MAGNETIC LOOP OR SMALL FOLDED DIPOLE?

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INTRODUCTION

The single turn tuned loop antenna as shown in Fig. 1 is a remarkably effective 'small' transmitting antenna. Small is defined as having a maximum physical dimension which is considerably less than a wavelength of the frequency of operation. Effective performance can be obtained for loops with diameters down to as little as two percent of the wavelength. For the HF band of 1.5 to 30MHz two percent corresponds to loop diameters of 4 metres down to 20 centimetres. For deployment in the confined space allowed to many if not most of the users of the HF band such a small size is of great importance.

A further important feature is that a tuned loop antenna can have a frequency tuning range of up to 3 to 1, depending primarily on the minimum to maximum capacitance ratio of the tuning capacitor. Furthermore the on tune match resistance is remarkably constant over the whole of this tuning range providing the antenna is not used in an electromagnetically confined space.

For reception the tuned loop antenna also performs effectively. In this case there is little difference in performance between a small and a full size antenna, where full size means having dimensions around half a wavelength or more. This is because the most important criterion for effective reception is signal to noise ratio and not antenna efficiency. In the HF band, particularly at the low frequency end, external manmade and galactic noise is dominant, and large antenna losses can be tolerated with little or no loss in received signal to noise ratio.

Most of the evidence for the 'effective' performance of the tuned loop is anecdotal and qualitative, not quantitative. Some of the reasons for this are explained in the following sections. Theoretical predictions remain pessimistic with respect to the performance actually being achieved by users of the antenna, but it is not easy to find evidence adequate to be able to prove whether there is a real discrepancy between theory and practice.

A number of tuned loop antennas are assessed in reference 1. Measurements of Standing Wave Ratio (SWR) are given at various frequencies and used as a means of indicating the typical operating bandwidth and (loaded) Q values. At present other than from amateur radio publications there is very little published information on the performance of transmitting loops. The AMA3 from Advanced Antennas and Ancillaries Ltd has been used for most of the measurements given in this paper.

The main purpose of this paper is to propose an additional and different mode of operation for the loop antenna and to predict and give some initial evidence that this mode is usually the dominant mode of operation. This 'dipole' mode has a polarisation and radiation pattern which differs significantly from the normal loop mode. However, as will be seen this difference can in practice be expected to be less apparent in the confined environment of typical deployment and operation.

THE BASIC LOOP ANTENNA

Fig. 1 shows the basic loop antenna. It consists of a conducting loop which is tuned by a high voltage tuning capacitor at one point on the loop circumference. The tuning capacitor is usually adjusted to resonance by remote control using some form of servo and motor drive on the capacitor. A subsidiary coupling loop is positioned usually on the opposite side of the mains loop to the tuning capacitor. The diameter of this smaller loop is adjusted for an optimum match to the feed cable. Small adjustments to the matching can be achieved by moving the coupling loop towards the centre of the tuned loop to decrease the coupling a small amount.

The main advantage of the tuned loop antenna is its small size. The main disadvantages are its narrow fractional bandwidth and when used for transmitting the high rf voltages which occur across the tuning capacitor. Both disadvantages are a consequence of the intrinsically high Q of this, and in fact of any, small antenna.

The Inductance of a Single Turn Loop

As will be seen the inductance of a single turn loop is a crucial parameter for loop antenna design. There are numerous formulas for inductance in standard textbooks but most of these have been derived to express the inductance of coils which have more than one turn and are dimensionally small with respect to a wavelength. Even for small coils, where turn to turn phase shift effects due to time delayed coupling can be neglected, the most accurate inductance formulas would appear to have been derived mainly empirically rather than entirely from first principles. An explanation for this could be that, just as claimed by some authors, more than one electromagnetic coupling mode usually contributes to the actual inductance, and the relative contributions from each mode depend on frequency, coil shape wire diameter, number of turns and the turn spacing of the coil.

The single turn inductance formula used in this paper is semi-empirical. It is derived by consideration of the coil as a transmission line so that variation of inductance with frequency can be taken into account. The calculated characteristic impedance (Z_o) of the transmission line experimentally proves to be about 20% in error with respect to measured results with no obvious way of theoretical reconciliation. Therefore an empirical scaling factor is introduced to align theory with practice. For a reason which as yet has not been explained the correction is found to correspond with the assumption that the tube of the antenna is 'electrically' twice as large as the physical diameter of the tube. The main utility of the theoretical formula is then for interpolation and extrapolation of the variations of inductance and inductive reactance with frequency.

The single turn loop is taken to be a short circuit transmission line stub. The length of the stub is taken to be half the wire length. In the case of a circular loop this is half the loop circumference or pi times the radius.

