2)] + 10/(f^{\,2}C^2) \quad \text{ohms} \quad \text{(Dye)} \qquad 1.4$$

$$R_{cap} = 0.02 + 0.022\ [10^3/(f\ C^2)] \quad \text{ohms} \quad \text{(Wilmotte)} \qquad 1.5$$

where f is in MHZ and C in pf

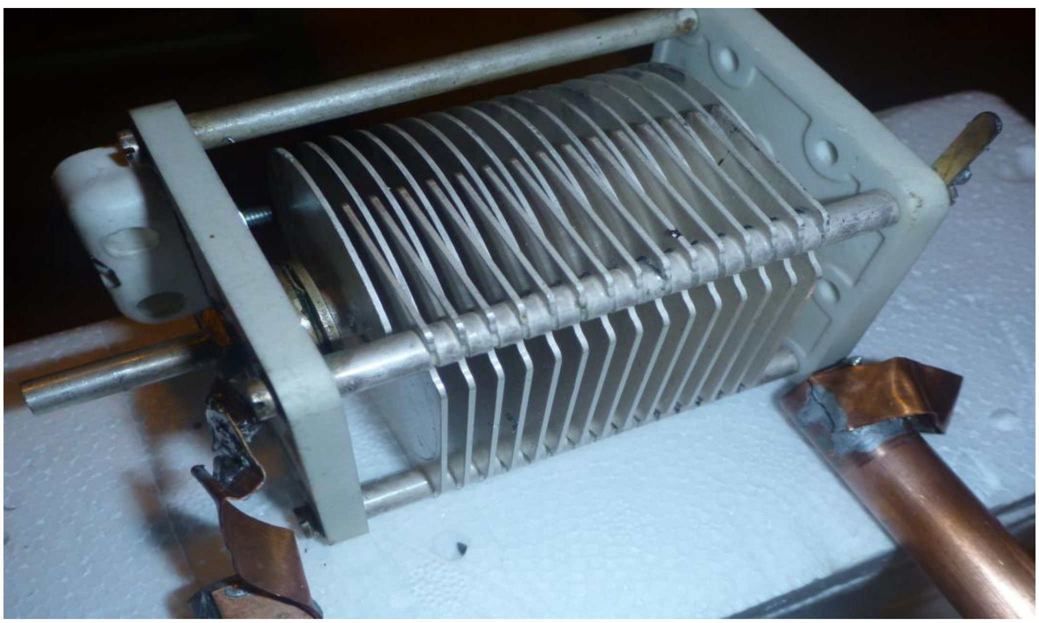
Notice that Dye has an additional term related to a constant resistance across the dielectric, but this term is only significant at very low frequencies, and so is not considered here.

2. EVALUATION OF LOSS COMPONENTS

So the measured value of R_s 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 R_s, 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.



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 (≈ 0.04Ω), 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 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 R_s, 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 (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 stack 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 $L = \mu\, l / 2\pi\ [\ \ln (2\, l / a) - 1\ ]$ henrys, then this will have a free space inductance of 0.05 μH, about 3.5% of the total inductance.

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):

$$L = L_o / [1-(f/f_r)^2] \qquad 4.3.1$$

where L_o is the low frequency inductance
f is the frequency
f_r 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 f_r is chosen which is slightly lower than the actual SRF. A good fit with the measured inductance above is given by f_r = 60 MHz (about 85% of the actual SRF), and the results of the equation with this value are shown in the table above.

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

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% :

$$L = \mu_o\, p/(2\pi)\ [\ \mathrm{Ln}\,(2p/r) + 0.25 - \mathrm{Ln}\,(p^2/A)]\ S_L \qquad 4.4.1$$

where p is the perimeter and A the area
S_L is the inductance SRF factor $1/[1-(f/f_r)^2]$
$f_r = 0.85(150/\ \text{length})$

