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DESIGN OF A PORTABLE MEASURING SYSTEM CAPABLE OF QUANTIFYING THE LF AND HF SPECTRAL EMISSIONS FROM TELECOMMUNICATIONS TRANSMISSION NETWORKS AT FIELD STRENGTHS OF 1 µV/METRE AND BELOW

Report prepared for
the Radiocommunications Agency

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1. Purpose of this document

This document is the draft final report (Deliverable 4) under a contract awarded to the University of Hertfordshire by the Radiocommunications Agency, Ref. AY3430 (510002080). The contract was for the development of a portable measuring system for LF and HF field strength measurements. This report incorporates the contents of Interim Report No. 1 which described the theoretical design of the antennas and Interim Report No. 2 which described the practical design.

The solution adopted is a set of measuring loop antennas with integral pre-amplifiers for use with a conventional EMC measuring receiver or spectrum analyser. In order to cover the required frequency range of 100 kHz - 30 MHz, which is a 300:1 ratio, it is necessary to use three separate antennas.

The target sensitivity was to be able to measure field strengths of 1μV/m (0 dB(μV/m)) or less in 9 kHz bandwidth over a frequency range of 100 kHz - 30 MHz. The Invitation To Tender (ITT) acknowledged that such an ideal was unlikely to be achievable in practice and that the design should aim to meet these objectives as is feasible and acceptable to the Project Manager. Some compromises were necessary as follows:

(A) The system measures the 'H' field strength with a noise floor of approximately -51.5 dB(μA/m) from 400 kHz - 30 MHz. This is equivalent to an 'E' field strength of 0 dB(μV/m) under far field conditions, i.e. where E/H = 120π. This level is achieved with 9 kHz receiver bandwidth and a CISPR 16 average responding detector (the ITT did not specify the type of detector). Thus the minimum measurable signal would be approximately 4 dB higher than the noise floor at -47.5 dB(μA/m) or +4 dB(μV/m) with average detection and -43.5 dB(μA/m) or +8 dB(μV/m) with Quasi-Peak (QP) detection.

(B) Below 400 kHz, the sensitivity is reduced for reasons explained in 3.5 below. The sensitivity is 6 dB lower at 200 kHz and 12 dB lower at 100kHz.

Development work has shown that the above design goals can be met or exceeded. In particular, a lower noise floor has been achieved in practice.

2. Possible types of antenna

2.1 Electrically short dipoles

In principle, an electrically short dipole or monopole could be used. Ref. [1] shows antenna factors for short dipoles for R_L = 50 Ω. This shows for example, that at 3 MHz (λ = 100 m), a dipole with a length *l* of 1m and a diameter *d* of 33 mm has an antenna factor of 53 dB with a 50 Ω load. This would increase to 73 dB at 300 kHz. As shown in 3.7 below, an antenna factor of approximately 20 dB with a 50 Ω load is required to meet the sensitivity requirement.

At frequencies below 30 MHz, a 1 m long dipole is electrically short and can be modelled as an ideal voltage source of E/2 (where E is the electric field strength in V/m) in series with an antenna capacitance C_a which would be of the order of 5 pF in the above example. Kanda [2] shows that a constant antenna factor can be achieved with a short dipole by connecting it to a capacitive load C_L. A practical value of C_L would be 4 C_a which would result in an antenna factor of approximately 20 dB in the above example but would require a buffer amplifier to match into a 50 Ω load. The antenna factor would be constant down to the frequency at which the reactance of C_L equals the amplifier input resistance R_{in}. For example, R_{in} = 20 kΩ would result in an antenna factor which is constant down to 400 kHz. Below that frequency, the sensitivity rolls off at 20 dB/decade.

The thermal noise power kTB in 9 kHz bandwidth at a temperature of 290 K is $1.38 \times 10^{-23} \times 290 \times 9000 = 3.6 \times 10^{-17}$ W. Hence for R = 20 kΩ the r.m.s. noise voltage $\sqrt{(4kTBR)}$ is 1.69 µV e.m.f. or +4.6 dB(µV). If this appears at the input of an ideal noiseless unity gain buffer amplifier, the same noise voltage appears at the 50 Ω output of the amplifier. An antenna factor of 20 dB can be achieved but the noise floor referred to input field strength is +24.6 dB(µV/m) which falls far short of the requirement.

There are also two further problems with this technique. First, the buffer amplifier is untuned and is therefore susceptible to overload by strong signals at any frequency within its passband. Secondly, due to the high input impedance and low load capacitance, the measuring antenna is sensitive to the proximity of personnel, conductive objects and the ground in a typical working environment. Such ground proximity effects were observed with a portable field strength meter which used an electrically short balanced dipole with a pre-amplifier with R_{in} = 10 kΩ. It was found that the indicated field strengths at MF varied significantly with distance above ground.

In order to achieve the required sensitivity using an electrically short dipole, it would be necessary to achieve the required antenna factor of approximately 20 dB by matching to a 50 Ω load with a passive network to avoid the noise introduced by a buffer amplifier with high input impedance. This requires complex conjugate matching where the source capacitance is resonated with an inductor. An inductance of 5 mH would be required to resonate with 5 pF at 1 MHz for example. The inductor would also require a very low self capacitance and its inductance would need to be variable if a constant antenna factor is to be achieved over a wide range of frequencies. Although end loading a short dipole with large 'capacitive hats' would increase the capacitance, allowing a smaller inductance to be used, short dipoles were not considered to be a practical solution.

2.2 Electrically small loops

An electrically small loop responds primarily to the 'H' field component of the incident electromagnetic field. Measurement of 'H' fields below 30 MHz using an electrically small loop is specified in standards such as CISPR 11. It is generally considered to be more satisfactory than 'E' field measurement in the LF/MF/HF bands because 'H' field measurements are less affected by the proximity of conductive objects. It was therefore decided that electrically small loops should be used.

It can be shown that in order to meet the sensitivity requirement, the loops must be resonant. The loops are tuned manually to give a maximum output at the frequency at which measurements are being made. Relays are used for band switching to make the design compatible with possible future development to provide a digitally controlled remote tuning facility using a motor driven variable capacitor. This remote tuning facility does not form part of the current development project however.

It is not considered feasible to use the conventional design of 'shielded loop' which is normally used for untuned loops due to the additional capacitance between the shield and inner conductor. Furthermore, according to Ref. [3], the type of antenna normally known as a 'shielded loop' is not truly shielded. The outer 'shield' is actually the antenna and the conductor inside the shield is the inner conductor of the coaxial transmission line leading to the load. The conventional 'shielded loop' design moves the effective feed point to the top of the loop and allows significant 'E' field rejection to be achieved without the use of a balun.

For the resonant loops proposed, the 'E' field response is minimised by making the loops electrically balanced with balanced loads and by ensuring that they are 'electrically small'. The design target is for the ratio of electric dipole current to magnetic dipole current to be such that the response is primarily to the 'H' field component with at least 20 dB less sensitivity to the 'E' field under 'far field' conditions where $E/H = 120\pi$.

3. Theoretical design of the loop antennas

3.1 Required sensitivity

As shown in 2.1 above, the thermal noise power kTB in 9 kHz bandwidth at 290 K is 3.6 x 10^{-17} W or -134.4 dBm. For a 50 Ω resistive source at room temperature, the r.m.s. noise voltage, √(4kTBR) is 84.9 nV or -21.4 dB(μV) e.m.f. If connected to the input of an ideal noiseless amplifier with a 50 Ω input resistance, the noise p.d. would be -27.4 dB(μV). As shown in 3.6 below, a noise floor of -22 dB(μV) with average detection and -16 dB(μV) with quasi-peak (QP) can easily be achieved. Thus to achieve a noise floor of 0 dB(μV/m) field strength with average detection, an antenna factor of 22 dB is required.