The characteristic impedance of the stub is taken to be the same as for a piece of two wire transmission line with spacing equal to the loop diameter and with the same wire diameter as for the wire (or tubing) of the loop. This is clearly open to question as a model. But the assumption of constancy of Z_o even when the spacing of the conductors varies can be justified in at least two ways. First as the wires come closer together at the two ends of the stub in both bases the angle between the wires progressively departs from the parallel condition. It is argued that the angle variation and the separation distance variation substantially cancel in effect to provide essentially constant coupling between the two equivalent wires and so the Z_o stays constant. The velocity along the wires is also assumed to be constant and this is confirmed by the fact that the quarter wave open circuit resonant frequency for a loop is indeed very close to the frequency at which the total length of the wire is a half wave.

A further justification of the constancy of Z_o assumption is that theoretically Z_o varies as the logarithmic ratio of the wire spacing to the wire radius, or in this case the logarithmic ratio of the loop diameter D to the wire radius a. The logarithm of a ratio actually varies little for large ratio changes and so the value of Z_o is little affected even by large changes in wire radius and spacing.

The theoretical value of Z_o is taken to be as derived in Kraus (2) and in ohms it is

$$Z_o = 276 \log (D/a) \qquad (1)$$

where D is the loop diameter and a is the wire (or tube) radius. (The constant 276 is actually $(1/\pi)\ \sqrt{(\mu_o/\varepsilon_o)}$ $\lambda o \gamma_\varepsilon$ 10 $\alpha v \delta \sqrt{(\mu_o/\varepsilon_o)}$ = 377 ohms, the intrinsic impedance of free space).

The input impedance of a shorted stub of length x (= half the loop circumference pD/2) is ZL = jZo tan(2px/λ) = jZo tan (2pf/x/c) where f = c/λ = jZo tan (pf/2fo) (2)
for a frequency $f = \omega/2\pi$ this gives an equivalent inductance of

$$L = Z_L/j\omega = Z_L/j2\pi f = Z_o \tan(\pi f/2f_o)/2\pi f \qquad (3)$$

We have for the loop
$$f_o = c/\lambda_o = c/4x = c/2\pi D \qquad (4)$$

If $\pi x/\lambda$ is small, for example less than $\pi/8$, or for $f \leq f_o/2$ the approximation tan y = y can be made.

$$L_o = Z_L/j\omega \approx Z_o/4f_o$$
(5)

The AMA3 loop antenna dimensions are diameter D = 833mm and tube radius = 16mm. At 14.0MHz the loop was found to resonate with a capacitance of 73pF corresponding to an inductance value of 1.77mH.

For the AMA3 loop equation 1 gives the characteristic impedance Z_o as

$$Z_o(\text{theory}) = 276 \log (833/16) = 474 \text{ ohms} \qquad (6)$$

The calculated self resonant frequency f_o of the AMA3 loop is thus given by equation 6 as
$$f_o(\text{AMA3}) = c/2\pi D = 57.3\text{MHz} \qquad (7)$$

Equations 6, 7 and 8 then predict a low frequency loop inductance which is calculated

$$L_o(\text{calc}) = 474/(4 \times 57.3) = 2.07\ \mu\text{H}. \qquad (8)$$

The calculated inductance at 14MHz is obtained using equations 2 and 3 as

$$L = Z_0 \tan(\pi f/2f_0)/2\pi f \quad (9)$$

At 14 MHz the AMA3 inductance is therefore calculated as

$$L(\text{calc: } 14\text{MHz}) = 2.18\ \mu H \quad (10)$$

This value is about 1.23 times larger than the value measured at 14MHz which is 1.77 Z_0.

It is interesting to note that tube diameters would have to be increased by about 2.09 times in order to reconcile the measured result with equations 1 and 2. It is therefore tentatively proposed that a better formula for the Z_0 of a single turn loop is

$$Z_0 (\text{empirical}) = 276 \log (D/d) \quad (11)$$

where d is the diameter of the tube and D as before is the diameter of the loop. (The diameter of the tube has replaced the radius of the tube which was used in equation 1).

This is the value of Z_0 which should therefore be used in equation 3 for inductance taking frequency variations into account, and equation 5 which gives the low frequency (asymptotic) inductance.

On this basis the AMA3 loop inductance at 14.0MHz is calculated to be 1.79 µH which is less than 2% in error with the measured value. The value of Z_0 reduces from 474 to 391 ohms when the formula of equation 11 is used. The modified equation would appear to give results well within the experimental error estimated as better than ± 3%.

Theoretical Loop Mode Radiation Resistance

Balanis (3) shows that the radiation resistance of a small single turn circular loop of circumference C with uniform current can be derived to be

$$R_r = 20\pi^2 (C/\lambda)^4 = 20\pi^2 (\pi D/\lambda)^4 \quad (12)$$

This formula is generally accepted in most if not all text books to be a valid expression for the radiation resistance appearing in series with the loop inductance.

