BBC

Research & Development

White Paper

WHP 403

December 2022

The Measurement of Magnetic Loop Antenna Performance

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BRITISH BROADCASTING CORPORATION

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Abstract

Despite the advent of the internet and streamed services, there is still an interest in traditional broadcasting. With that interest comes a need to measure the associated electromagnetic field strengths, both of the wanted transmissions and of atmospheric noise that could be interfering with them. For instance, at the time of writing, the BBC intends to transmit a trial DRM service on 3,955 kHz. It is also concerned about interference from wireless power transfer (WPT) devices to AM transmissions in the LF and MF bands. An accurately calibrated and sensitive antenna is needed for such work.

This White Paper looks at the characterisation of magnetic loop antennas — those sensitive to the magnetic component of an electromagnetic field. Generally, such antennas are intended for use below 30 MHz. Unfortunately, generation of a travelling electromagnetic wave at these frequencies — the ideal test signal — is impracticable within any reasonable enclosed space; the wavelength at 200 kHz is 1,500 m, for instance. To make matters worse, these antennas are often too large to be tested satisfactorily in a substitute such as a GTEM- or PTEM-cell.

An alternative to these traditional methods involves the use of a ‘small coil’ placed at the centre of the loop antenna. This is described here, along with the work needed to eliminate possible sources of error. The result is a reliable plot of antenna factor (the field strength divided by the antenna output) versus frequency. By combining the antenna factor with the noise measured at the output of the antenna, it is then possible to calculate the equivalent noise field — an essential parameter in deciding whether an antenna is capable of measuring the expected levels of environmental noise.

There is also a short discussion on the use of magnetic loop antennas and possible shortcomings such as spurious sensitivity to electric fields.

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The Measurement of Magnetic Loop Antenna Performance

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1. Introduction and Background

Despite the advent of the internet and streamed services, there is still an interest in traditional broadcasting. With that interest comes a need to measure the associated electromagnetic field strengths, both of the wanted transmissions and of atmospheric noise that could be interfering with them. For instance, at the time of writing, the BBC intends to transmit a trial DRM service on 3,955 kHz. It is also concerned about interference from wireless power transfer (WPT) devices to AM transmissions in the LF and MF bands. An accurately calibrated and sensitive antenna is needed for such work.

This White Paper will look at the characterisation of low-frequency magnetic loop antennas. Generally, such antennas are intended for use below 30 MHz. Unfortunately, generation of a travelling electromagnetic wave at these frequencies — the ideal test signal — is impracticable within any reasonable enclosed space; the wavelength at 200 kHz is 1,500 m, for instance. To make matters worse, these antennas are often too large to be tested satisfactorily in a GTEM- or PTEM-cell. (These devices will be described later.)

An important parameter of antenna performance is the antenna factor (AF). This equals the field-strength of the electromagnetic field expressed in volts per metre divided by the output of the antenna in volts. Hence the unit of AF is metre^{-1}. It is often expressed in dB/m, with the AF in dB/m equalling 20 × $\log_{10}$ (AF in metre^{-1}). Also of interest is the equivalent noise field of the antenna: the noise, in volts per metre, that the antenna appears to be receiving from the environment even when there is none. This is defined as the noise output in a given bandwidth times the antenna factor at the frequency of interest. It is generally converted into dB relative to 1 µV/m, or $\mathrm{dB}_{\mu V/m}$. As will be explained, the measured noise field of the environment is a function of the directional properties of the antenna, and is not absolute; it is distinct from the environment's excess noise factor, which is truly independent of the antenna.

It is important to appreciate that the loop antennas under consideration are (or should be) only sensitive to the magnetic field, or H-field. However, the convention is to convert the H-field into an equivalent electric field, or E-field, on the assumption that the relationship between the two is the same as for an electromagnetic wave propagating through free space. Such a wave comprises both E- and H-fields, with the ratio E/H equalling 120 π ohms — the impedance of free space.

2. Receiving Antenna Calculations

As an introduction to the behaviour of loop antennas, a simplified model is given below. An E-field antenna, based on a capacitor, is also shown as a comparison.

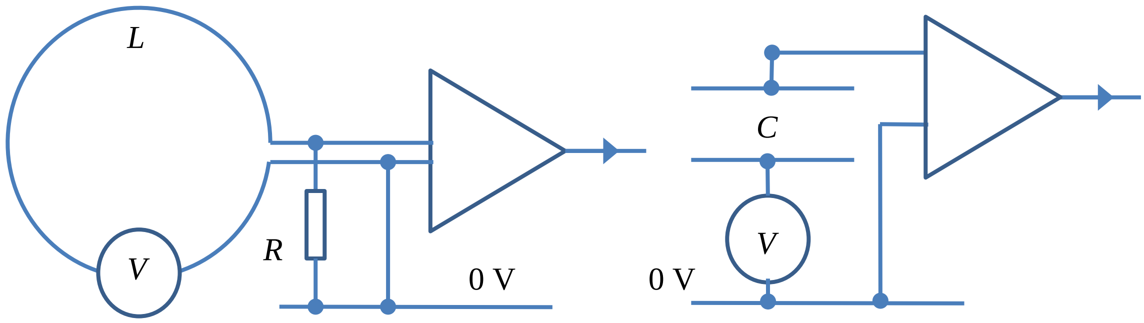

Figure2.1: Models of H- and E-Field Antennas

The loop itself has an inductance L, and we assume that it is loaded with a resistor of value R. Thereafter, the output of the loop is amplified noiselessly. The voltage V induced in the loop by a magnetic field H is given by the formula

$$V = \mu_0 \pi b^2 \ dH/dt \ ,$$

where πb^2 is the area of the loop (of radius b) and μ_0 is the permeability of free space. Letting the equivalent E-field equal $120\pi H$, and letting dH/dt equal $2\pi f H$, gives us

$$V = \mu_0 \pi b^2 \times 2\pi f \ (E/120\pi),$$

where f is the frequency. The antenna being used as an example in this White Paper has a radius of 0.465 metres. Putting figures into the equation:

$$V = (4\pi \times 10^{-7}) \times (\pi \times 0.465^2) \times 2\pi f \ (E/120\pi) \ .$$

The antenna factor of the loop alone is then

$$AF_{loop} = E/V_{out} = 70.3/f_{MHz}.$$

Expressed in dB, this becomes $AF_{loop\ (dB/m)} = 36.9 - 20\ log_{10}\ f_{MHz}$.

Note that the output of the loop is proportional to frequency. That is because the output is proportional to the *rate of change* of the magnetic field. On the other hand, the output of an E-field antenna (capacitor) is independent of frequency. Calculations for the E-field antenna are very straightforward: if the field is E and the capacitor plates are separated by distance d, $V = E/d$. Unfortunately, the output impedance, $1/(2\pi f C)$, is very high for any reasonable values of f and C, and the loading imposed by the following amplifier is likely to have a serious effect.

For the loop antenna, the resistor R has the effect of giving a constant antenna factor above a given corner frequency, since R works with the inductance L to give a first-order low-pass filter; the falling response of the filter above its corner frequency neatly cancels the rising voltage induced in the loop. The overall output (for any frequency) is given by

$$V_{out} = \mu_0 \pi b^2 \times 2\pi f \ (E/120\pi) \ \times \ R/\sqrt{(R^2 + (2\pi f \ L)^2)} \ .$$

The right-hand term represents the attenuation introduced by loading the loop with R. At high frequencies, where $2\pi f L$ is much greater than R, the expression becomes

$$V_{out} = \mu_0 \pi b^2 \times (E/120\pi) \ \times \ R/L \ ,$$

which is independent of frequency. As we shall see, the inductance of a loop of radius 0.465 m is about 2.3 μH. If we want the output of the loop to be independent of frequency above say 1 MHz, R becomes $2\pi \times (1\times10^6) \times (2.3 \times 10^{-6}) \approx 15\ \Omega$. Although loading the loop in this way is not the ideal method of obtaining a flat response, it serves to show that H-fields and the loops used to detect them are essentially ‘low impedance’, and that makes them easy to work with.

We have already said that the output impedance of the E-field antenna is ‘very high’. If we (optimistically) take the antenna’s capacitance as 10 pF, the impedance $1/(2\pi f C)$ at 1 MHz is $1/\{2\pi \times (1\times10^6) \times (10 \times 10^{-12})\} \approx 16\ k\Omega$. This is certainly a high value compared with that of the loop antenna, and the situation becomes worse as the frequency decreases. We also have to contend with the input capacitance of whatever the antenna is connected to. This will attenuate the antenna’s output by an amount that might be difficult to determine.