It is intended that a low noise OEM pre-amplifier module will be used. It would be possible to achieve a further improvement by using a pre-amplifier whose noise figure improves when driven from the relatively high source impedance of the loop antennas.

3.2 Theory of operation

Kanda [1] gives the following equivalent circuit for an electrically small loop antenna.

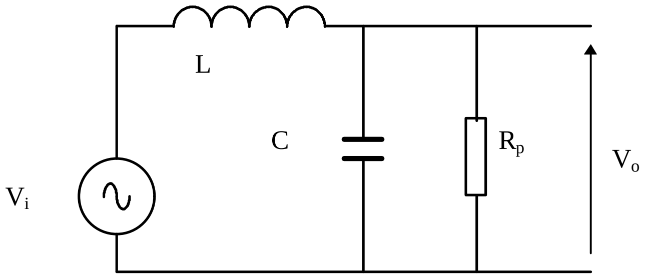

Fig 1. Equivalent circuit for an electrically small loop antenna

L is the self-inductance of the loop, C is the total capacitance in parallel with the loop and R_p is the parallel load resistance. V_i, the voltage induced in the loop by the incident electromagnetic wave is determined by the rate of change of magnetic flux and is given by:

$$V_i = j\omega\mu H_i NS \tag{1}$$

where μ is the permeability of the loop core, H_i is the component of the magnetic field normal to the plane of the loop, ω is the angular frequency of H_i, N is the number of turns and S is the area in m^2. Kanda states that at angular frequency ω,

$$\frac{V_o}{V_i} = \frac{-j\frac{1}{\delta}}{\frac{1}{Q} + j(\delta - \frac{1}{\delta})} \tag{2}$$

Where

$$Q = \frac{R_p}{X_0} \qquad X_0 = j\omega_0 L \qquad \delta = \frac{\omega}{\omega_0} \qquad \omega_0 = \frac{1}{\sqrt{LC}} \tag{3}$$

If $\omega = \omega_0$, i.e. when the loop is at resonance, $\delta = 1$ and (2) reduces to:

$$\frac{V_o}{V_i} = -jQ \qquad (4)$$

Combining (1) and (4) yields the following transfer function at resonance:

$$\frac{V_o}{H_i} = \omega\mu NSQ \qquad (5)$$

If far field conditions are assumed where E/H = 120π, $E_i/120\pi$ can be substituted for H_i so that:

$$\frac{V_o}{E_i} = \frac{\omega\mu NSQ}{120\pi}$$

Substituting $\mu = \mu_0 = 4\pi \times 10^{-7}$ and taking the reciprocal yields the following expression for the inverse transfer function E_i/V_o:

$$\frac{E_i}{V_o} = \frac{120\pi}{\omega.4\pi \times 10^{-7}.NSQ} = \frac{3 \times 10^8}{\omega NSQ} \qquad (6)$$

It is also necessary to maintain a constant value of E_i/V_o over a range of frequencies. One way to achieve this is to make the load resistance R_L much less than the inductive reactance X_L at all frequencies of interest, i.e. $Q<<1$. Hence the increase of V_i with frequency is compensated by the increasing reactance X_L. This is known as a short-circuit current loop but does not yield sufficient sensitivity for this application.

An alternative approach is to use a 'Q' substantially greater than 1 and to maintain a constant ωQ product, i.e. to make Q inversely proportional to frequency. Increasing Q as frequency decreases compensates for the decreasing rate of change of magnetic flux at lower frequencies. This can be achieved if C in Fig 1 is variable and R is constant and large compared to X_L. As $Q = R_p/X_L$ and X_L is directly proportional to frequency, Q is inversely proportional to frequency.

Although Kanda's analysis is in the context of 'Q' factors less than 1, it can be applied to larger 'Q' factors. The maximum loaded 'Q' factor which can be achieved in practice is limited by series losses. These are ohmic losses which swamp the radiation resistance of the loops. The radiation resistance of the loops ranges from 2.5×10^{-2} Ω at 30 MHz to 6×10^{-9} Ω at 100 kHz. The series losses in the resonant circuit can be represented by an equivalent parallel load resistor connected in parallel with R_p in Fig. 1.

Equation (6) only represents the antenna factor if $R_p = 50$ Ω. As L is determined by the geometry of the loop and Q is determined by the required antenna factor, making $R_p = 50$ Ω will not generally give the required 'Q'. Hence, it is necessary to match the parallel load resistance R_p to the required load resistance R_L (e.g. 50 Ω) using a transformer with N:1 turns ratio as shown in Fig. 2. The transformed load N^2R_L therefore defines the loaded 'Q'. A consequence of this is that the antenna factor depends on R_L and if the loaded 'Q' is small compared to the unloaded 'Q', R_L is driven by a constant current source.

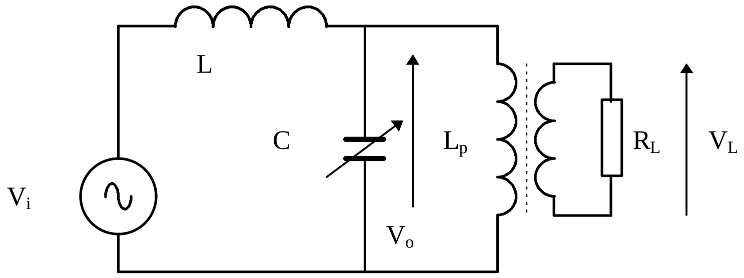

Fig. 2 Equivalent circuit of an electrically small loop with transformer matching.

If the primary inductance L_p is large compared to L so that primary magnetising current is insignificant, then:

$$\frac{V_o}{V_L} = \sqrt{\frac{R_p}{R_L}} \qquad (7)$$

where V_L is the p.d. across load R_L. Hence by substituting for V_o, the 'E' field antenna factor for a load R_L is given by:

$$\frac{E_i}{V_L} = \frac{3 \times 10^8}{\omega NSQ} \sqrt{\frac{R_p}{R_L}} \qquad (8)$$

R_p can be eliminated by substituting $R_p = Q\omega L$

$$\frac{E_i}{V_L} = \frac{3 \times 10^8}{\omega NSQ} \sqrt{\frac{Q\omega L}{R_L}} = \frac{3 \times 10^8}{S} \sqrt{\frac{Q\omega L}{N^2 \omega^2 Q^2 R_L}} = \frac{3 \times 10^8}{S} \sqrt{\frac{L}{N^2 \omega Q R_L}} \qquad (9)$$

It is required to maximise V_L for a given value of E_i, hence the ratio E_i/V_L should be as small as possible. The inductance of a single turn loop is determined by the loop geometry and wire diameter. The inductance of an N turn loop is directly proportional to N^2, hence L/N^2 is constant. This shows that at a given frequency, changing the number of turns does not change the antenna factor nor the required Q. Changing the number of turns does not change the amount of power captured by the loop. Nevertheless, it is desirable to minimise the inductance of each turn in order to reduce the required Q.

3.3 Loop parameters

E_i/V_o is required to be of the order of 10 to achieve the required sensitivity. The design is subject to the following constraints:

(A) The antenna factor should be substantially constant with frequency.