5. CALCULATED TOTAL LOSS OF RESONANT CIRCUIT

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

$$R_s = R_{wall}\,[l_{pipes} / (\pi\, d_w)]\, S_R\, P + R_r + R_{leads} + R_{cap} \qquad 5.1$$

where d_w is the diameter of the copper pipes
$R_{wall} = (\rho\, \pi\, \mu_o\, \mu_r\, f)^{0.5}$
ρ is the resistivity of the pipes (see below)
$\mu_o = 0.4\pi$ μH/m
μ_r is the relative permeability of the copper pipes (see below)
S_R is the resistance SRF factor $=1/[1-(f/f_r)^2]^2$
P is the proximity factor $1/[1-(d_w/D)^2]^{0.5}$ (see Moullin ref1)
D is the spacing of the pipes (centre to centre)
R_r is the radiation resistance = $31200\,(A/\lambda^2)^2$
R_{leads} is the resistance of the leads connecting the capacitor
R_{cap} 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 μ_r of pure copper is unity, and it is assumed to be the same for plumbing copper.
The frequency f_r 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 :

$$R_{strip} = R_{wall}\; l / [2(w+t)]\; K_{fringing} \qquad 5.2$$

where $K_{fringing} = 1.06 + 0.22\,\mathrm{Ln}\, w/t + 0.28\,(t/w)^2$

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

6. MEASUREMENTS

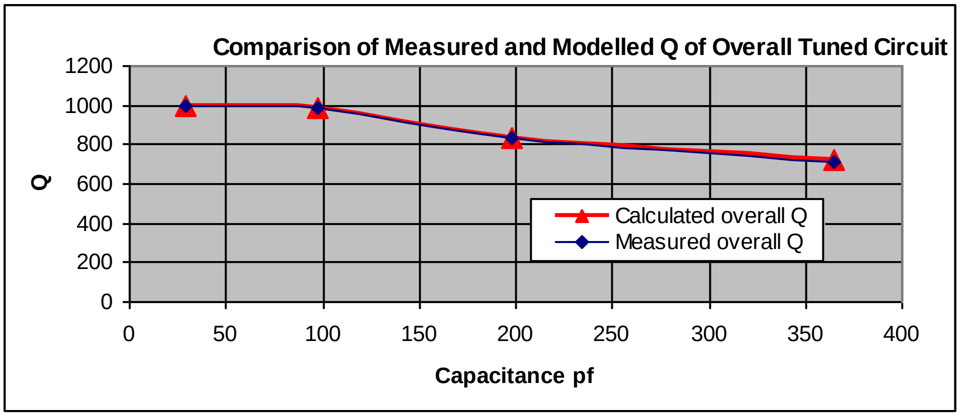
6.1. Procedure and Initial Measurements

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
R_s is given by Equation 5.1

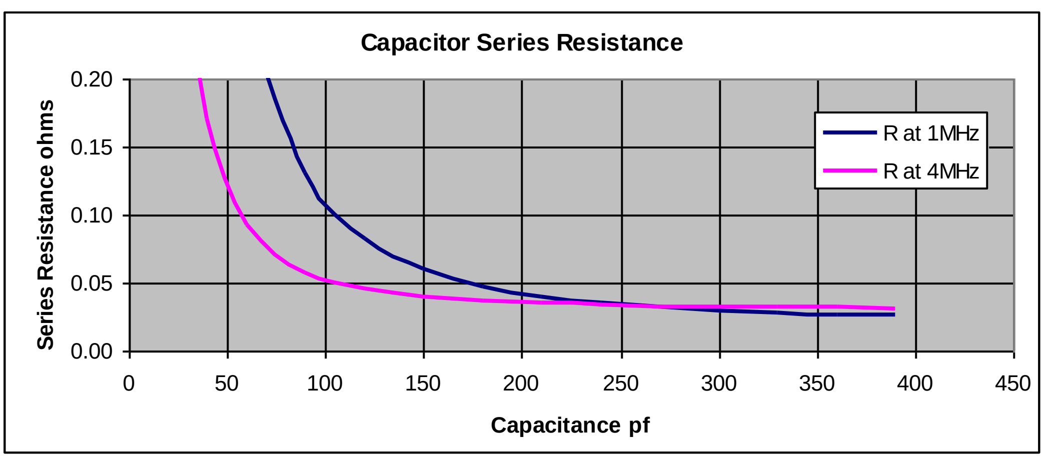
The capacitor resistance R_c was assumed to conform to Equation 2.1, and the constants R_s, α and β were adjusted in the model to give the best correlation between the measured Q and the theoretical Q, and this was achieved with $R_s = 0.01$, $\alpha = 800$, and $\beta = 0.01$, to give the following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