It is instructive to calculate the relative outputs of the two types of antennas. Suppose we take the frequency to be 1 MHz as before, and the E-field to be 1 V/m. In free space, the associated H-field is 1 V/m divided by 120π ohms, or 2.65×10^{-3} A/m. The output of our loop antenna is 1/70.3 V, or 14.2 mV, from our earlier formula $AF_{loop} = E/V_{out} = 70.3/f_{MHz}$. If the plates of our E-field sensing capacitor are 0.1 m apart, the formula $V = E/d$ gives us 100 mV. The E-field sensor is clearly the winner here.

In summary, where the field strength of a remote transmission is to be measured, and where the relationship $E/H = 120\pi$ applies, an H-field antenna is to be preferred, because the low impedance makes it easier to work with. Of course, locally generated E-fields could well be of interest, in which case an E-field antenna is essential.

3. Calibration of Loop Antennas: Methods

As mentioned in the Introduction, the ideal method of calibrating an antenna is to place it in a 'free space' electromagnetic field and measure the output. At the low frequencies in question, the very long wavelengths make this impracticable, and we must seek an alternative. The possibilities are as follows.

The GTEM-Cell Plenty of literature about GTEM-cells is available on the internet; see **[1]**, for instance. However, a simplified description is as follows. 'TEM' stands for transverse electromagnetic, since a transverse electromagnetic wave can be thought of as propagating down the length of the cell. The 'G' refers to gigahertz, because such cells are usable at microwave frequencies, despite their large size. In essence, the cell comprises a large metal box shaped like a giant wedge of cheese, with a plate known as the septum suspended from the roof. The septum combined with the metalwork of the cell behaves as a transmission line, with the test signal being applied to the narrow end, and being absorbed in terminations at the wide end. Usually, the equipment under test (the EUT) is placed beneath the septum at the wide end.

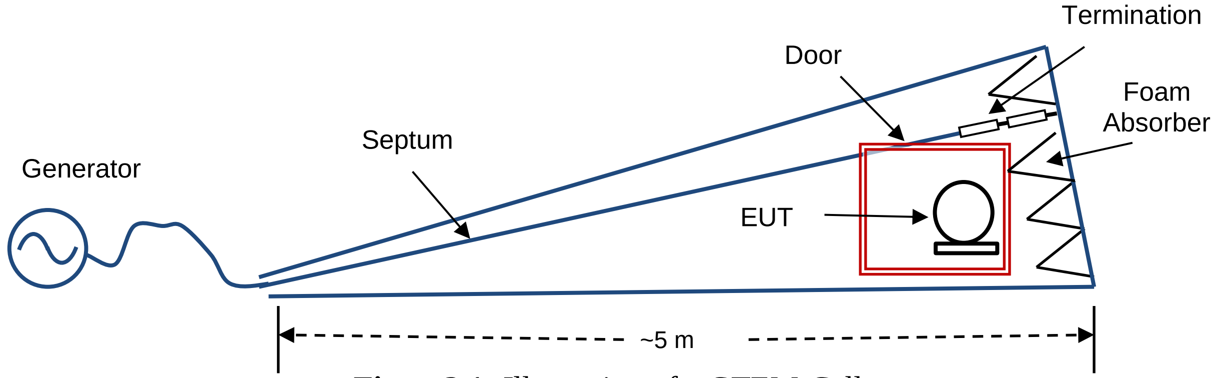

Figure3.1: Illustration of a GTEM-Cell

In principle, the calculations are simple: the electric field equals the voltage on the septum divided by the distance between the septum and floor in the test area. By a wonder of physics, the magnetic field is in the correct proportion to the electric field ($E/H = 120\pi$) provided that the septum is correctly terminated.

In practice, there are sources of error. Since the septum is at an angle to the floor, the field cannot be completely uniform or purely transverse, and the sidewalls also have an effect. Moreover, reflections arise because of imperfect terminations. Manufacturers do not claim spectacular accuracy: **[2]**, for instance, quotes ±3 dB up to 1 GHz, and ±4 dB above that. Yet another source of error is the effect that EUT has on the field. The rule-of-thumb is that the EUT should not occupy more than one third of the distance between floor and septum. For our example antenna, the septum would need to be some 3 metres from the floor — a big cell indeed!

The PTEM-Cell The PTEM-cell, or pseudo-TEM-cell, is a poor man's GTEM-cell. It was developed in BBC R&D over 30 years ago, originally for the testing of portable radios. It comprises a screened room with a pair of open transmission lines — one near the floor and one near the ceiling. These are fed with the test signal in antiphase, with the EUT being placed about halfway between the upper and lower lines. The electromagnetic field between the lines can be both measured and calculated, as described in an earlier White Paper **[3]**. As for the GTEM-cell, the ratio of E-field to H-field is the same as it would be in free space, with the same proviso of the transmission lines being correctly terminated.

The PTEM-cell is most suitable for measuring small items, since the field varies significantly with position between the transmission lines. A loop such as our example takes up nearly half the distance between the lines (assuming a separation of about 2 m), making the effective field difficult to calculate — although not impossible. There is the further complication that the presence of the loop distorts the field, just as it does with the GTEM-cell. For the present job, the PTEM-cell makes a useful confidence check on the results of the 'small-coil' method to be described, but the measurements should not be taken as exact.

The Meguro Loop The Meguro loop is a professional device intended for the testing of AM portable radios. It comprises a single turn of wire 0.25 m in diameter enclosed in a metal tube to provide electrostatic screening. There is an 86 Ω resistor in series with the single turn, since this, together with the 50 Ω output impedance of the generator, gives a convenient calibration factor when the loop is at the 'standard' distance of 0.6 m from the EUT. Although the device works up to 30 MHz or so, it is of limited use on the HF band, as it only generates an H-field; HF portable receivers are mostly fitted with telescopic antennas that respond to the E-field.

Calculations of the H-field and the equivalent E-field are as follows:

If I is the current flowing through the loop, a is the radius of the loop, and r is the axial distance between the loop and the item under test, the H-field is given by

$$H = \frac{I}{2}\ a^2/(r^2 + a^2)^{\frac{3}{2}}$$

The current I equals the generator EMF (open-circuit output voltage) V divided by the total circuit resistance of 136 Ω (50 Ω from the signal generator plus 86 Ω from the loop). Since the equivalent E-field is calculated on the assumption that E/H = 120 π, we have

$$E = 120\pi\ \frac{V}{136}\ \frac{1}{2}\ a^2/(r^2 + a^2)^{\frac{3}{2}} \approx 1.386\ \ a^2/(r^2 + a^2)^{\frac{3}{2}}$$

Where r is much greater than a, we can simplify this to $E = 1.386\,V\ \ a^2/r^3$. Putting in a = 0.0625 and r = 0.6 gives

$$E = 1.386\,V \times (\ 0.125^2/0.6^3\ ) = \ V/10\,.$$

In other words, the equivalent electric field equals the generator EMF divided by 10.

Implicit here is that the dimensions of the EUT are small compared with the separation r. If they are not, r should be increased. The equivalent field is then calculated using the inverse cube law: doubling r reduces E by a factor of 8. Alternatively, a more sophisticated formula is available, as described for the 'small coil, extended' overleaf.

Sometimes the Meguro loop is used in the same plane as EUT, rather than axially. If so, the field is half that calculated above.

The 'Small Coil' An attractive proposition is to place a small coil (a miniature version of the Meguro Loop) at the centre of the EUT. In that position, the effective field is maximised, and is less likely to be influenced by reflections from the surroundings (since these will be much smaller than the direct field). Calculating this effective field appears difficult, but there is a useful trick to help us. Any flux generated by the small coil and circulating within the area of the EUT has no effect. This is because equal amounts of flux pass in both directions (in and out of the paper in Figure 3.2). It follows that we only need to calculate the flux existing *outside* the EUT, since this must equal the remaining effective flux within the EUT.