(B) The largest practicable size of loop for portable use is considered to be approximately 1 m square.

(C) The total conductor length should not exceed approximately 0.1 wavelength at the highest frequency of operation, for two reasons. First, to ensure that the loop is 'electrically small', i.e. the current is in the same phase all round the loop and the response is primarily to the 'H' field. Secondly, the self-resonant frequency (SRF) should be significantly higher than the highest frequency of operation to minimise proximity effects which occur near SRF where the loop resonates with stray capacitance.

(D) Tuning should be by means of variable capacitors which are commercially available.

(E) The unloaded 'Q' factor should be significantly higher than the loaded 'Q' but must be achievable in practice.

(F) The loaded 'Q' should be such that the -3 dB bandwidth of the loop is always significantly greater than the 9 kHz -6 dB bandwidth of the measuring receiver.

(G) The impedance transformation ratios for matching to 50 Ω should be achievable in practice with acceptably low loss.

The required frequency range of 100 kHz - 30 MHz is a 300:1 frequency ratio. The requirement can be met using a set of three measuring antennas designated 'A', 'B' and 'C'. Four switched frequency ranges are required for loop 'A' and two for loop 'B'. Two options for loop 'A' were offered in the tender, a 0.6 m loop or a 1m loop. The RA Project Manager selected the 1 m option which offers a higher sensitivity than the 0.6 m option from 100 kHz - 1.76 MHz.

The design characteristics of the three loops are summarised in Table 1.

Loop:	A	B	C
Dimensions:	1 m x 1 m	0.6 m x 0.6 m	0.3 m x 0.3 m
No. of turns:	4	1	1
Inductance:	65 µH	2.6 µH	1.3 µH
Maximum tuning capacitance:	730 pF	730 pF	182 pF
Conductor length:	16 m (0.094 λ at f_{max})	2.4 m (0.096 λ at f_{max})	1.2 m (0.12 λ at f_{max})
Frequency ranges:	100 kHz - 220 kHz 200 kHz - 440 kHz 400 kHz - 880 kHz 800 kHz - 1.76 MHz	1.6 - 5.2 MHz 4.5 - 12 MHz	11 - 30 MHz
Self-resonant frequency:	>2 MHz	>18 MHz	>45 MHz
Minimum bandwidth: (See also Fig 3)	15 kHz.	45 kHz.	275 kHz.
Loaded Q (max):	30	33	40
Parallel load:	4000 Ω above 400 kHz	800 Ω	3200 Ω
Series load:	6 Ω below 400 kHz		

Table 1. Summary of loop characteristics.

3.4 Loop tuning

In principle it would be possible to cover the lowest frequency range, 100 kHz - 200 kHz using a 32 turn 4 mH loop tuned directly with a variable capacitor of approximately 600 pF. In practice however, the stray capacitance of the winding would restrict the tuning range unduly unless the turns were widely spaced. There is also a need to cover the whole 100 kHz - 1.76 MHz range with the same loop. In principle, it would be possible to switch turns in series or parallel to make a 4, 8, 16 or 32 turn loop. In practice however the stray capacitance would result in a self-resonant frequency well below the maximum frequency required.

A 4 turn loop 1 m square is found to have an SRF of over 2 MHz provided adjacent turns are spaced 10 mm apart. To ensure that the interwinding capacitance has a low dielectric loss, the cable insulation should be Low Density Polyethylene (LDPE) rather than PVC. In principle, switched series loading coils could be used to tune a 4 turn loop to resonance over the frequency range 100 kHz - 1.76 MHz with a variable capacitor of 600 - 700 pF. It can be shown however that if the total inductance is N times the loop inductance, the 'Q' must be increased by a factor of N^2 in order to achieve the required antenna factor. Similarly, if a parallel loading coil is used to reduce the inductance of a loop by a factor of N to achieve resonance on a higher frequency, the required 'Q' is also increased by a factor of N^2.

The solution adopted is to use a 4 turn loop, 1 m square with a step-up transformer. The loop can then be resonated from 100 kHz upwards using a dual gang 365 pF variable capacitor with both gangs connected in parallel. For 100 - 220 kHz, a 1:8.6 step up transformer is used between the loop and the variable capacitor. This effectively transforms a 4 turn loop into a 34.4 turn loop. The capacitor resonates with an inductor whose effective inductance is transformed up by a factor of 8.6^2 from approximately 65 μH at the loop terminals to 4.6 mH at the capacitor terminals. An alternative view is that the loop resonates with a capacitor whose effective capacitance is transformed up by a factor of 8.6^2. The turns ratio is switched between 1:8.6, 1:4.3, 1:2.15 or 1:1.08 to select one of four ranges. It has been shown that the required ratios and 'Q' factors can be realised in practice using two separate ferrite cored transformers, one for 100 kHz - 440 kHz and one for 400 kHz - 1.76 MHz.

For loop 'B', a step-up autotransformer wound on a ferrite ring core is used to resonate the single turn loop on the 1.6 - 5.2 MHz range using a 3:1 turns ratio. On the 4.5 - 12 MHz range, the variable capacitor is connected directly across the loop.

Loop 'C' has a single range covering 11 - 30 MHz. A dual gang 365 pF variable capacitor is used with the two gangs in series. With this arrangement, the wiping contact is not in series with the resonant which ensures a high unloaded 'Q' for loop 'C' at frequencies where the effect of series resistance of the wiping contact would be more significant than for loops 'A' or 'B'.

3.5 Loop loading and bandwidth.

From 400 kHz - 30 MHz, the loops are loaded with parallel load resistances as listed in Table 1. This results in a substantially constant antenna factor as is shown in Fig 3. The antenna factors shown in Fig 3 are calculated and take account of the effect of finite unloaded 'Q'.

Although a near constant antenna factor could be maintained down to approximately 200 kHz before unloaded 'Q' becomes a limiting factor, the bandwidth at 200 kHz would be 9 kHz or less which is considered too narrow. The loading is therefore changed from parallel to series below 400 kHz. This results in a Q which is directly proportional to frequency so that as frequency is reduced, the bandwidth remains constant but the sensitivity is reduced. As explained in 3.7 below, even if a constant antenna factor could be maintained down to 100 kHz without unduly narrow bandwidth, there would be little need for such high sensitivity due to atmospheric noise levels. The -3 dB bandwidth of the loop antennas and the target loaded 'Q' factors are shown in Fig 4.