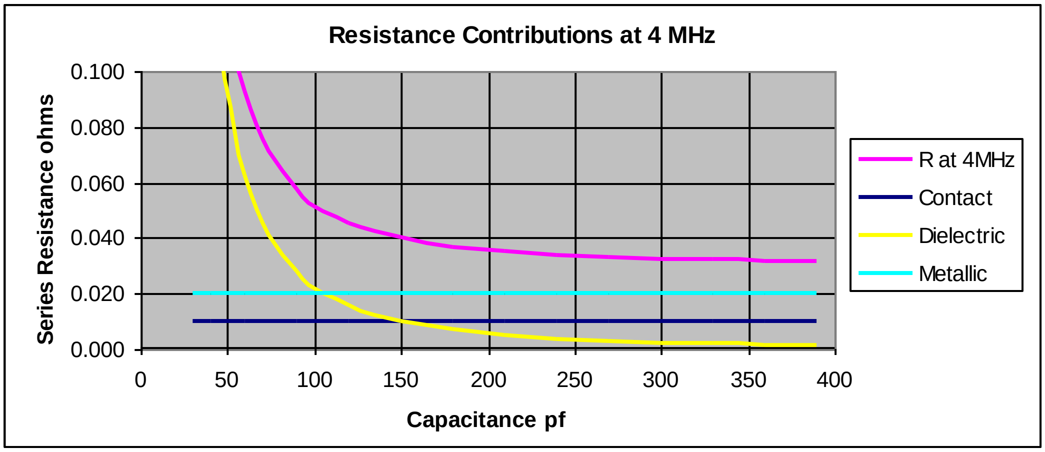
The agreement is within ±1.5%.
It should be noted that the frequency is not constant in the above graph, and is 24.5 MHz at 27pf, 10.2 MHz at 198 pf, 14 MHz at 98 pf and 7.56 MHz at 365 pf, all frequencies corresponding to resonance with the inductor. These combinations of capacitance and frequency give 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:

$$\mathbf{R_{cap} = 0.01 + 800 / (f\,C^2) + 0.01(f^{0.5})} \qquad \mathbf{6.1.2}$$

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

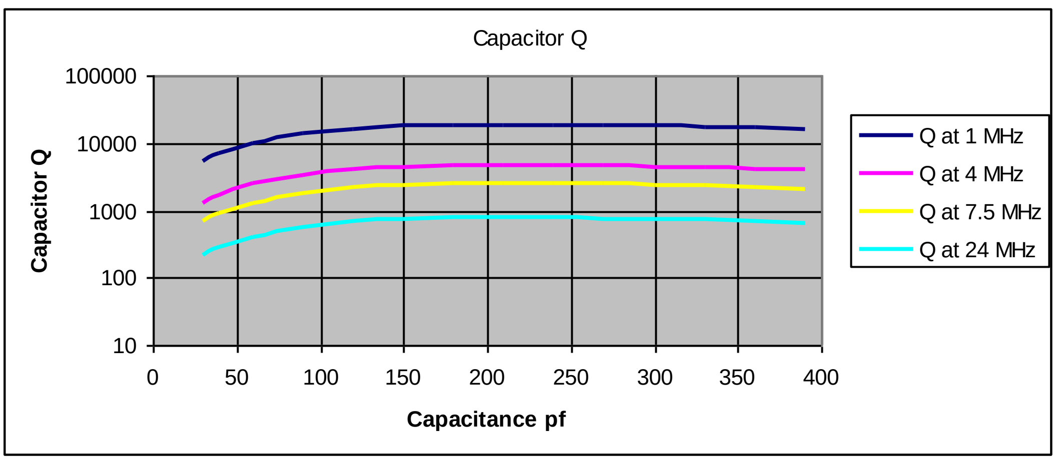


It is useful to know the contributions which each factor in Equation 6.2.1 makes to the overall resistance, and for f = 4 MHz this is shown below:



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 R_{cap}), and is given below:



6.2. Test of Ground Effects

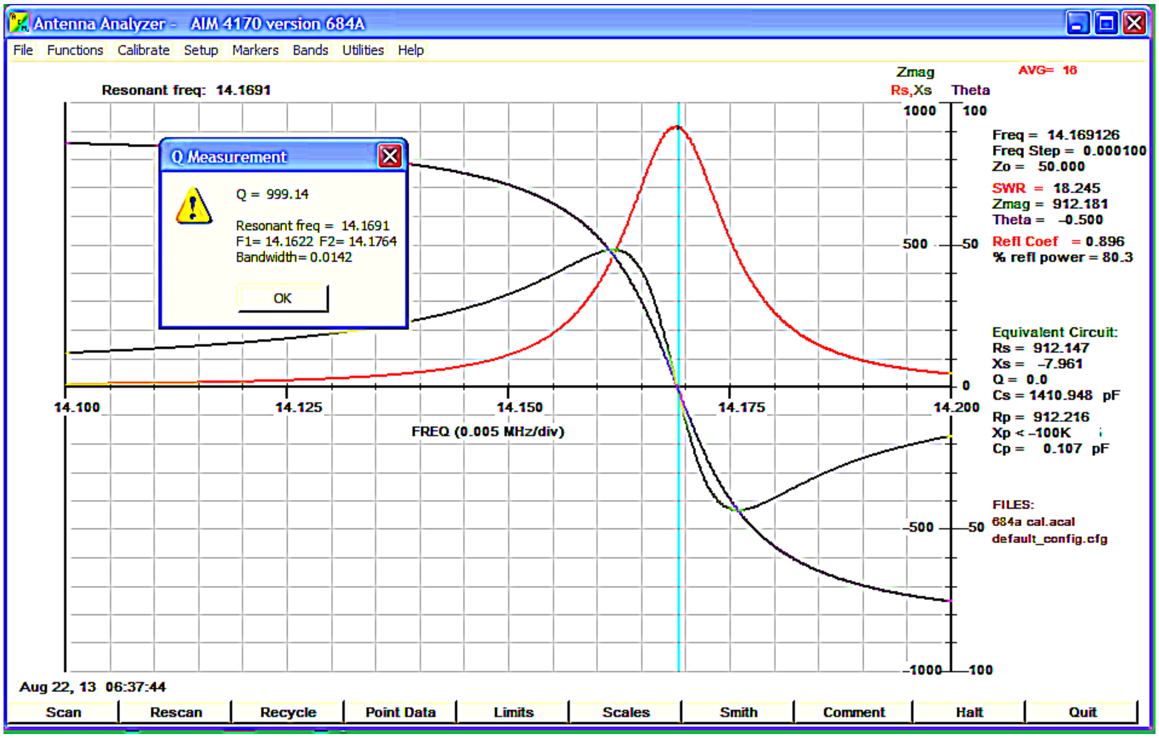
The measured Q was as large as 1000, and so it is possible that the floor of the room and other surfaces could be introducing loss. To evaluate this an aluminium sheet, approx 0.8mx 0.5m, was placed on the floor below the tuned circuit, a distance of about 0.8m. A measured Q of 953 was unaffected, to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

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

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 R_{wall} is ±6 %. Also there was an uncertainty in the room temperature of ± 5^{o} 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 45^o, 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 $\pm 1^o$, this will give a 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, 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 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

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 C_1 due to the supporting insulation, in parallel with a variable air portion devoid of energy loss. The power factor tan δ of the portion C_1 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, C_1. This can be determined from Equation 6.1.2, by setting $C = C_1$, since then the loss is not diluted by the air capacitor and the series resistance due to the dielectric loss will then be :

$$R_c = 800/(f\, C_1^2) \qquad 8.5.1$$

Also from the definition of tan δ :

$$R_c = \tan\delta /(2\,\pi\, f\;\, C_1)$$
for f in Hz and C in farads

$$= 10^6 \tan\delta /(2\,\pi\, f\;\, C_1) \qquad 8.5.2$$
for f in MHz and C in pf (as equation 8.5.1)

Equating 8.5.1 and 8.5.2 :

$$C_1 = 2\,\pi\, 800 / (10^6 \tan\delta)\ \text{pf} \qquad 8.5.3$$

The material of the insulator is unknown but is possibly Steatite, for which tan δ = 0.003, and then the capacitance C_1 due to the supporting insulation is 1.7 pf.