The field generated by the small coil falls with the inverse cube of the distance r, and can be integrated over the area A outside the EUT to give the flux linkage $\emptyset$. If the EUT has radius b,

$$H \approx I\ a^2/4r^3\,.$$

$$\emptyset = \int_{r=b}^{r=\infty} H\,dA = \int_{r=b}^{r=\infty} I\,a^2\,(1/4r^3)\ 2\pi r\,dr$$

Integrating,

$$\emptyset = (\pi I a^2/2)\ [-1/r]_b^\infty \ = \ (\pi I a^2/2b)$$

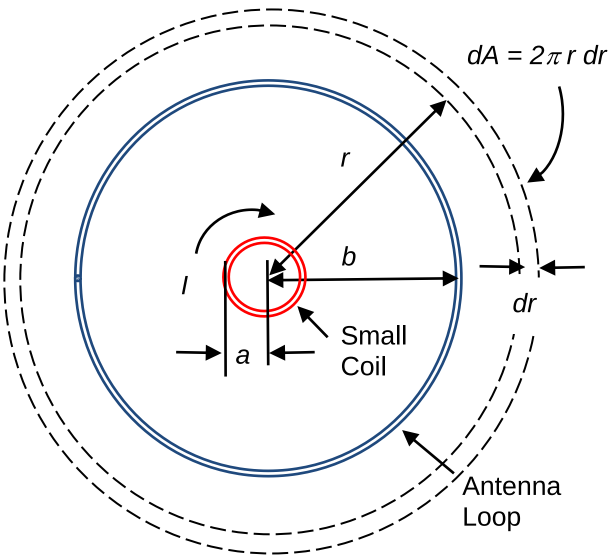

Figure 3.2: Calculating the 'Small Coil' Field

The effective field H_e is then the flux linkage divided by the area of the antenna loop:

$$H_e = \emptyset / \pi b^2 = I\,a^2\,/2\,b^3$$

Multiplying by 120 π gives the equivalent electric field:

$$E_e = 60\pi\,I\,a^2\,/\,b^3$$

This formula can be extended slightly to give E_e in terms of the generator EMF V, the total resistance in the circuit R, and the small coil's number of turns n (so far, we have assumed a single turn):

$$E_e = n\,(V/R)\,120\pi\,(a^2/2b^3)$$

The small coil method is looking to be the most promising technique so far, but even that is subject to errors. These will be discussed later, but we can mention now that the self-inductance of the small coil can be significant at higher frequencies, especially if the coil possesses more than a single turn.[1] If so, R needs to be replaced by the total circuit impedance Z (or strictly its modulus). For a self-inductance L,

$$Z = \sqrt{(2\pi f\,L)^2 + R^2}\,.$$

BBC R&D White Paper 091 **[4]** has a companion spreadsheet which is useful for calculating the self-inductance of coils.

The 'Small Coil' Extended Our 'small coil' formula is a useful result, but it makes the (usually valid) approximation that the diameter of the small coil is much less than that of the antenna loop. It also assumes that the coil and loop lie in the same plane and on the same axis (usually easy to arrange). According to an article discovered with the help of Google, **[5]** a more exact formula is

$$E = \frac{60\pi\,a^2\,I}{(a^2 + b^2 + r^2)^{3/2}}\left[1 + \left(\frac{2\pi r}{\lambda}\right)^2\right],$$

where r is the axial distance between the coil and loop, and λ is the wavelength at the test frequency.

If r is zero and a is much smaller than b, the equation reduces to our earlier formula:

$$E = \frac{60\pi\,a^2\,I}{(0^2 + b^2 + 0^2)^{3/2}}\left[1 + \left(\frac{2\pi 0}{\lambda}\right)^2\right]$$

$$= \frac{60\pi\,a^2\,I}{b^3}$$

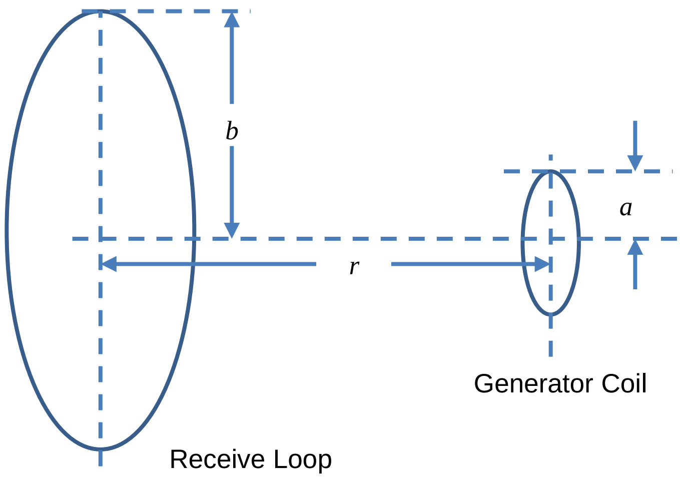

Figure 3.3: The Extended Small Coil

The quantity in the square brackets dominates when r is greater than $\lambda/2\pi$, and describes what happens in the far-field region. At high frequencies, where $(2\pi r/\lambda)^2 >> 1$,

$$E = \frac{60\pi\,a^2\,I}{(0^2 + 0^2 + r^2)^{3/2}}\left[\left(\frac{2\pi r}{\lambda}\right)^2\right] = \left(\frac{2\pi}{\lambda}\right)^2 \times \left(\frac{60\pi\,a^2\,I}{r}\right).$$

Notice that E falls away as $1/r$ rather than $1/r^3$. This E-field really does exist, and is not simply a convenient conversion of the H-field. Of course, in the far-field region, E and H are linked together, and E/H equals 120π. The receive loop antenna is still only responsive to the H-field.

[1] Remember that the self-inductance of a coil is proportional to the *square* of the number of turns.

4. Confidence Checks and High-Frequency Effects

For the rest of this report, the emphasis will be on using the 'small coil' method of determining the AF of loop antennas. However, the other methods will be given the chance to agree or disagree with the results, and so provide confidence checks.

The results from an early pair of measurement runs are plotted below. The set-up was straightforward: the EUT (i.e. the antenna loop under test) was suspended from the ceiling of BBC R&D's Screened Room, and the small coil hung in the middle of the loop. The coil was driven from the output of an RF signal generator. At that stage, the effect of the Screened Room walls had not been calculated, and so, to find out, the loop/coil arrangement was moved around, away from and closer to the walls. Doing this made very little difference to the output of the loop, suggesting that any such effect could be ignored — at least for these initial measurements.

The output levels of the RF generator and the EUT were measured with the same instrument (a spectrum analyser). This eliminated the need to determine absolute levels accurately, since only the *difference* in level (in dB) was of interest. The current through the generator coil was taken to be the generator EMF divided by 100: 50 Ω from the generator output impedance and 50 Ω built into the coil. Screened cable, earthed at one end, was used for the winding of the coil — screened to avoid the possibility of electrostatic pick-up by the EUT.[2] The coil former was a short length of plastic drainpipe of approximate diameter 70 mm.

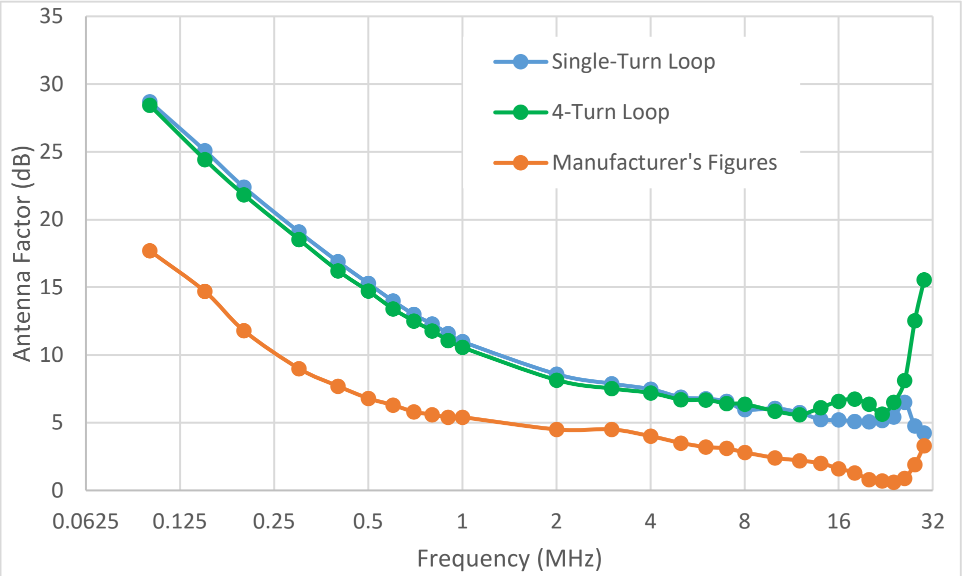

Figure 4.1: Early Results for Antenna 'A'

The need for confidence checks is obvious from Figure 4.1. The two sets of results, using single-turn and four-turn coils, agree well with one another up to about 12 MHz, but not with the figures given by the EUT's manufacturers. All three plots show the expected 6 dB per octave fall in sensitivity below a certain corner frequency, as explained in Section 2. It is odd that the measured corner frequency differs from that quoted by the manufacturer — a calibration error would most likely affect the sensitivity over the whole band rather than the corner frequency.

The AF 'coggles' beyond 16 MHz give rise to suspicion. As the EUT is a fairly simple device — just a loop followed by a low-noise amplifier — there is no likely mechanism for sharp changes such as these. The worst offender is the 4-turn coil, but its larger self-inductance cannot be blamed entirely, since the apparent sensitivity changes too rapidly with frequency; it should not exceed 6 dB per octave. It turned out that the main culprit was coupling between the test interconnections, where cable lengths could approach a quarter-wavelength (and resonance) at the top end of the band.