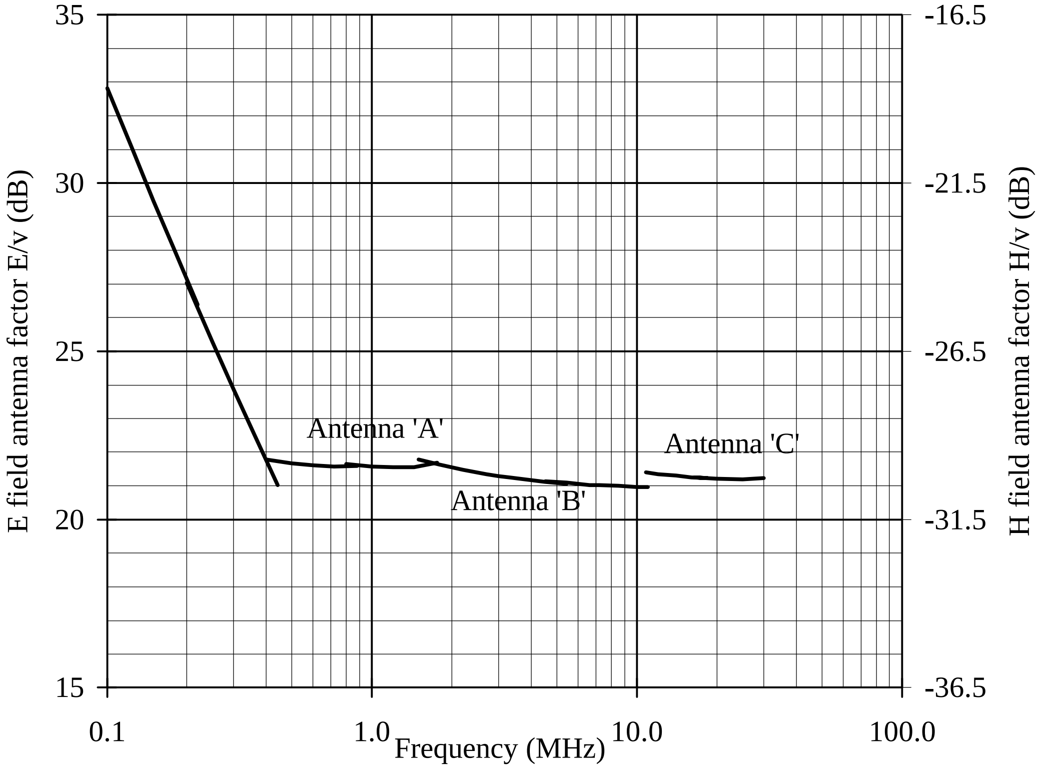

Fig 3. Predicted antenna factors of antennas 'A', 'B' and 'C' in passive mode

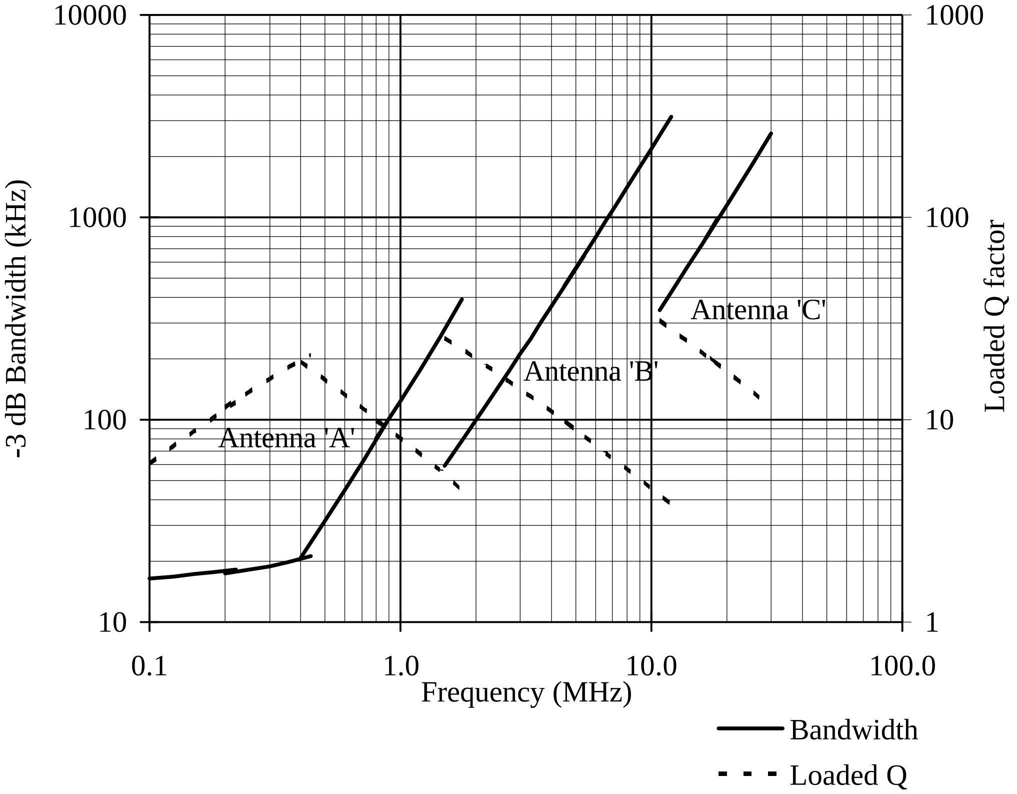

Fig 4. Design bandwidth and loaded 'Q' factors of antennas 'A', 'B' and 'C'

3.6 Pre-amplifier

The outputs of the antennas are matched into 50Ω using passive matching techniques. If a measuring receiver is available with a noise figure of 3 - 4 dB, the antenna factors shown in Fig 3 would allow a noise floor of 0 dB(μV/m) or lower to be achieved without any pre-amplifier, using average detection. As EMC measuring receivers and spectrum analysers typically have a noise figure of 12 dB or more, pre-amplifiers are included for each loop but can be switched out to enable the antenna to be used in passive mode if required.

In view of the short timescale, an OEM broad band pre-amplifier module, Mini Circuits ZFL 500LN was selected for the prototype antennas. The gain of this unit is specified as 24 - 27 dB, depending on supply voltage. The maximum gain and maximum output level are achieved at the maximum specified supply voltage of 15 V. This gain is sufficient to ensure that the overall measuring system noise floor is largely independent of the noise figure of the measuring receiver or spectrum analyser.

Table 2 shows some measurements of noise floor with various pre-amplifiers and measuring instruments. The noise floor referred to the input is based on the nominal gain specified for each pre-amplifier.

Measuring instrument	RF atten.	Pre-amplifier	Input load.	Noise floor dB(μV) (QP)	Noise floor dB(μV) (Average)
R & S ESH3	0 dB	None	O/C	-11	-13.8
HP 8591 EM	0 dB	None	O/C	-4.0	-10.3
HP 8591 EM	0 dB	HP 8447F (28 dB)	50 Ω	-16.7	-23.3
HP 8591 EM	10 dB	ZFL 500LN (27 dB)	50 Ω	-18	-24.3

Table 2. Noise floor measurements

A further reduction in noise floor can be achieved if the HP 8591EM is used with 0 dB RF attenuation and the Mini Circuits ZFL 500 LN pre-amplifier. Furthermore, the ZFL 500 LN exhibits a lower noise figure when the input is driven from an impedance higher than 50 Ω. This should further improve the noise performance above 400 kHz where the source impedance of the loop antennas is substantially higher than 50 Ω and the thermal noise power in the source resistance is not well matched into the amplifier input.

3.7 System noise floor

When using antennas 'A', 'B' and 'C' in active mode with their built-in pre-amplifiers and an EMC measuring receiver or spectrum analyser, the predicted system noise floor expressed as an equivalent field strength is as shown in Fig. 5. This is based on a measuring receiver noise floor of -22 dB(μV) (Average) and -16 dB(μV) (QP). As shown in 3.6 above, it may be possible to achieve a lower noise floor in practice.

Although the system noise floor increases below 400 kHz, the atmospheric noise field strength also increases at lower frequencies. Details of atmospheric noise levels are included in Ref. [4]. There are substantial variations with season, time of day and geographical location. The man-made and atmospheric noise curves in Ref [4] are for a lossless vertical monopole antenna and have been converted to RMS field strength in 9 kHz bandwidth using equation (7) in Ref [4].