For the AMA3 at 14.0MHz the radiation resistance is calculated from equation 12 to be

$$R_r (\text{loop: } 14\text{MHz}) = 0.0439 \text{ ohms} \quad (13)$$

To extrapolate to adjacent frequencies equations 13 and 12 can be combined to give

$$R_r (\text{loop: fMHz}) = 0.0439 (\text{fMHz}/14)^4 \quad (14)$$

This shows a radiation resistance which increases as the fourth power of the frequency.

From the inductance value measured at 14MHz we can derive the loop antenna unloaded Q_u as

$$Q_u (\text{loop}: 14\text{MHz}) = \omega L/R_r = 3546 \quad (15)$$

which as will be seen is much higher than can be measured. It would also imply that the power dissipated in the loss resistance of the tubing is likely to be much higher than the radiated power.

Extrapolating towards the higher end of the AMA range at 28MHz the inductance value doubles while the radiation resistance increases by 2^4 which is a factor of 16 times. On this basis we have for the 28MHz unloaded Q_u.

$$Q_u (\text{loop}: 28\text{MHz}) = 3546/8 = 443.3 \quad (16)$$

Also R_r (loop : 28MHz) = 0.7024ohm

The conclusion is that this particular loop mode cannot contribute significantly to radiation at 14MHz although at 28MHz significant loop mode contribution to the radiation might be expected.

A more quantitative evaluation can only be reached by comparing the radiation resistance of this loop mode with both the loss resistance and with the radiation resistance of any other mode which may be excited at the same time.

PROPOSED DIPOLE MODE

The main purpose of this paper is to propose that a dipole mode of radiation is usually dominant for a small tuned loop antenna. The dipole mode is essentially the same as for the folded dipole for example as described in Balanis (4).

Radiation Resistance of Proposed Dipole Mode

The dipole mode is assumed to be excited by the voltage which appears across the capacitor when power is transformer coupled into the tuned loop from the source connected to the subsidiary loop.

Because the inductance per unit length can be assumed to be constant around the loop the rf voltage between the two sides of the loop decreases linearly from the capacitor terminals along the conductors until the zero voltage point opposite the capacitor. Making the assumption that the field radiated from the loop depends on the average of the voltages seen in the direction of radiation it can be seen that the effective voltage for radiation is as if half the capacitor voltage had been placed across the loop diameter at the point half way from the capacitor to the opposite zero voltage point. Fig. 2 gives an example showing that 4kV across the capacitor is equivalent to 2kV across the ends of the equivalent (horizontal) dipole.

It is now assumed that the dipole mode radiation resistance can be calculated as if for a folded dipole with the usual much closer spacing of the conductors. The equivalent dipole mode current is assumed to be equal and in the same parallel direction in the two sides of the loop at the capacitor at the top and opposite to the capacitor at the bottom. The dipole current distribution is assumed to decrease essentially linearly from these points around the various loop conductor paths to the sides of the loop. The distribution is actually the tip of a sinusoidal current distribution which for a small fraction of a wavelength can be taken to be linear. But the angle of the current to the loop horizontal diameter also has to be taken into account.

The far field is calculated from the horizontal current component which has the effective mean value of

$$i_{eff} = \frac{\int_0^{\frac{\pi}{2}} \theta \sin\theta . . d\theta}{\int_0^{\frac{\pi}{2}} \theta} = \frac{2}{\pi} \qquad (17)$$

The mean value of a linearly decreasing current is 0.5 and so for the same current the far field of the loop acting as a dipole is $4/\pi$ times stronger than a small dipole of length equal to the diameter. This means the radiation resistance is $(4/\pi)^2 = 1.62$ times higher.

Thus the short dipole radiation resistance formula given by Balanis (5) can be multiplied by a factor $(4/\pi)^2$ to give the estimated dipole mode radiation resistance seen in series with the antenna loop.

$$R_r \text{ (loop dipole)} = (4/\pi)^2 \times 20\,\pi^2\,(D/\lambda)^2 = 320\,(D/\lambda)^2 \qquad (18)$$

At 14MHz for the AMA3 we have

$$R_r \text{ (loop-dipole : 14mhz)} = 0.484 \text{ ohm} \qquad (19)$$

At 14MHz Z_L = j155.7 ohms giving a Q_u of

$$Q_u \text{ (loop-dipole : 14MHz)} = 155.7/0.484 = 321.7 \qquad (20)$$

The corresponding values predicted for 28MHz assuming the loop inductance is little changed are

$$R_r \text{ (loop-dipole : 28MHz)} = 1.94 \text{ ohms} \qquad (21)$$

$$Q_u \text{ (loop-dipole : 28MHz)} = 163.6 \qquad (22)$$

By comparing equations 19 to 22 to equations 13 to 16 it can be seen that the dipole mode always has the higher radiation resistance, by a factor of 11 times at 14MHz reducing to a factor of 2.76 times at 28MHz. Thus at 14MHz only 8% of the radiation is predicted for the loop mode and at 28MHz this is expected to rise to 27%. The dipole mode will thus predominate over this band of operation. This would also be predicted to be the case for all the antennas reviewed in reference 1.