2 This is the same philosophy as adopted for the Meguro loop.

An example of a confidence check is given below. The measurements were made in the large GTEM-cell owned by the Digital Television Group.

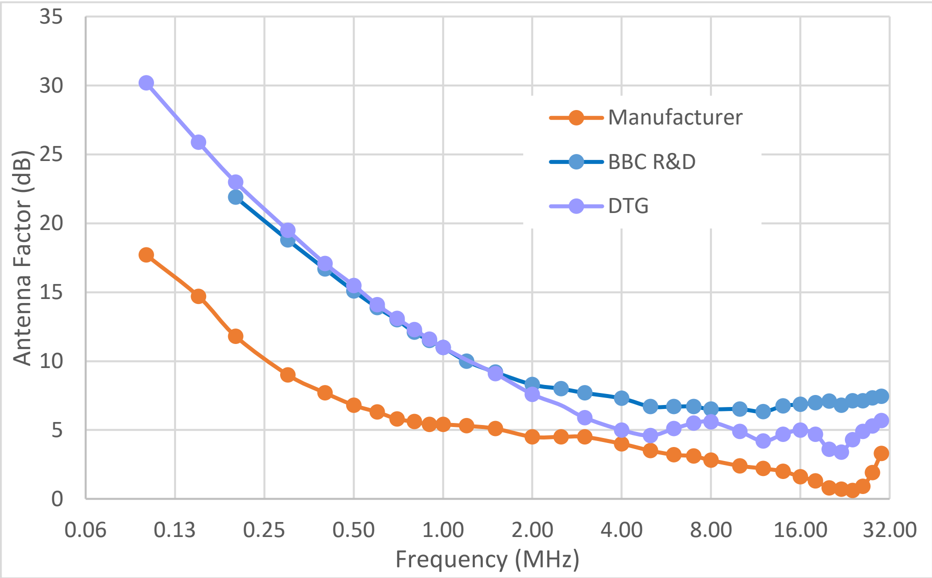

Figure 4.2: A Confidence Check on Antenna ‘A’, Using a GTEM-Cell

Agreement is good below 2 MHz, and at least bears out the probability that the manufacturer’s stated corner frequency is wrong. The ripples beyond 2 MHz are attributable to the GTEM-cell, and possibly the test interconnections; as mentioned before, such ripples are unlikely to be a property of a simple device such as the EUT. It is interesting to note that the ‘anomaly’ at the top-end of the band has disappeared.

A similar check was made with the EUT placed in BBC R&D’s PTEM-cell. No great accuracy was expected, for the reasons given in Section 3; also, the cell was in need of calibration. However, it was hoped that there would be agreement on the corner frequency. Fortunately, there was!

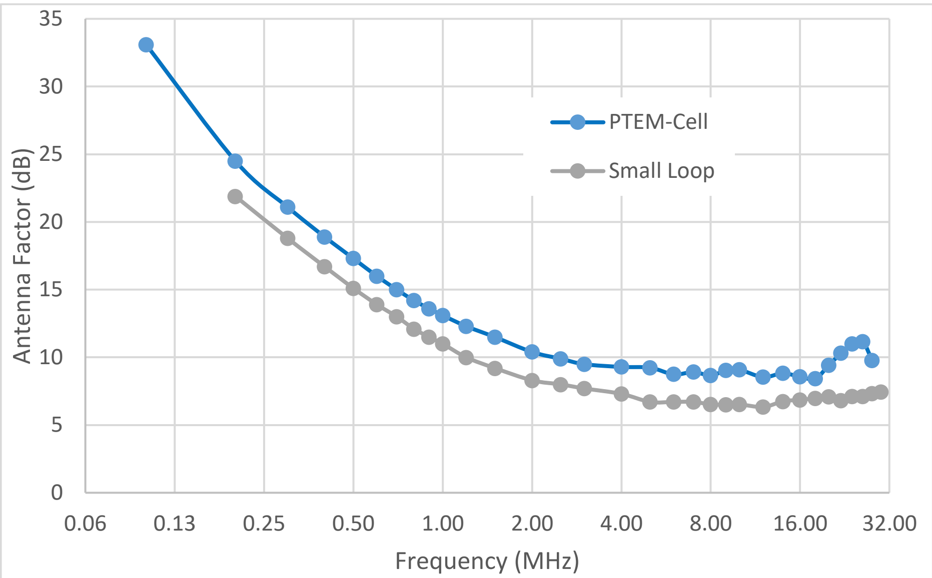

Figure 4.3: A Confidence Check on Antenna ‘A’, Using the PTEM-Cell

Some far-field checks were made by using the EUT to measure the field-strengths of broadcast MF transmissions. The measurements were then compared with those made using a professional field-strength meter. Agreement was good.

5. An Optimised Test Arrangement

So far, the evidence favours of the small-coil method, at least at lower frequencies. It is now time to refine the method and reduce uncertainties. To start with, we need direct monitoring of the current flowing through the small coil, and we need to eliminate unwanted coupling between the cables. The arrangement adopted is illustrated below:

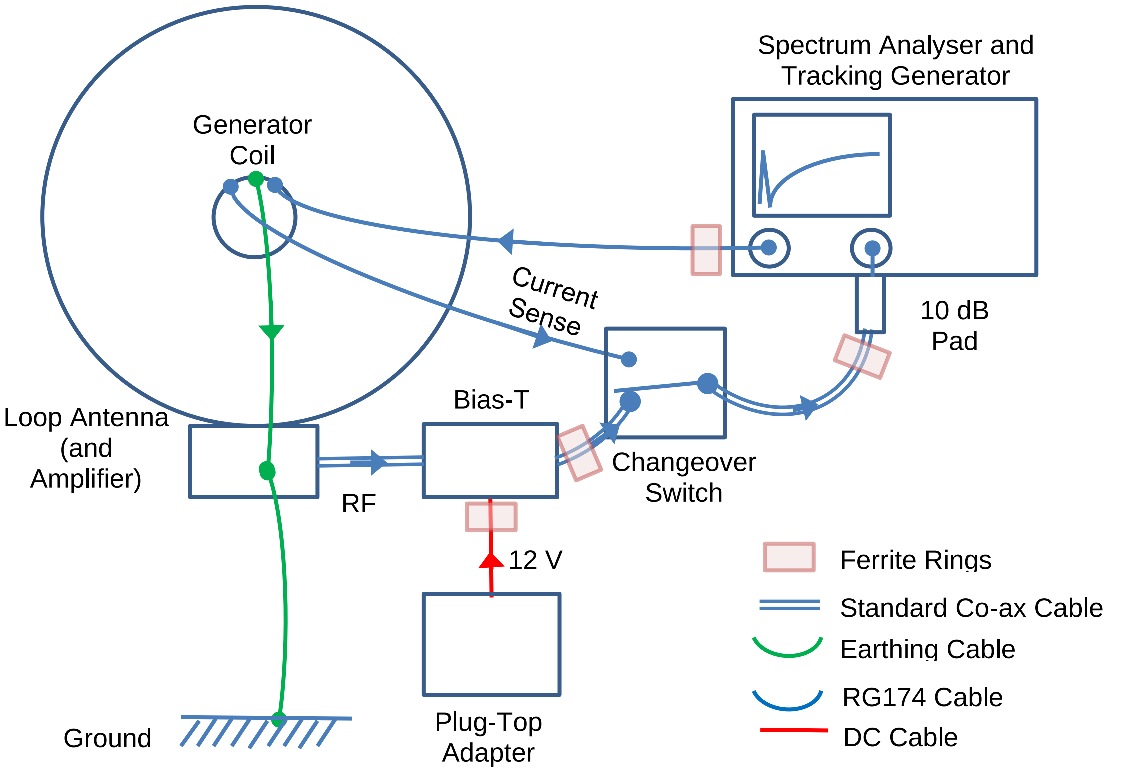

Figure 5.1: The Optimised Test Arrangement

The most important thing is secure earthing of both the generator coil and loop antenna, with the earthing cable lengths being kept well below $\lambda/4$ at the highest frequency of interest. (At $\lambda/4$, the cables make efficient monopole antennas — something not wanted here!) The remaining test interconnections are generously loaded with ferrite rings to ensure good isolation. A final refinement is to provide a changeover switch so that an easy comparison can be made between the current flowing through the coil and the output of the EUT.

The coil arrangement and current monitoring is shown in Figure 5.2. The essential idea is to measure the voltage across a low-value resistor placed in series with the coil. In this case, 1 A flowing through 2.35 Ω gives rise to 2.35 V. Because of the 47 Ω termination resistor, this is halved when measured with a device having a 50 Ω input impedance.

The low-value resistor is chosen so that the signal level remains nearly the same when the changeover switch in Figure 5.1 is operated. This avoids possible errors caused by range-changing in the measurement device.