Fig 5 shows two atmospheric noise levels plotted in terms of median r.m.s. 'E' field strengths in 9 kHz bandwidth.. The 50% confidence curve is comparable to night time atmospheric noise levels in the UK whereas the 20% confidence curve is comparable to UK atmospheric noise levels at 08.00 - 12.00 UTC.

At 200 kHz for example, the atmospheric noise level does not exceed +7 dB(µV/m) with 20% confidence and does not exceed +21 dB(µV/m) with 50% confidence. Hence even if a measuring system noise floor of 0 dB(µV/m) could be maintained from 400 kHz down to 100 kHz, this would be below the background atmospheric noise level for much of the time.

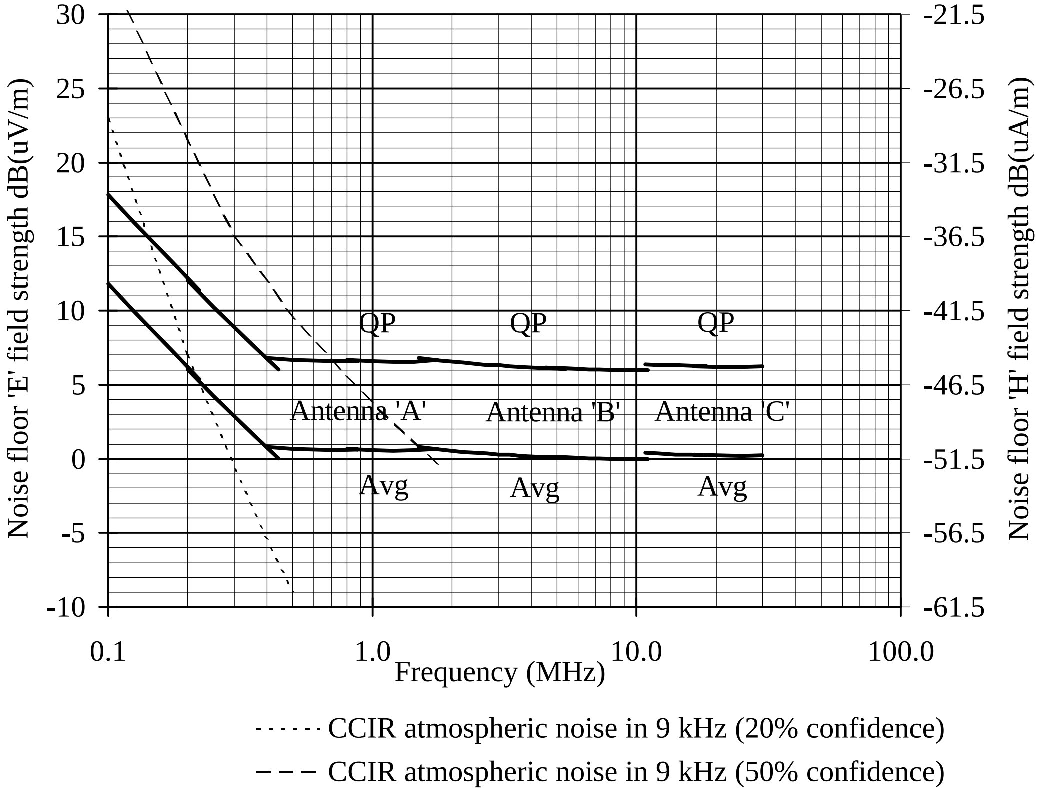

Fig 5. Noise floor 'E' and 'H' field strength for loops 'A', 'B' and 'C' in active mode (QP and average detection)

3.8 Dynamic range

The output 1 dB compression point of the pre-amplifier is at least +5 dBm or +112 dB(µV). If the gain is 27 dB, this corresponds to an input level of 85 dB(µV) or a field strength of approximately 106 dB(µV/m) from 400 kHz - 30 MHz (increasing below 400 kHz). An overload warning facility is provided with an adjustable threshold set at around -5 dBm at the pre-amplifier output which corresponds to a field strength of 95 dB(µV/m).

As the tuned loop acts as a passive pre-selector, it gives additional rejection of unwanted signals outside the loop bandwidth. This can improve the dynamic range significantly, depending on the spacing between the wanted and unwanted signals and the loaded 'Q' factor of the loop at the measurement frequency (see also Fig 4). If a measurement requirement arises where the dynamic range of the pre-amplifier is a limiting factor, the antennas can be used in passive mode where the pre-amplifier is bypassed.

It should be noted however that the pre-amplifier is specified to withstand a maximum input of +5 dBm. If any loop is tuned to resonance at the frequency of a signal whose field strength exceeds 133 dB(µV/m), the pre-amplifier should not be switched in.

3.9 Calibration accuracy

The source impedance of the antennas in passive mode is substantially higher than 50 Ω at frequencies above 400 kHz (and substantially lower than 50 Ω below 400 kHz). Hence the antenna factor in passive mode is affected by the external load impedance which should be 50 Ω resistive. A target figure of +/-3 dB is specified for combined calibration accuracy and repeatability. To improve calibration accuracy, the antennas should be calibrated in active mode (with built-in pre-amplifiers in use).

Range selection is by means of sealed relays to ensure consistently low contact resistance and long operating life.

In order to improve repeatability of measurements, a slow motion drive is fitted to the tuning control to allow fine adjustment of the resonant frequency of the loop at frequencies where the 'Q' factor is relatively high.

5 Mechanical construction

5.1 General description

The loop antennas are designed for field use and are of robust mechanical construction and reasonably weatherproof. Fig 8 shows the principle of construction of all three loops but is not to scale. Fig 9 shows a photograph of the finished loop antennas.

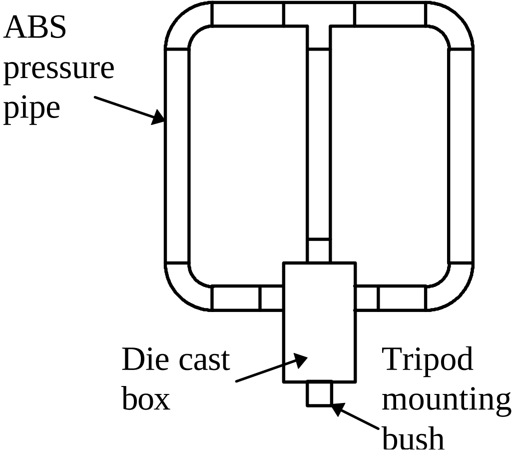

Fig 8. Principle of mechanical construction of loops 'A', 'B' and 'C'.