The accuracy of these predictions can be approximately assessed by comparing the predicted Q_u value of 322 at 14MHz to the measurements of Q around 250 obtained with the AMA3 antenna about five metres above ground in an attic of a house. Assuming no antenna losses the difference in Q values could be attributed to an additional loop radiation resistance of 0.14ohm, implying a loop mode contribution of a maximum of 22% rather than 8%. However this additional loop mode contribution is considerably larger than predicted by equation 12.

Determination of the Radiation Mode

In principle the measurement of the plane of polarisation of the radiated signal in all directions can be used to determine the relative strengths of the dipole and loop modes. However the tuned loop antenna is usually only used in relatively confined spaces or close to the ground or buildings. In such cases it is unlikely that the free space antenna pattern will be retained because of reflections from, and coupling to, adjacent surfaces.

Simple measurements with dipole and loop field probes do in fact indicate that the dipole mode fields are considerably stronger in the near-field region. However this is no guarantee that the dipole field dominance will be maintained in the far field. Coupling to nearby objects can also be expected to cause polarisation changes in the far field. For example it is well known that the vertically polarised surface wave on a reasonably conductive earth is propagated preferentially to considerable distances. Thus any measurements based on the relative strengths of horizontal and vertical polarisation will be suspect in the far field, at least at low angles of radiation.

Some far field measurements of relative polarisation strengths are planned. These can be obtained by rotating the loop antenna in the horizontal plane. It is expected that these will be able to provide further confirmation of the dipole mode dominance although a high degree of accuracy cannot be expected with any confidence.

Determination of Antenna Efficiency

It is not easy to determine antenna efficiency with any accuracy. One definition of efficiency η is the ratio of the actual antenna gain to the antenna directivty in the preferred direction. The antenna directivity is best determined theoretically. In free space this is easy because the antenna pattern is well behaved. In confined spaces it is not practical to calculate the antenna pattern with any degree of confidence. Even given a value for the directivity the antenna gain has to be measured with great accuracy particularly if he efficiency is high. For example if the efficiency is 90% the antenna gain and directivity differ by less than 0.5dB.

"Height gain" is another effect which can give rise to great confusion in the calculation of antenna gain. An antenna placed at the top of a reasonably conductive building will appear to have a gain which can appear to be much in excess of the free space directivity or gain. The explanation is that the radiation field lines are compressed together at the top of the building and the fields are considerably enhanced. In fact a short monopole at the top of the building couples strongly into the building which is in effect the major part of the combined antenna system. This is still true for a horizontal dipole at the top of a broad building. A loop, which can couple into the currents induced in the building structure, can be expected to have best enhanced performance when placed about a quarter wave down from the top of the building. When height gain is present apparent antenna efficiencies can easily exceed 100% because the distortion in the antenna pattern actually is such as to give increased gain usually in a horizontal direction.

The environment can either enhance of degrade the antenna radiation resistance. If reflections from, or coupling to, the environment increase radiation from the antenna the antenna Q will be lowered. But then the enhanced radiation from the antenna may be being absorbed by the environment and not being radiated into free space or it may actually enhance radiation at least in some directions. On the other hand losslesss reflections from the environment can either enhance or degrade the radiation resistance and Q but in general in either of these cases radiation is enhanced in at least some directions.

The environmental variations in Q can in practice be quite large, up to two to one increases or decreases are not uncommon. Although the tuned loop is a small antenna its radiation resistance can be expected to vary in just the same way in which larger antennas such as the half wave dipole or quarter wave monopole vary in the real environment. For example the presence of a lossless ground can potentially halve the (series) radiation resistance and double the Q. In this case half the potential radiation pattern is suppressed. The remaining pattern is likely to be enhanced in at least some directions. The results obtained so far would appear to confirm this, although fortunately the reactive de-tuning of the loop by the environment is found to be much smaller than for a larger antenna.