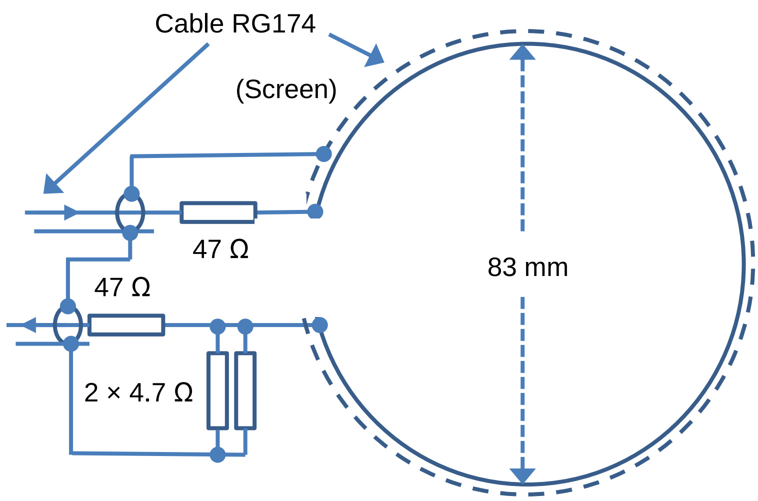

Figure 5.2: Details of the Coil and Current Sensing

A slight error remains because of the small self-inductance associated with the current-sensing resistors. This causes the apparent current to rise with frequency. However, the error can be measured and allowed for; here it amounted to 3 dB at 40 MHz, and was below 2 dB at the highest frequency of interest (30 MHz).

In the previous section, we mentioned the possible effect of the screened room's metal walls, and decided that, to a first approximation, it was negligible. However, we can now be more scientific about it. The starting point is our well-worn formula

$$E = \frac{60\pi\, a^2\, I}{(a^2 + b^2 + r^2)^{3/2}} \left[1 + \left(\frac{2\pi r}{\lambda}\right)^2\right].$$

Assuming that we can ignore the term in the square brackets, we see that the relative strength of the reflection is

$$= \frac{60\pi\, a^2\, I}{(a^2 + b^2 + r^2)^{3/2}} \div \frac{60\pi\, a^2\, I}{(a^2 + b^2)^{3/2}} = \left\{\frac{(a^2 + b^2)}{(a^2 + b^2 + r^2)}\right\}^{3/2}.$$

where a is the radius of the generator coil, b is radius of the antenna loop, and r is the distance between the (co-sited) coil and loop and the reflection of the coil. We may take r as twice the distance between the loop and the wall, since the reflection is sited at distance r behind the wall. The experimental and theoretical results are plotted below for reflections off the end-wall of the Screened Room:

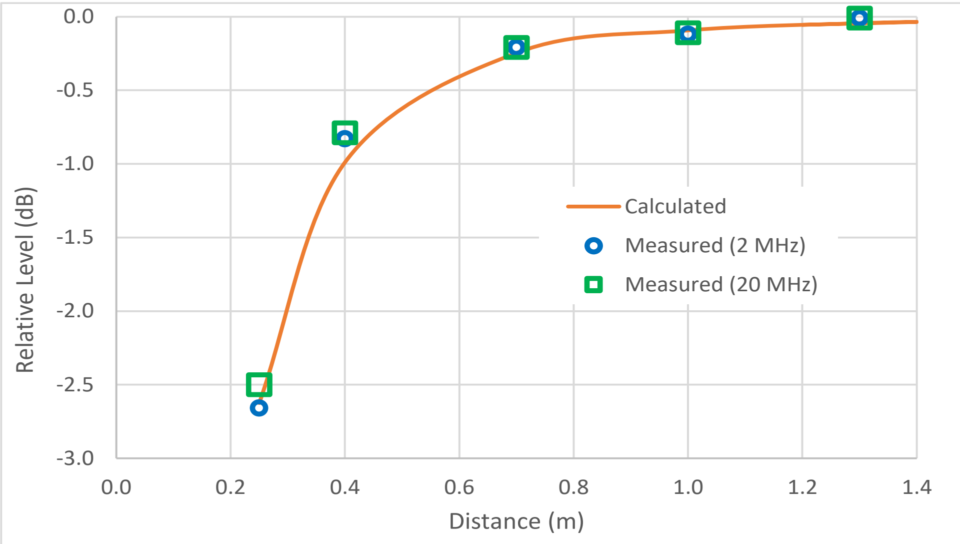

Figure 5.2: The Effective Loss Caused by Reflection in a Conducting End-Wall

The 'relative level' in Figure 5.2 is calculated by subtracting the image field from the direct field (before conversion to dB). For the measured values, the distance has been 'corrected' by 30 mm to give the best fit to the calculated figures. This tactic can be excused by the difficulty in determining the effective centre of the generator coil. As can be seen, the calculation gives the right answers, and there is no evidence of a frequency dependence.

The corresponding results for reflections from the sidewalls are plotted overleaf. Here the antenna loop is perpendicular to the wall, and so the minimum distance is restricted to the radius of the antenna. The combined effect of *both* sidewalls of the Screened Room has been calculated: the wall separation of 2.4 m makes it impossible to measure the effect of one wall in isolation. When the distance is given as 1.2 m, the two walls are making equal contributions.

Where the reflection was from the end-wall, the virtual coil (the reflection) and the antenna loop were coaxial, and the resulting field strength was as just calculated. The reflection from a sidewall is co-planar, and the effective field-strength is half as great, as mentioned under 'Meguro Loop' in Section 3. An already small quantity is now even smaller and hence difficult to measure accurately. One can only say that the results are consistent with theory.

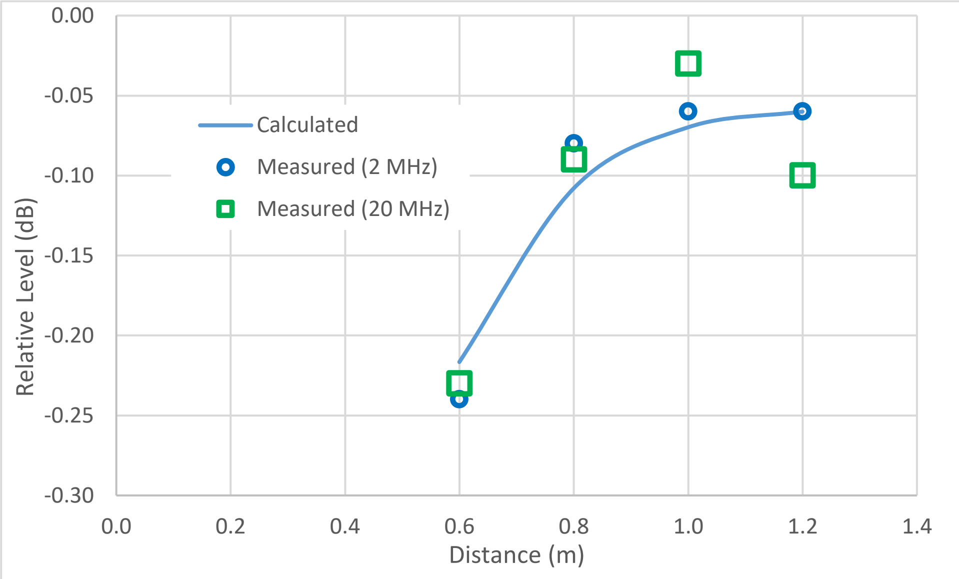

Figure 5.3: The Effective Loss Caused by Reflections in Conducting Sidewalls

In summary, we can take the total loss to be 0.15 dB: 0.05 dB because of the end-wall, 0.05 dB from the two sidewalls, and 0.05 dB from the ceiling and floor. These are very minor corrections, and they do not need to be known with great accuracy.

The presence of the end-wall could also reduce the inductance of the antenna loop, in turn affecting the corner frequency (below which the response falls at 6 dB per octave). It would be surprising if the effect were appreciable, but we must make sure. Calculation is beyond the capability of the author, but he could make a practical measurement! To do this, the loop was disconnected from the low-noise amplifier and connected to a component bridge. The test frequency was 1 MHz.