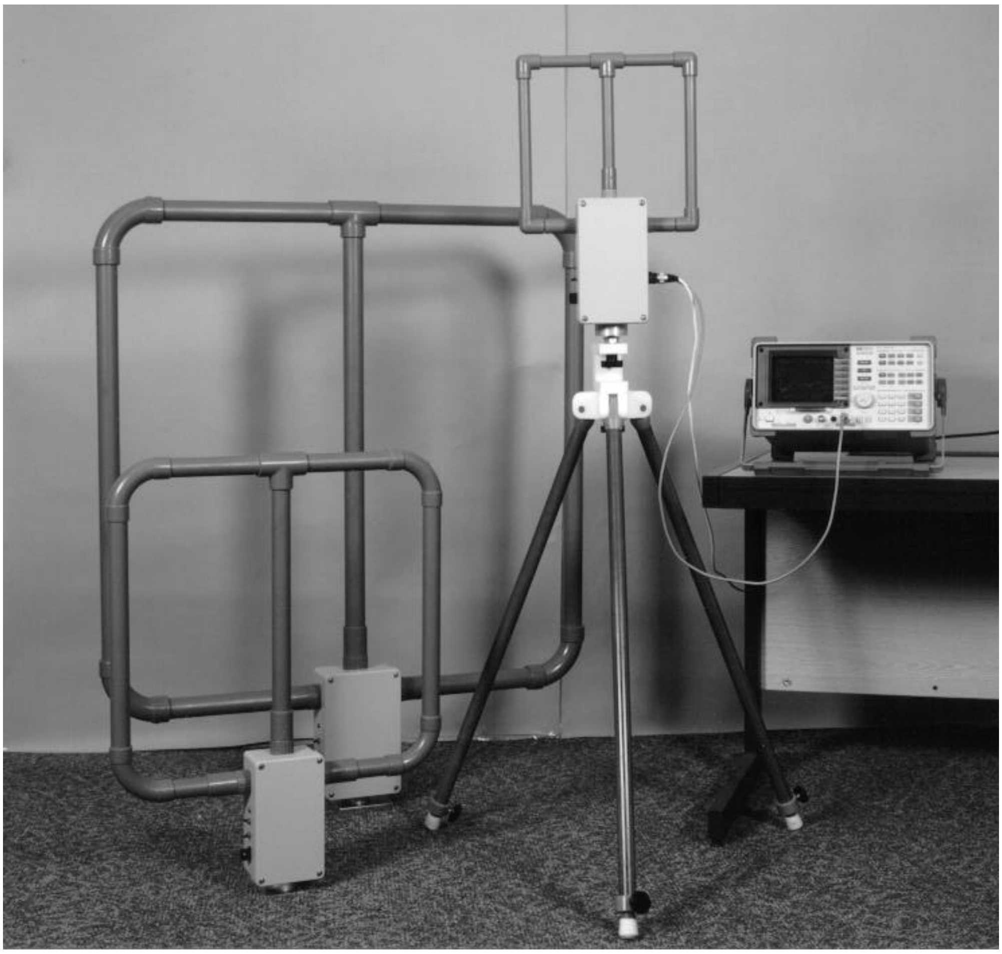

Fig 9 Photograph of loops 'A', 'B' and 'C'

The circuitry is housed in Rolec heavy duty pre-painted aluminium die cast boxes. These have sealing gaskets in the lids which are designed so that they make good electrical contact with the boxes at many points. Loop 'A' is housed in a 266 x 166 x 100 mm box. Loops 'B' and 'C' are housed in a 225 x 126 x 90 mm box.

The loop cables are contained in a square loop of heavy duty self supporting ABS pressure pipe, BS5391, Class 'E'. ABS has been selected in preference to PVC-U as the ABS type has better properties at low temperatures and is less likely to become brittle if used outdoors in very cold weather.

The central vertical tube provides additional support and prevents the ends of the loop tubing from rotating where they join the box. Standard pipe fittings such as 90° bends and tee pieces are used and the whole ABS pipe assembly is fixed together using ABS cement. Loop 'A' uses 1.25 inch inside diameter pipe, loop 'B' uses 1 inch i/d and loop 'C' uses 0.5 inch i/d.

5.2 Tube lengths

Fig 10 and Table 4 show the cutting details for the lengths of tubing

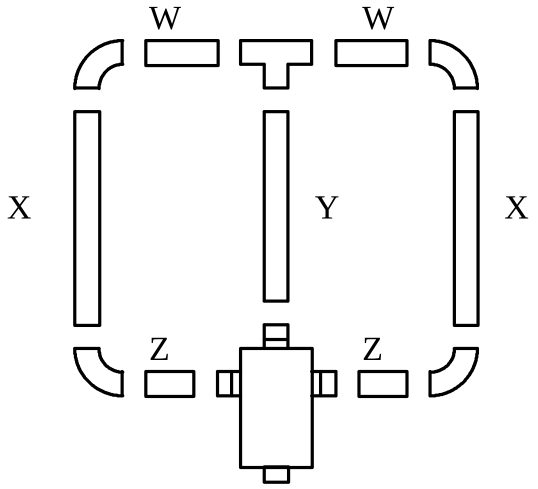

Fig 10. Tube reference letters

Tube reference letter	Loop 'A' tube length (mm)	Loop 'B' tube length (mm)	Loop 'C' tube length (mm)
W (2 off)	398	200	125.5
X (2 off)	839	560	272
Y (1 off)	864	443	200
Z (2 off)	279	106	38

Table 4. Tube lengths for Loops 'A', 'B' and 'C'.

5.3 Tube end fittings

The fittings used to join the tubing to the box are shown in Fig 10. Item 'Q' is a plain male spigot to BSP female adaptor. Item 'P' is a straight coupler. Item 'R' is a PVC-U threaded male taper plug. This item does not appear to be available in ABS. A hole is drilled in the two plugs 'R' at the ends of tubes 'Z' but is not required at the end of tube 'Y'. To avoid an excessively tight fit, it is necessary to file the tapered thread on item 'R' to a parallel thread and then restore the thread profile.

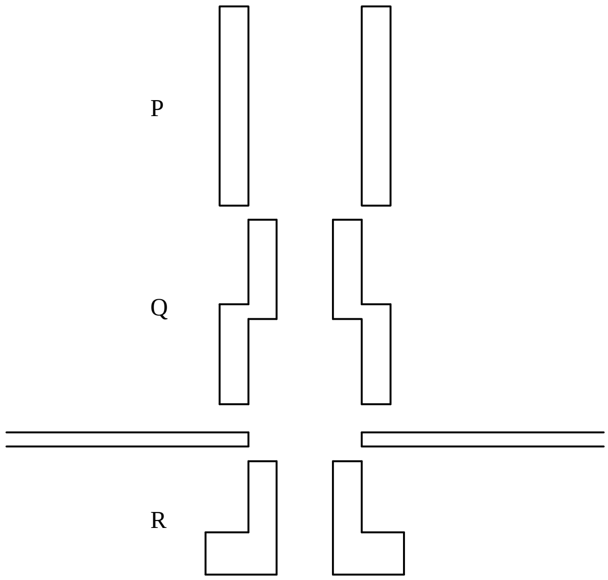

Fig 11. Detail of fittings used to join tubing to box

An alternative method of fitting the tubing to the box would be to use a plain female socket to threaded taper spigot instead of items 'P' and 'Q' in Fig 11. A metal BSP back nut would then be required inside the box. The outside diameter of the adaptor may be too large to clear the flange of the box for Loops 'A' and 'B' however and the taper thread may need to be filed to avoid an excessively tight fit.

5.4 Order of assembly

It is not practicable to make a 'dry run' trial assembly without solvent weld cement because the tubing is a very tight fit into the fittings without cement but can be fitted easily with cement. The recommended order of assembly is as follows (See Figs 10 and 12):

Fit the adaptors to the ends of tubes 'Y' and 'Z'.
Join tubes 'W' and 'Y' with a 'T' piece and fit tube 'Y' to the box.
Thread the cable(s) through tubes 'W'.
Thread the top two 90 degree bends onto the cable but do not join them to tubes 'W' yet.
Thread the cable(s) through tubes 'X'.
Thread the lower two 90 degree bends onto the cable then thread tubes 'Z' onto the cable.
Join tubes 'X' and 'Z' with the lower two 90 degree bends. The sub-assemblies should now be as shown in Fig 12.
Fit the 90 degree bends to the ends of tubes 'W', noting that that due to the taper of the box, tubes 'X' are not exactly parallel with tube 'Y'.
Join tubes 'W' and 'X'
Fit tubes 'Z' to the box.