Antenna efficiency can in principle be measured accurately by the "Wheeler Cap" method attributed to H A Wheeler (6) The antenna Q at the operating frequency is measured in free space as Q_f and inside a sealed perfectly conducting enclosure as Q_e. The radiation resistance is theoretically reduced to zero in the enclosure if it is perfectly reflective and lossless. The efficiency can then be calculated as

$$\eta = (Q_e - Q_f)/ Q_e \qquad (23)$$

This technique was used to provide a worst case efficiency estimate of the AMA3 loop at 14MHz. The antenna Q was measured in three different screened rooms. Surprisingly the highest value was recorded in an anechoic chamber which had just been constructed. This room was lined with aluminium foil under the anechoic material which has a low frequency cut off frequency (for absorption) around 1GHz. At 14MHz the Q value was measured as 790 against the Q of 718 in the best of the other two screened rooms. However both of these rooms contained extraneous material (books, cupboards and desks), which probably contributed significantly to the room losses.

Since a network analyser was used for most of the SWR measurements Q ratios could also be obtained from the ratios of the measured resistances. In general this was found to be less accurate than direct bandwidth measurement.

The Q_e value of 790 when compared with a free space measured Q of around 250 gives from equation 23 a worst case efficiency of

$$\eta\,(\text{worst case : 14MHz}) = 1 - 250/790 = 0.68 = 68\% \qquad (24)$$

The value of 68% corresponds to a antenna gain reduced by about 1.7dB.

This predicted value is probably pessimistic. Some evidence is that the AMA3 antenna does not get perceptibly warm to the touch even after prolonged operation with 400 watts (PEP) of peak SSB speech power. Further measurements using thermometers to measure temperature rises are planned although it will be difficult to measure the heat capacity of the antenna with any accuracy so the overall measurement of accuracy will not be particularly good.

Q Measurement Methods and Results

The 3dB bandwidth of a reasonably high Q antenna can be obtained by standing wave ratio (SWR) measurements. At the 3dB frequencies the SWR can be shown to be $(\sqrt{5} + 1)/(\sqrt{5} - 1) = 2.62$ provided that the antenna has been matched by a lossless antenna matching unit (AMU) at the centre frequency.

In principle this is a very good method for measuring the Q of a tuned loop antenna. In practice because the antenna has high Q any matching unit losses markedly degrade the measurement accuracy. In fact none of the commercial available AMUs were found to be sufficiently low loss. The main problem would appear to be contact resistance in variable capacitors. On receive, or for the low powers used in network analysers, "roller coaster" mechanically variable inductors are unusable. Contact resistances were found to be highly variable and a few watts of power obviously are needed to "microweld" the contact points to a low resistance condition. The best variable capacitors are those which have, or have been modified to have flexible wire connections direct to the rotor.

A further difficulty with the SWR method for measuring Q is that the 'on tune' SWR should be better than 1.02 : 1 for results to be obtained to about 3% accuracy and repeatability. The L network AMU as described in reference 7 is found to be suitable in this application.

Some results from the SWR method are shown in Fig. 3. The AMA3 antenna was located on a flat roof for these results.

Because of initial problems with variable contact resistance a 'direct' method of Q measurement was investigated. The coupling loop was removed from the antenna and a 'gamma' match top made to the lower part of the AMA3 loop opposite the tuning capacitor. A feed was taken from the tracking generator of an HP spectrum analyser to the gamma match through a 1k resistor. The input to the spectrum analyser was then taken from a small 5 cm diameter two turn loop placed to couple inductively with the main tuned loop. A 1k resistor was placed in series with the output of this loop in order to minimise any coupling.

Some results from this method are shown in Fig. 4. For frequencies around 14MHz the Q values would appear to be considerably higher than obtained by the SWR method. There is some evidence that there may be more than one mode present each of which is radiating and coupling into the environment. Partial radiation pattern cancellations of these two modes could account for the much higher Q values than expected, and different relative excitation of the two modes could account for the difference in measured Q values for the two methods. Clearly further investigation is needed to reconcile this method of Q measurement with the SWR measurement method. As shown in Fig. 4 the measurement frequency was again extended down to 3.9MHz by connecting additional high value variable capacitors. The substantial variation in the results in this case is attributed to variable contact resistance, although an alternative explanation is if more than one mode is being excited.

In both Figs. 3 and 4 of particular interest is the region below about 10MHz. In both cases the general shape of the curve would appear to agree with the assumption of a rf variation in loss resistance in the tuned loop which is proportional to $f^{\frac{1}{2}}$.

Thus in conclusion it is felt that the SWR Q measurement method is the most appropriate method for assessment of tuned loop performance. For this method the excitation of antenna modes is likely to be the same as for normal operation.

Loop Dipole Mode Equivalent Circuit

Fig. 5 shows an equivalent circuit which would appear to be a good representation of the tuned loop antenna over the 2 to 3 : 1 frequency band of typical operation in an approximately free space environment.