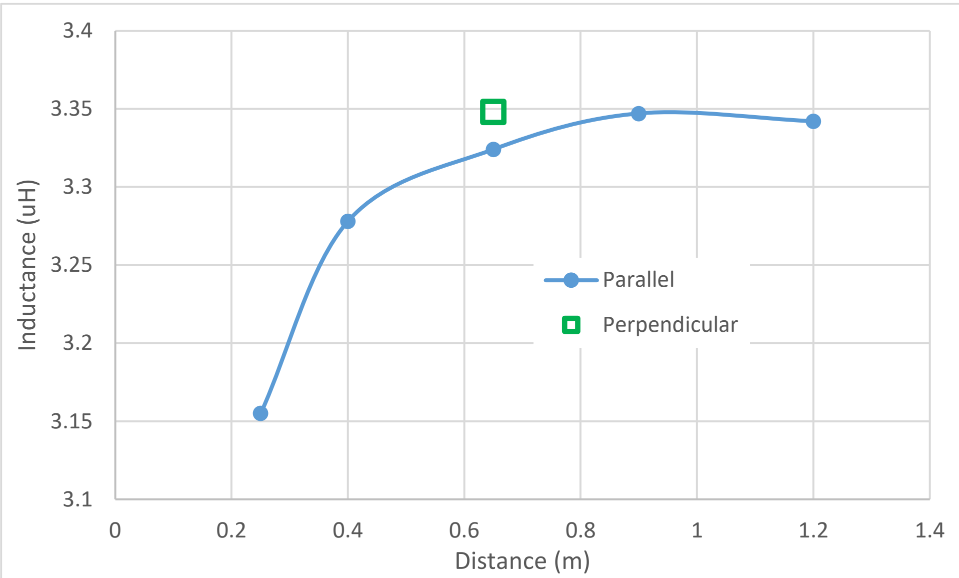

Figure 5.4: The Effect of Conducting Walls on Loop Self-Inductance

For the ‘perpendicular’ measurement, the loop was moved towards the sidewall, with its rim 0.15 m from the wall. It is evident that the inductance is not appreciably affected by the position of the loop, and no allowance needs to be made for any sensible distance.

As will be seen in the next section, the true loop inductance is rather less than the measured value of 3.34 µH given in Figure 5.4. This is likely to be the result of the unavoidable connections between the loop and component bridge. However, the *change* in inductance remains true.

6. Antenna Output Impedance

The manufacturers make it clear that their antennas should be terminated in 50 Ω. This requirement has been observed for all the measurements quoted so far: for instance, the 'optimised test arrangement' of Figure 5.1 shows a 10 dB pad on the input of the spectrum analyser, in case the analyser itself does not offer a good match to 50 Ω. However, there is no implication that the output impedance of the antenna itself is 50 Ω. The actual impedance is an important factor if the antenna is to be used with a receiver whose input impedance is not so well controlled. Serious errors can then arise.

To calculate the output impedance, measurements of the output level were made with the output terminated (V_{term}) and unterminated (V_{unterm}). The output impedance Z_{out} is then given by

$$Z_{out} = 50 \times \{(V_{unterm} / V_{term}) - 1\} .$$

This is only strictly true if Z_{out} is resistive, but the errors are small provided that the ratio V_{unterm} / V_{term} is not close to unity (and Z_{out} is far removed from 50 Ω). As the results for Antenna 'B' show, Z_{out} indeed varies widely:

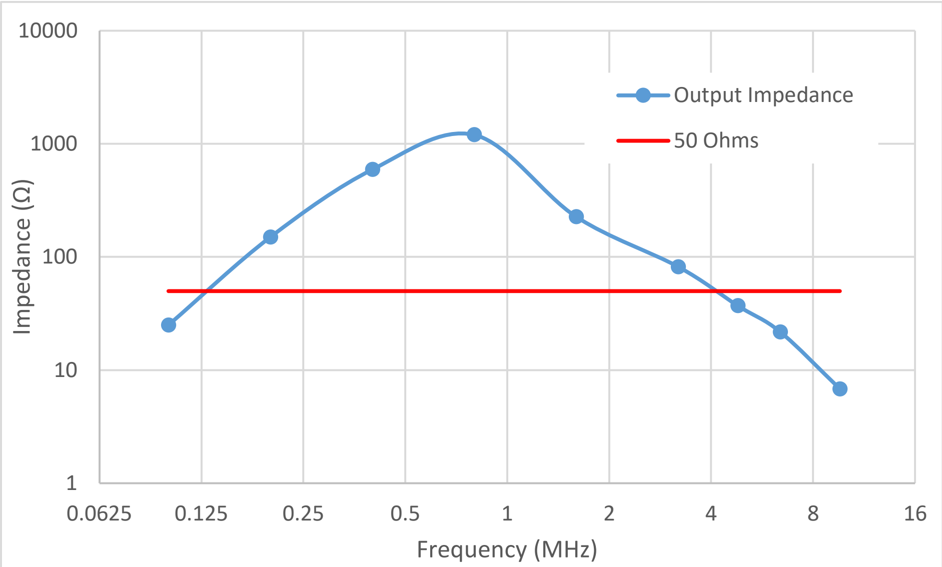

Figure 6.1: The Output Impedance of Antenna 'B'

These results are rather alarming. If whatever the antenna is driving has a high input impedance at around 1 MHz, the antenna output will be about 20 times the correct level, whereas it would only be twice as great for an antenna with a true 50 Ω output impedance. For the antenna in question, there were some signs of instability at 0.8 MHz when its output was unloaded.

Where the signal loss can be permitted, it is a good idea to add a low-value attenuator at the input to the driven device. A 3 dB pi-pad offers an impedance of about 150 Ω to the antenna, and so the output from the antenna will be 3 times that into 50 Ω. This is a considerable improvement: a well-behaved antenna with a 50 Ω output impedance would still give nearly twice the correctly terminated output.[3]

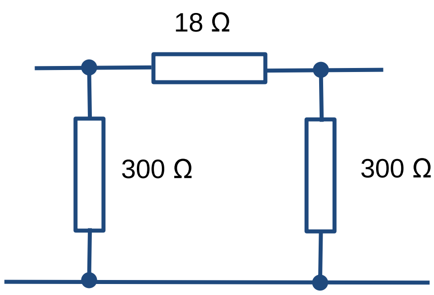

Figure 6.2: A 6 dB Pi-Pad

[3] It is assumed that a receiver system would be calibrated by driving the receiver from a generator with a true 50 Ω source impedance, even if the receiver did not provide a good match to 50 Ω. In other words, if the generator were set to '0 dBm', the receiver would be adjusted to indicate '0 dBm', even though that was not the true input level.

7. Antenna Noise Performance

The noise performance of the antennas is also an important issue, and hence something that we must determine. Many ‘professional’ antennas covering this frequency range concentrate on accuracy of calibration, but their self-generated noise can swamp the environmental noise that it is hoped to measure.

Making antenna noise measurements needs care: antennas tend to be good at picking up signals from the environment — that is their job! Obviously, the use of a screened room is essential. The test set-up used for the present measurements is illustrated below:

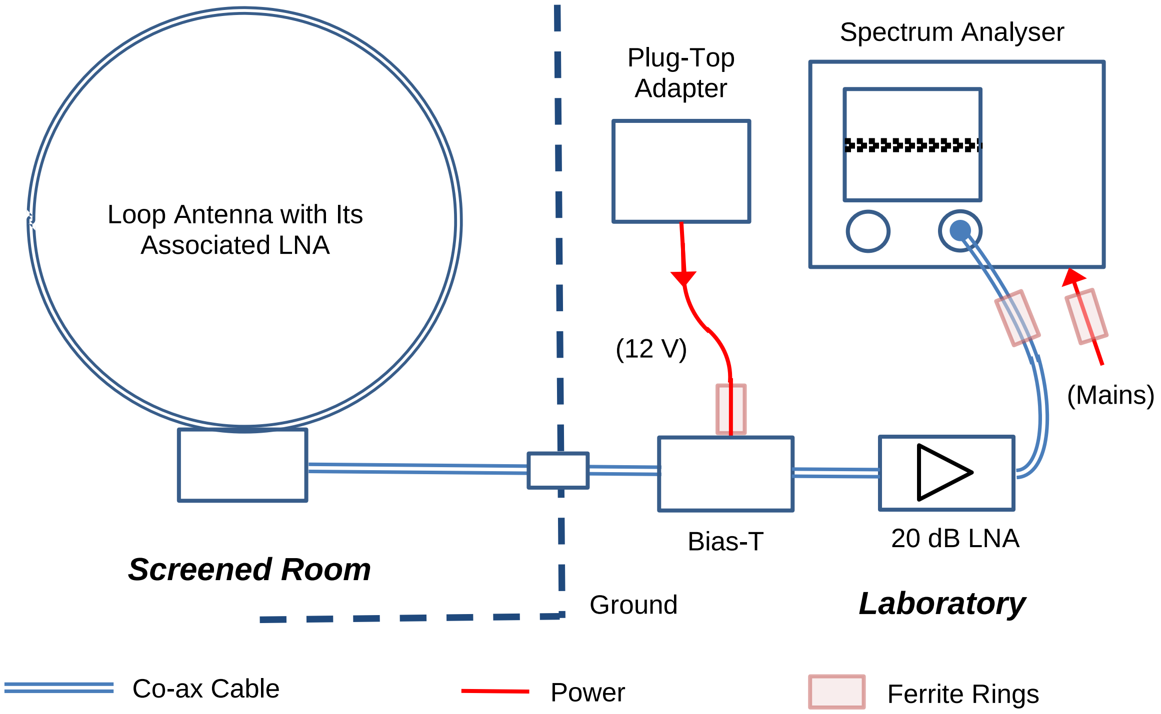

Figure 7.1: Measurement of Antenna Noise

Further details are as follows:

- To prevent the possibility of interference, all power to the Screened Room was switched off. Even the antenna’s plug-top adapter and the bias-T were kept outside.
- The connection to the antenna was made through a BNC feedthrough adapter in the wall of the Screened Room.
- The noise from the antenna was measured with an R&S FSH3 spectrum analyser. This analyser can measure noise within a given bandwidth (9 kHz here) using an RMS detector (hence giving a true power indication).
- Although the FSH3 has a built-in preamplifier, an external low-noise amplifier (LNA) is still needed for sensitive measurements such as these.