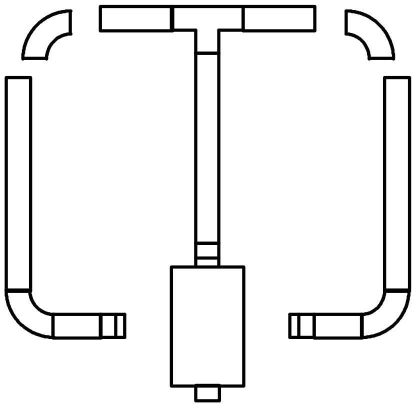

Fig 12. Order of assembly

5.5 Tripod mounting bush

The tripod mounting bush consists of a 20 mm length of 2 inch diameter aluminium alloy rod. This is drilled in the centre and tapped with a 5/8 inch UNC thread to fit the type of tripod normally used for supporting EMC measuring antennas. In practice however, it has been found that a larger diameter bush would be more suitable for fitting to certain types of tripod.

The upper face of the mounting bushes is machined at an angle of 2.5° to keep the central support tube 'Y' vertical when the loop is rotated on the tripod. This angle is correct for Loop 'A' but should be reduced for Loops 'B' and 'C' because the smaller die cast boxes have a smaller taper angle.

5.6 Variable capacitor reduction drive

A reduction drive is provided for the tuning capacitor. The preferred type is a Jackson type 4511F epicyclic ball drive with 6:1 reduction ratio but these were unobtainable when the prototype antennas were constructed. A 35mm vernier dial was used instead. The vernier dial is mounted inside the box as it is not waterproof. Nevertheless, there would be some advantage in making the scale visible to the user so that the setting could be recorded when measurements are made. The control knob is removed from the vernier dial and the shaft is extended by means of a length of 6.35 mm diameter nylon shaft and a shaft coupler. The shaft coupler must be drilled out to 7 mm i/d to fit the stub shaft on the vernier dial. An insulated extension shaft is essential for Loop 'B' where neither side of the variable capacitor is grounded and is useful of the other two loops where the slight flexibility of a nylon shaft assists smooth rotation.

6. Testing

6.1 Tuning range test

A coupling loop approximately 100 mm diameter is placed close to the top of the loop under test. The coupling loop is driven by an RF signal generator, preferably via a balun. The loop under test is tuned to the limits of each range. The frequency of the signal generator is adjusted while monitoring the output from the loop, in order to find the resonant frequency. The measured tuning range of each loop should exceed the design range. The results for the prototype loops are shown in Table 5.

Loop:	A	B	C
Design frequency ranges:	100 kHz - 220 kHz 200 kHz - 440 kHz 400 kHz - 880 kHz 800 kHz - 1.76 MHz	1.6 - 5.2 MHz 4.5 - 12 MHz	11 - 30 MHz
Measured frequency ranges:	<100 kHz - >240 kHz <180 kHz - >450 kHz <375 kHz - > 900 kHz <750 kHz - >1.8 MHz	1.53 - 6.75 MHz 4.37 - >12 MHz (4.37 - 19 MHz without C13)	10.4 - 31 MHz (10.4 - 41.5 MHz without C1/C12)

Table 5. Design and measured tuning ranges of the prototype loop antennas.

6.2 Loaded and unloaded 'Q' test

The loops should be tested to ensure that a sufficiently high unloaded 'Q' factor is achieved at the limits of each frequency range and that the loaded 'Q' values are close to those indicated. To perform this test, the loops should be operated with the pre-amplifiers switched out. For the unloaded 'Q' test, a load resistance in excess of 500Ω is required. For the loaded 'Q' test, an accurate 50Ω termination is required.

A test signal from a signal generator is coupled into the loop under test via a small coupling loop as described in 6.1 above. The signal generator is set to the centre frequencies listed in Table 6. The loop under test is then tuned to resonance while monitoring the output signal from the loop. The signal generator frequency is then adjusted to find the upper and lower -3 dB points and the 'Q' factor is calculated from the -3 dB bandwidth and centre frequency. The typical loaded and unloaded 'Q' factors achieved on the prototypes were as follows:

Loop	Range	Nominal test frequency	Unloaded 'Q'	Loaded 'Q'
A	1	100 kHz	58	11
	1	200 kHz	65	20
A	2	200 kHz	74	20
	2	400 kHz	69	32
A	3	400 kHz	87	33
	3	800 kHz	49	21
A	4	800 kHz	50	21
	4	1600 kHz	36	9.8

Loop	Range	Nominal test frequency	Unloaded 'Q'	Loaded 'Q'
B	1	1.5 MHz	110	38.5
	1	3 MHz	104	19.2
	1	4.5 MHz	89	15
B	2	4.5 MHz	167	14.2
	2	9 MHz	103	7.1
	2	12 MHz	64	4.8
C		12 MHz	166	50
		18 MHz	140	47
		30 MHz	57	26

Table 6. Loaded and unloaded 'Q' factors measured for the three loops.

The unloaded 'Q' values on subsequent sets of antennas may be higher than the values listed in Table 6 but if they are more than 20% lower, it is recommended that the cause should be investigated, particularly at frequencies where the unloaded 'Q' values are less than three times the loaded 'Q'. The loaded 'Q' values should be within +/-15% of the values shown in Table 6 when the loops are operated in passive mode with an accurate 50Ω resistive load.

6.3 Electrical balance test

This test uses capacitive coupling of a test signal onto the loop conductor to find the electrical centre. If the electrical centre coincides with the physical centre, this indicates optimum electrical balance and hence maximum rejection of vertically polarised 'E' fields.

A length of coaxial cable with a 50Ω through termination is fitted with a wire probe approximately 100 mm long. The braid of the cable is temporarily grounded to the box as shown in Fig 13. The cable is connected to an RF signal generator set to a test frequency near the maximum frequency of each loop. The loop is tuned to resonance by monitoring its output using a spectrum analyser. The loop pre-amplifier may be switched on if required. The probe is held in contact with the tubing of the loop using an insulated tool and is moved from side to side as shown in Fig 13. The loop output level should pass through a minimum when the probe passes the electrical centre of the loop. If no minimum is found or it is far from the physical centre of the loop, the cause of the unbalance should be investigated.

For Loop 'A', the electrical centre of the loop was close to the physical centre. There was a small unbalance on Loops 'B' and 'C' but this is not considered significant. For Loop 'B', the electrical centre was offset by approximately 100 mm. This could be corrected by adjusting C13 which could be changed to a 2 - 22 pF trimmer capacitor instead of a fixed capacitor. For Loop 'C', the electrical centre of the loop was offset from the physical centre by approximately 25 mm. This is due to the loop connector not being exactly in the centre of the box and could be improved by a modification to the PCB layout.