9. SUMMARY

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

$$R_{cap} = R_s + \alpha /(f\, C^2) + \beta\, (f^{\,0.5}) \quad \text{ohms} \qquad 9.1$$

where R_s , α and β are constants for any particular capacitor.
f is the frequency and C the capacitance

The first factor R_s 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:

$$R_{cap} = 0.01 + 800/(f\,C^2) + 0.01\,(f^{0.5})\ \text{ohms} \qquad 9.2$$

where f is in MHz and C in pf

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

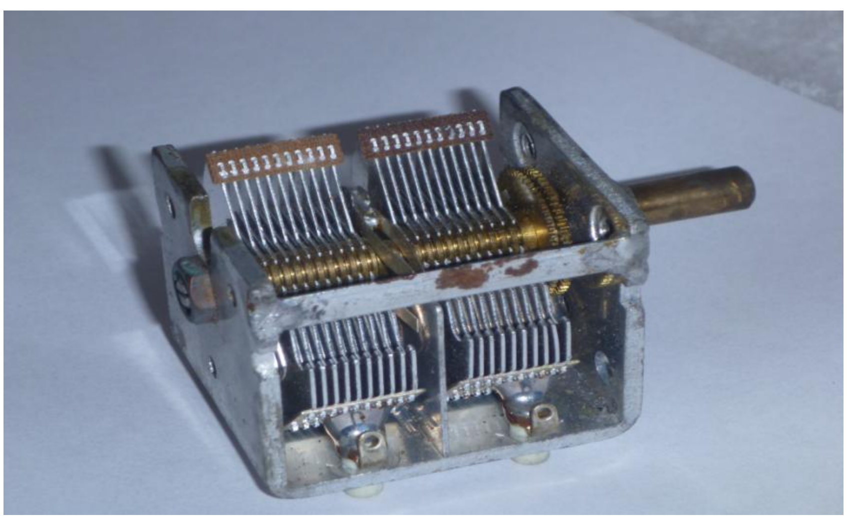
10. ANNEX 1 : BROADCAST CAPACITOR

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:

$$R_{cap} = 0.032 + 5800/(f\,C^2) + 0.0039\,(f^{0.5})\ \text{ohms} \qquad 10.1$$

where f is in MHz and C in pf

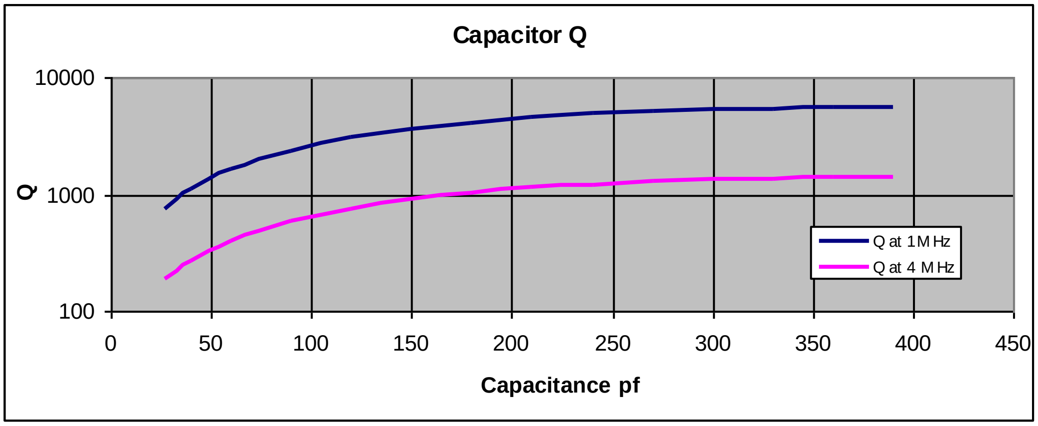
The accuracy of the above equation is estimated at ± 13%.



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 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 mΩ compared with 10 mΩ. 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 R_v will be :

$$R_v = 2\, Z_{wall}\, l / (N\, w) \qquad \text{A1.1}$$

where $Z_{wall} = \rho / \delta = (\rho\, \pi\, \mu\, f)^{0.5}$

ρ is the resistivity of the metal

μ 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

If the plate is silver (as here) , then $\rho=1.629\ 10^{-8}$, and $\mu = 4\pi 10^{-7}$, so $Z_{wall} = 8\ 10^{-4}$ 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:

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

So at maximum capacitance we can expect the loss resistance in the plates to be, $R_v = 2\, Z_{wall} / (\pi N) =$ 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.

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.

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

Enquiries to paynealpayne@aol.com