Calibration was carried out at each frequency step. The antenna factor measurements did not require an accurate measurement of absolute power; all that was needed was a comparison of the power emerging from the small-coil’s current monitor and the power from the antenna itself. With noise measurements, we do need to know the true power. To carry out the calibration, a signal generator was set to the required frequency, and its output level checked with a thermal power meter. The output, suitably attenuated, was then passed through the LNA and measured on the spectrum analyser. In that way, the effective gain between the antenna and analyser could be determined, and this value subtracted from the noise level displayed on the analyser.

Even with a good LNA, the system noise is still likely to be appreciable and would spoil the apparent performance of the antenna. This is particularly true at low frequencies where the antenna output falls away and the analyser noise increases. To overcome the problem, the noise was measured with a termination applied in place of the antenna's output. This residual noise was then subtracted from the apparent antenna noise, so giving the true noise output.

Calculations are then straightforward. The corrected noise power from the antenna, measured in dBm, is converted to $dB_{\mu V}$ by adding 107 dB. (1 V into 50 Ω is equivalent to +13 dBm, and 1 μV is 120 dB below 1 V.) The equivalent noise field (in $dB_{\mu V/m}$) is this figure plus the antenna factor (in dB/m).

The results for Antenna 'B' (the low-noise version of the antenna) are shown below:

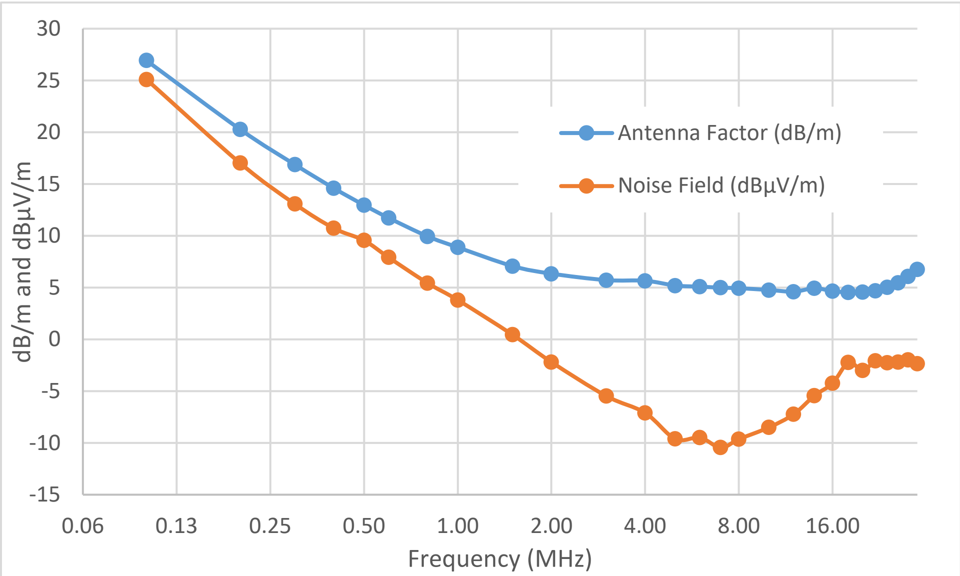

Figure 7.1: AF and Equivalent Noise Field for Antenna 'B'

It is worth asking whether the performance of the antenna is 'good'. In Section 2, we calculated that, provided $2\pi f L$ is appreciably greater than R, the output of the antenna V_{out} is related to the equivalent electric field E by

$$V_{out} = \mu_0 \pi b^2 \times (E/120\pi) \times R/L\,.$$

where L is the inductance of the loop and R is the load resistance. We can rearrange this expression to give the antenna factor:

$$AF = E/V_{out} = 1\,/\,(\mu_0\, b^2 \times (1/120) \times R/L)$$

L is fixed at 2.3 μH, and we choose R to be 15 Ω so that the expression is true above a corner frequency of 1 MHz. (As discussed earlier, beyond that frequency the increasing reactance of L counteracts the increasing voltage induced by the magnetic field.) Putting numbers in,

$$AF = E/V_{out} = 120\,/\,(4\pi \times 10^{-7} \times 0.465^2 \times 15/(2.3 \times 10^{-6})) = 68.$$

The thermal noise voltage generated in R is given by

$$V_n = \sqrt{4\,kT\,BR},$$

where k is Boltzmann's constant, T is the absolute temperature and B is the bandwidth. Hence,

$$V_n = \sqrt{4 \times 1.38 \times 10^{-23} \times 290 \times 9{,}000 \times 15} = 4.65 \times 10^{-8}\ \text{V}.$$

Multiplying this by the antenna factor gives the noise field:

$$E_{nf} = 4.65 \times 10^{-2} \times 68 = 3.16\ \mu\text{V/m}\,,$$

which is equivalent to 10 $dB_{\mu V/m}$. Since the antenna comfortably beats this, the performance is indeed good! Note that we can overcome this apparent theoretical limitation on the performance by amplifying the output of the loop before applying the frequency shaping.

8. Noise Fields and Excess Noise

We now consider how we can use an antenna to measure atmospheric noise.

The plots given in Figure 8.1 are taken from **[6]**, and show atmospheric noise levels in dB above the thermal noise-floor of 290 K. Note that the noise is only significant below about 30 MHz, and climbs steeply with decrease in frequency. The man-made noise plots C and E are based upon historic work and could possibly be out-of-date — one reason for calibrating the antennas is to come up with new figures! For the antenna to be useful, it should generate less noise than that being measured.

We need to translate the noise levels given in Figure 8.1 into equivalent field-strengths, and hence voltages appearing at the output of the receiving antenna. Only an outline is given here; anyone interested can follow the rather complicated thermodynamic arguments in **[7]**, **[8]** and **[9]** for instance.

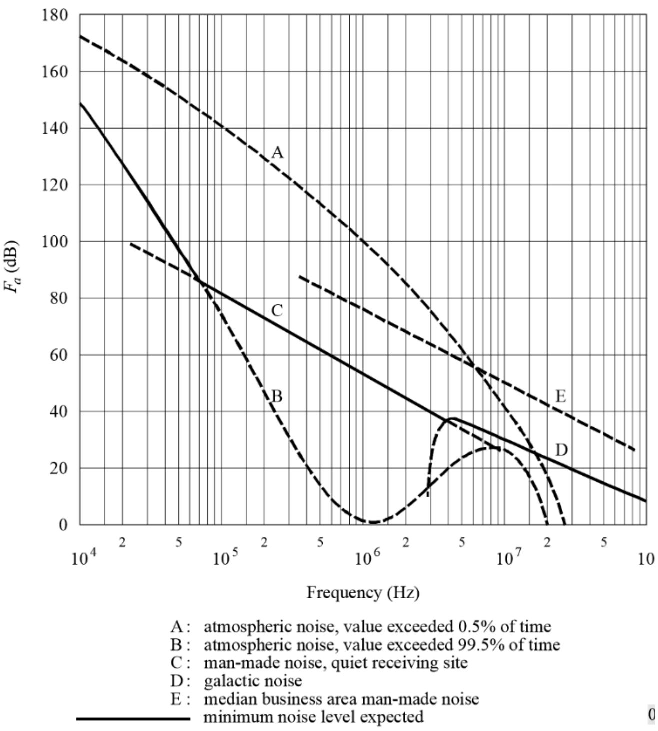

Figure 8.1: Atmospheric Noise Levels in the UK

We start with the idea that the environment is constantly emitting electromagnetic radiation — and absorbing it in equal quantities, since the system is in equilibrium (provided that it stays at the same temperature). It is usually satisfactory to assume *all* the energy falling on a surface is absorbed before being reradiated. The radiation is then said to be blackbody, and the total power per unit frequency radiated isotropically from unit area of the surface is given by Planck's Law:

$$P_\nu = \frac{2\pi h\nu^3}{c^2} \frac{1}{e^{h\nu/kT} - 1},$$

where h is Planck's constant and k is Boltzmann's constant.