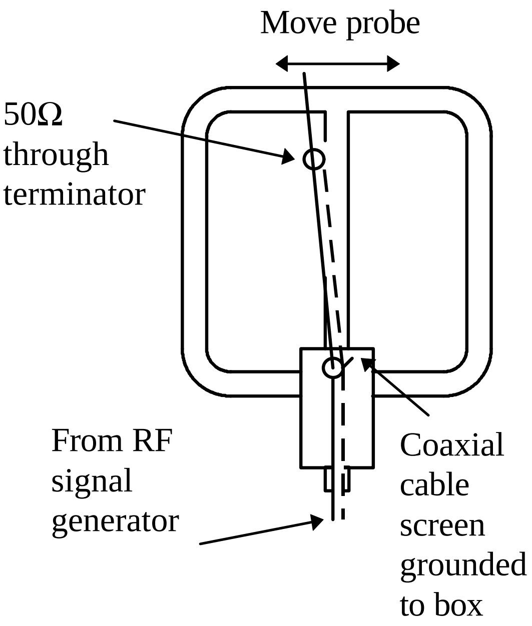

Fig 13. Principle of electrical balance test.

7. Calibration and operation

There are several possible test methods for calibrating a loop antenna at frequencies up to 30 MHz.

7.1 TEM Cell method

A Transverse Electromagnetic (TEM) Cell is used to generate a defined field. The field within the usable volume of a TEM cell approximates to a plane wave where the wave impedance E/H = 120π. This technique is used by the National Physical Laboratory (NPL) but their TEM cell can only accommodate a loop antenna whose overall height does not exceed approximately 700 mm. This method is therefore unsuitable for calibrating Loop 'A'. It may be possible to calibrate Loop 'A' in a larger TEM cell if available. The maximum operating frequency of any TEM cell is limited by resonances and higher order modes which propagate above a certain frequency cut-off frequency. A TEM cell large enough to accommodate Loop 'A' may have a maximum operating frequency below 30 MHz but this would still be far above the maximum operating frequency of Loop 'A'.

Another factor to be considered for calibration of an unshielded loop in a TEM cell is the response to the 'E'-field although this may not be significant for loops with good electrical balance.

7.2 Standard field method

This method, which is described by Kanda [2], consists of generating a field whose magnitude can be calculated in terms of the current flowing in a transmitting antenna and its dimensions. Kanda shows a small single turn balanced transmitting loop of 10 cm radius. The receiving loop being calibrated is positioned at a distance *d* from the receiving loop, on the same axis. The normal component of the magnetic field from the transmitting loop averaged over the area of the receiving loop is calculated using the following equation:

$$H = \frac{\mathrm{b} I r_1}{r_2} \sum_{m=0}^{\infty} \frac{1}{(2m+1)!} \cdot \frac{1 \cdot 3 \cdots (2m+1)}{2 \cdot 4 \cdots (2m+2)} \left[\frac{\mathrm{b} r_1 r_2}{R_0} \right] \cdot h_{2m+1}^{(2)}(\mathrm{b} R_0)^{\cdot}$$

Where:

H	r.m.s. value of the magnetic field, A/m
I	r.m.s. current in the transmitting loop, A
r_1	radius of the transmitting loop, m
r_2	radius of the receiving loop, m
R_0	$= \sqrt{(d^2 + r_1^2 + r_2^2)}$
d	axial distance between the two loops, m
b	= 2π/λ m^{-1}
l	free space wavelength, m
$h_n^{(2)}$	nth-order spherical Hankel function of the second kind

The above equation only applies to circular loops however.

7.3 Mutual Induction method

This is the method normally used by Schaffner-Chase EMC Ltd for calibrating conventional untuned H-field loops. A small radiating loop is placed at the centre of the loop under test. This is effectively a standard field method but differs from the method described by Kanda [2] because the two loops are in the same plane. The mutual inductance between the two loops is calculated and this result is used to predict the relationship between current in the radiating loop and the output of the loop under test. The result of this test method is not traceable to national standards because it involves a calculation of the mutual inductance. Nevertheless, it has been shown that for an untuned circular loop, the results agree closely with the results obtained by calibration in a TEM cell by NPL.

Two issues need to be investigated if this test method is to be used for resonant loops. First, the calculation of mutual inductance needs to be applied to a square loop instead of a circular loop. Secondly, it needs to be established whether this test method is valid for a resonant loop with a 'Q' factor of up to 50.

7.4 Standard antenna method

In the standard antenna method, the field generated by the radiating loop is not calculated but is measured by means of an antenna whose response can be calculated.

This method was used for an approximate calibration at University of Hertfordshire and it is understood that a similar method will be used for the final calibration by Schaffner-Chase EMC Ltd. A test signal is radiated by an antenna and the field strength is measured at a given distance using a calibrated antenna. The antenna under test is then substituted for the calibrated antenna.

For the tests at University of Hertfordshire, the radiating antenna was a loop 300 mm square driven by an RF signal generator. The radiation resistance of such a loop varies from $3 \times 10^{-12}\ \Omega$ at 100 kHz to $2.5 \times 10^{-3}\ \Omega$ at 30 MHz so the effective radiated power is extremely low. Consequently, it was necessary to use a spectrum analyser bandwidth of 200 Hz in order to detect the signal with an adequate signal to noise ratio at a distance of 10m. Measurements could only be performed on frequencies which were relatively free of ambient signals.

As a ground plane was not available, care was required to avoid misleading results due to conducted signals via mains cables and other interconnecting cables. Ferrite ring common mode chokes were fitted to various cables and the radiating and receiving equipment were powered from separate mains supplies.

The final calibration results are in Appendix E (to be added).

7.5 Tuning aid

It was found that when measuring modulated signals or signals close to the noise floor, some sort of test signal is useful in order to tune the loops accurately to resonance. A simple battery powered comb generator was designed with a 3.58 MHz crystal oscillator and divide by 4096. An output buffer drives a small square radiating loop 115 mm square made of 19 mm wide aluminium strip. The schematic diagram and a photograph are shown in Appendix D.

9. Conclusion

A portable measuring system for LF and HF field strength measurements has been developed. This comprises a set of three measuring loop antennas with integral pre-amplifiers for use with a conventional EMC measuring receiver or spectrum analyser. The prototype set of loop antennas was designed, constructed and demonstrated to the Radiocommunications Agency within 11 weeks of the contract start date.

The performance of the loop antennas is better than the design goals specified in the first Interim Report. The system noise floor referred to incident field strength is approximately -51.5 dB(μA/m) from 400 kHz - 30 MHz. This is equivalent to an 'E' field strength of 0 dB(μV/m) under far field conditions, i.e. where $E/H = 120\pi$. This level is achieved with 9 kHz receiver bandwidth and a CISPR 16 ***Quasi-Peak*** (QP) detector. Below 400 kHz, it is necessary to reduce the sensitivity for reasons explained in 3.5 above. The sensitivity is 6 dB lower at 200 kHz and 12 dB lower at 100kHz.

10. References

[1] BS727 : 1983. British Standard Specification for Radio-interference measuring apparatus. British Standards Institution 1983. Fig 7, Short Dipole Aerial Factors for $R_L = 50\Omega$

[2] Kanda, M. Standard Antennas for Electromagnetic Interference Measurements and Methods to Calibrate Them, IEEE Transactions on Electromagnetic Compatibility, Vol. 36, No. 4, November 1994, pp 261 - 273. (Reprinted in EMC/EMI Selected Readings, Ed. Kodali, IEEE Press 1996)

[3] Collin, R. & Zucker, F, Antenna Theory Part 1, McGraw-Hill 1969, Section 11.8, The Shielded Loop.

[4] International Telecommunications Union, ITU-R Recommendations, 1994 PI Series Volume, 'Propagation in Ionized Media' Recommendation ITU-R PI.372-6.