The radiation resistance has been transformed from its original position in series with the inductor L to be across the tuning capacitor C. The remarkable fact is that to good approximation the resistor R_p turns out to be almost constant, because the series radiation resistance R_s varies as the square of the frequency. The standard series to parallel transformation is

$$R_p = (X_L)^2/R_s = (2\pi fL)^2/R_s = X_LQ \qquad (25)$$

Using the measured Q value of 260 and $X_L = 155.7$ at 14MHz gives $R_p = 40.5k$. This gives for the parallel equivalent radiation resistance across the capacitor:-

$$R_p = 40k \qquad (26)$$

Not only is this resistance invariant with frequency it is also invariant if the loop size is scaled up or down at least until the self resonant frequency is approached. The loop inductance is directly proportional to loop size and the radiation resistance increases as the square of the loop size (see equation 12) In equation 25 these two variations cancel leaving R_p constant at 40k.

A consequence of the invariance of R_p is that it is possible to relate the voltage on the tuning capacitor directly to the antenna input power W. We have for

$V_{cap} = \sqrt{(R_p W)} = 200\sqrt{W}$ (27)

For 400 watts V_{cap} = 4kV irrespective of frequency or loop size, assuming no losses.

The equivalent turns ratio between the tuned loop and the matching loop is in ratio to their respective areas. (This is the ratio of the amount of flux coupled from the large loop into the small loop). If both loops are circular with respective diameters D and D_m we then have for the impedance transformation at loop resonance

$$\frac{R_p}{R_{in}} = n^2 = \left(\frac{A_c}{A_m}\right)^2 = \left(\frac{D}{D_m}\right)^4 \qquad (28)$$

For R_{in} = 50 :- D/D_m = 5.3 (29)

The theoretically correct diameter ratio is thus 5.3 under all circumstances. In practice coupling reduces when the matching loop is moved away from the tuned loop conductor towards the tuned loop centre. Coupling is also reduced if the match loop 'wire' diameter is reduced. In either case the match loop diameter should be increased.

Confirmation of all aspects of this very simple model is given by the results reported by P J Hart in reference 1. For each of the several loop antennas tested the on tune SWR values remained generally constant. The largest departures appear at the lower frequencies of operation in all cases. This is where resistive (skin resistance) losses can be expected to become appreciable. In theory if a loop is correctly matched at the higher frequencies, then the SWR at low frequencies is exactly the proportion by which the series radiation resistance is raised by the series loss resistance. On this basis all the loop antennas reported would have efficiencies of better than 50% since this corresponds to a 2 : 1 SWR at the low frequency band edge. (The worst case of 1.8 : 1 for the AMA3 at 14MHz actually corresponds to an efficiency of 56%).

This method of estimating loss is again considered to be rather pessimistic. Coupling into ground losses will also account for more severe degradation at low frequencies if the antenna is operated near the ground.

CONCLUSIONS

The tuned loop antenna does not appear to operate significantly in its loop mode. The small folded dipole mode proposed here gives much better agreement between theory and measured results. It provides an explanation of why the loop antenna performs so well in practice. A very simple equivalent circuit derived theoretically has been shown to give predicted results in reasonable agreement with experiment. The circuit has useful invariant properties which result in very simple design rules for matching loop ratio to tuned loop size. The voltage on the tuning capacitor is also shown to be an invariant for a given input power to the antenna. A new semi-empirical formula for loop inductance has been derived.

The radiation resistance of the loop antenna is sensitive to the environment but no more so than for a full size dipole or monopole. The tuning point is almost unaffected by environmental changes.

Because the dipole mode in practice always is the dominant mode and never the loop mode, it is proposed that the term 'magnetic loop' antenna should not be used. The term 'small folded dipole' is a better term to use. However the term 'tuned loop' antenna is more generic. It would remain correct if new modes of loop or dipole excitation prove to be possible with new as yet unexplored tuned loop arrangements.

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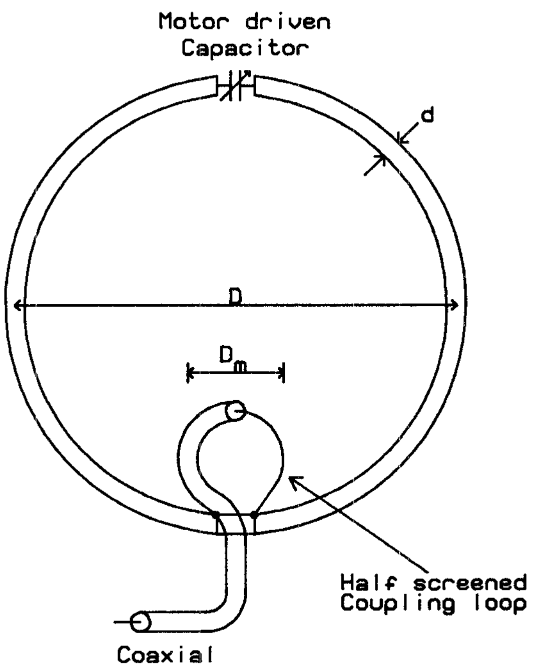


For AMA3
D = 833mm
d = 3.2mm
Dm = 170mm

Fig 1 Tuned loop antenna

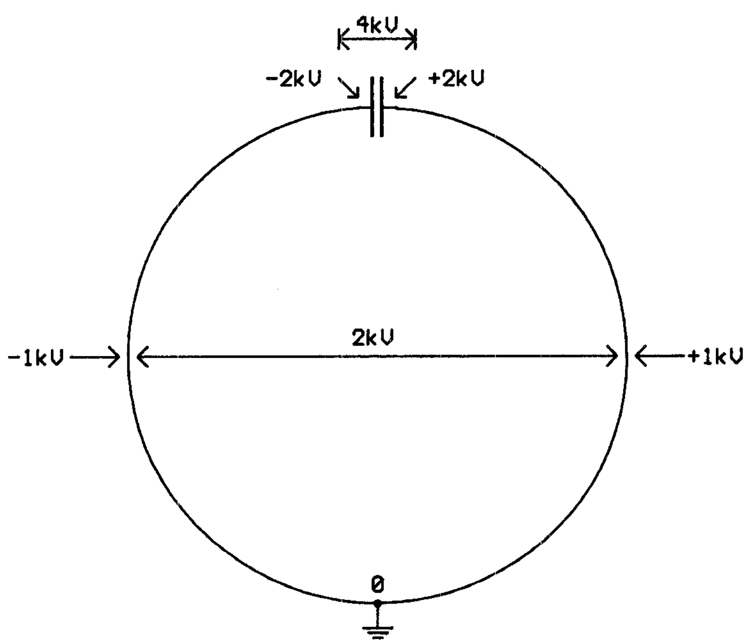


Fig 2 Voltage on Tuned loop

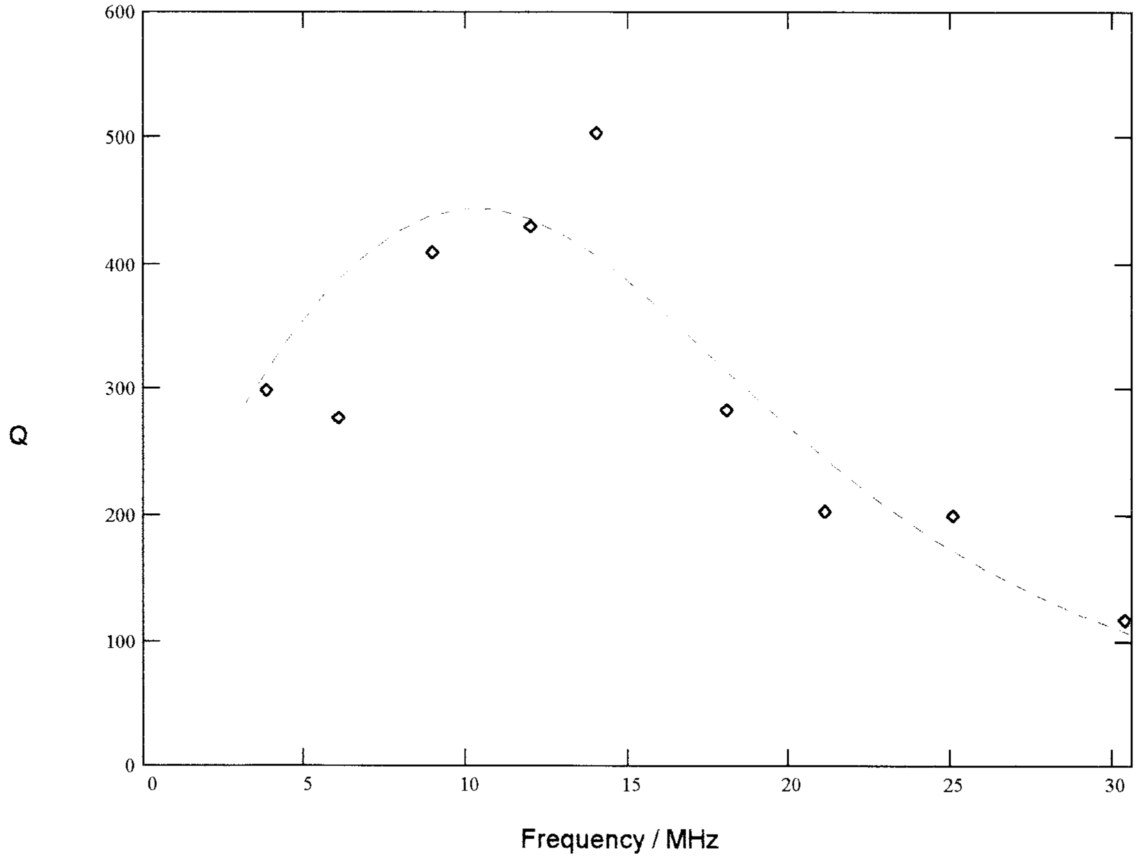


Fig 3 Loop Q by VSWR measurement

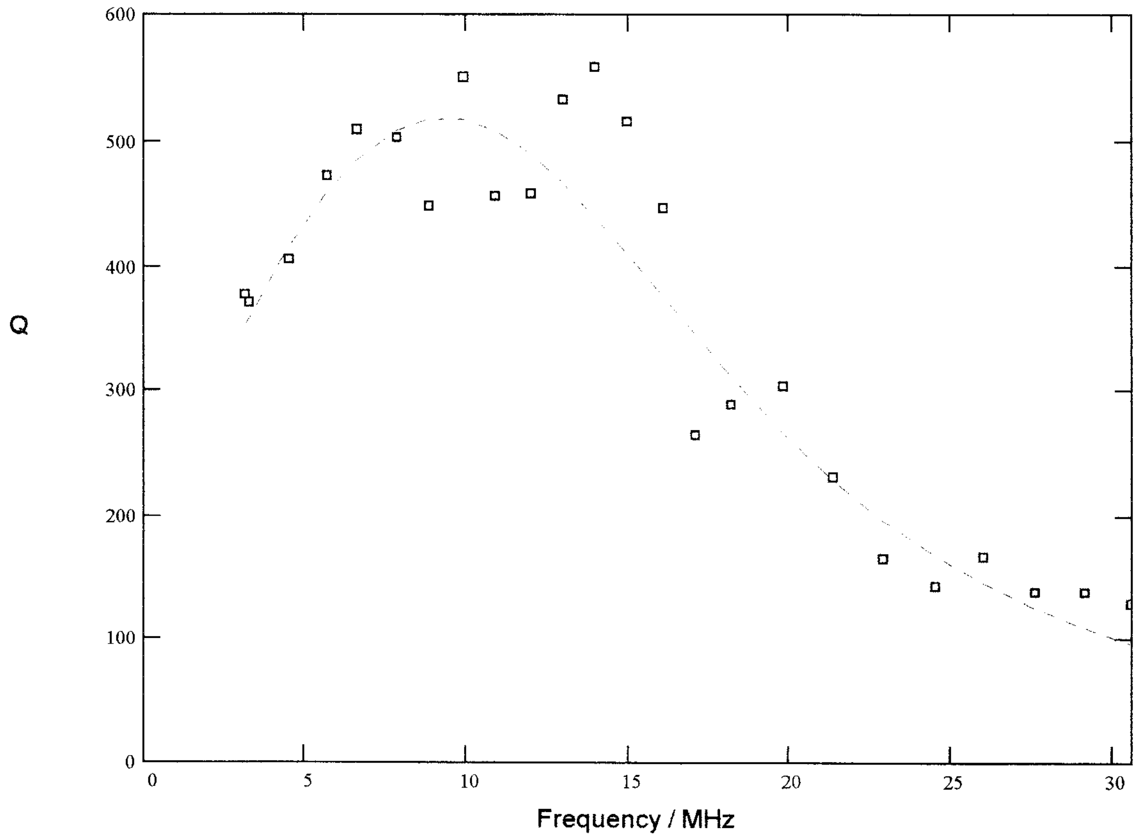
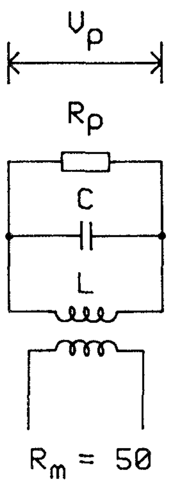


Fig 4 Loop Q by lightly coupled measurement



$Rp = 40k$

$L(uH) = Zo/4fo(MHz)$

$$\frac{D}{Dm} \approx \left[\frac{Rp}{Rm}\right]^{0.25} \approx 5$$

(But smaller for thin
wire matching loop)

D = Loop diameter
d = Tubing diameter
Dm = Matching loop diameter

$$Zo = 276\log_{10}(D/d)$$

$$\text{3dB Bandwidth} = Q = \frac{Rp}{2\pi fL}$$

Fig 5 Simple Design Model for Loop Antennas