At normal radio frequencies, $h\nu/kT$ is very small, and we can rewrite the equation as

$$P_\nu = \frac{2\pi h\nu^3}{c^2} \frac{kT}{h\nu} = \frac{2\pi \nu^2}{c^2} kT = \frac{2\pi kT}{\lambda^2},$$

where λ is the wavelength of the radiation.

This now needs to be translated into the received flux density B_υ. To do this, we imagine the radiation to be generated by a sphere S1 of radius a, and that this radiation is intercepted by a much larger concentric sphere, S2, of radius b. The radiated energy per unit frequency is then $P_\upsilon\, 4\pi a^2$; that is, P_υ times the surface area of S1. This must equal the energy $P'_\upsilon\, 4\pi b^2$ intercepted by S2, where P'_υ is the energy per unit frequency per unit area of that sphere. Thus

$$P'_\upsilon / P_\upsilon \;=\; 4\pi\, a^2 / 4\pi\, b^2 \text{, and so } P'_\upsilon = P_\upsilon\ (a^2/b^2)\,.$$

The solid angle subtended by S1 at a point on S2 is $\pi a^2/b^2$; hence the received flux density per unit frequency per unit solid angle, B_υ, is given by

$$B_\upsilon \;=\; P'_\upsilon\,/\,(\pi a^2/b^2) \;=\; P_\upsilon\ (a^2/b^2)/\,(\pi a^2/b^2) \;=\; P_\upsilon\ /\,\pi\,.$$

Finally,

$$B_\nu = \frac{2kT}{\lambda^2}.$$

The effective aperture of an isotropic antenna, A_{eff}, is given by $\lambda^2/4\pi$. The power captured over 4π steradians equals

$$\frac{1}{2} B_\nu \, 4\pi \, A_{eff} = \frac{1}{2} \frac{2kT}{\lambda^2} \times 4\pi \, \frac{\lambda^2}{4\pi} = kT.$$

The factor of ½ arises because the antenna only responds to one plane of polarisation. Finally, to determine the equivalent electric field E, we note that the power flow per unit area is given by the Poynting vector **E×H**, where H is the magnetic field. This is equivalent to E^2/Z_0, where Z_0 is the impedance of free space, or approximately 120π. We equate $A_{eff} \times E^2/Z_0$ with kT:

$$\frac{\lambda^2}{4\pi} \times \frac{E^2}{Z_0} = kT \text{ ; hence } E^2 = kT \, Z_0 \frac{4\pi}{\lambda^2} \text{ , or } E = \frac{1}{\lambda} \sqrt{4\pi \, kT \, Z_0} \, .$$

Letting kT equal 4.11×10^{-21} gives $(1/\lambda) \times 4.41 \times 10^{-9}$ V/m. For $\lambda = 1$, this comes to −47.1 $dB_{\mu V/m}$. The formula in Reference **[6]** expresses λ in 'dB' referenced to a frequency of 1 MHz. $1/\lambda$ becomes $10^6/ (3 \times 10^8)$, or −49.5 'dB'. For our isotropic antenna receiving thermal noise, the field-strength becomes

$$E = 20 \log f_{MHz} - 96.6 \text{ dB}_{\mu V/m}$$

This is for an 'ideal' environment and a bandwidth of 1 Hz. Where the bandwidth is given as B, with B being 10 log (actual bandwidth), and the excess noise figure of the environment is F_a, the expression becomes

$$E = F_a + 20 \log f_{MHz} + B - 96.6 \text{ dB}_{\mu V/m} \, .$$

Reference **[6]** quotes −95.5 $dB_{\mu V/m}$ for a short monopole above a ground-plane, and −99.0 $dB_{\mu V/m}$ for a dipole suspended in free space. However, we need a figure for a loop antenna (which we also assume to be suspended in free space. In the horizontal plane, the sensitivity varies as $\cos\theta$, as shown below. For the sake of argument, assume that power P is incident on the antenna over the range $\theta = \pm\pi/2$. The power dP' actually received by the antenna in small angle $d\theta$ is $P(d\theta/2\pi) \cos\theta$. Integrating,

$$P' = \frac{P}{\pi} \int_{-\frac{\pi}{2}}^{+\frac{\pi}{2}} \cos\theta \, d\theta = \frac{P}{\pi} [\sin\theta]_{-\frac{\pi}{2}}^{+\frac{\pi}{2}} = \frac{2P}{\pi} \, .$$

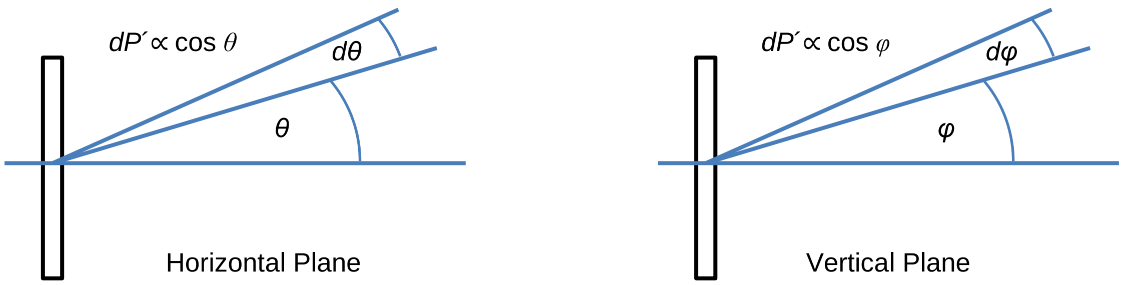

Figure 8.2: Response of a Loop Antenna in the Horizontal and Vertical Planes

Hence the fraction of the power received by the antenna is $2/\pi$. The same is true in the vertical plane: the sensitivity varies as $\cos\varphi$. In all, the fraction becomes $4/\pi^2$, or 0.405, or −3.9 dB. Our equation for E is then

$$E = F_a + 20 \log f_{MHz} + B - 100.5 \text{ dB}_{\mu V/m} \, .$$

At 1MHz, 20 log f_{MHz} (conveniently) equals 0, and B equals 39.5 (10 log 9,000). Hence,

$$E = F_a - 61.0 \text{ dB}_{\mu V/m} \, .$$

At that frequency, Antenna 'B' has a noise field of 4 $dB_{\mu V/m}$ — equivalent to an excess noise figure of 65 dB. A look at Figure 8.1 shows that the antenna is easily sensitive enough to measure man-made noise in a 'median business area', but not atmospheric noise at a quiet receiving site.

9. Comments and Conclusion

This report has looked at the calibration of ‘magnetic’ loop antennas, and — with luck — has demonstrated a simple and accurate way of determining antenna factors. With knowledge of the antenna factor and a measurement of the noise generated by the antenna, one can calculate an equivalent noise field, and hence the excess noise figure of the environment. There are some further points we should mention:

Accuracy It is normal good practice to include an estimate of errors and uncertainties when making measurements. This is hard to do with electromagnetic fields, since these are so easily affected by their environment, and there could be factors that are simply not known about. With the small-coil method, the known parameters such as the dimensions of the loop and coil can be measured accurately and, judging by the consistency of the results, it is reasonable to say that the antenna factor is within ±1 dB up to 10 MHz.

At higher frequencies, ‘anomalies’ occur, caused by such things as unwanted coupling between the generator and antenna output, in turn caused by connecting cables becoming resonant. Good earthing helps, but it is difficult to know for sure how great the remaining anomalies are. Again judging by consistency, the uncertainty can be supposed to increase from ±1 dB at 10 MHz to ±3 dB at 30 MHz. It is important to stress that good experimental practice is needed to achieve this.

E-field Sensitivity An ideal loop antenna would have no response to an electric field. The manufacturers of the antennas used as examples in this report claim good E-field immunity and hence rejection of much local (E-field) interference. This immunity arises from the loop feeding an amplifier with balanced inputs and high common-mode rejection.

There is a further reason for wanting E-field immunity. The small-coil calibration method (ideally) only generates an H-field, whereas broadcast transmissions give rise to both E- and H-fields. Any E-field sensitivity will invalidate the calibration if the antenna is used to measure such transmissions. An attempt was made to measure the E-field immunity by adapting the PTEM-cell. Removing the terminations at the ends of the transmission lines should, in principle, stop any current flowing in the lines, hence killing the H-field but leaving the E-field. Similarly, replacing the terminations with short-circuits should kill the E-field.

In practice, this method is imperfect. Even with the terminations removed, current still flows into the transmission lines, thanks to the lines’ capacitance to ground. Even an ideal antenna therefore appears to possess increasing E-field sensitivity with frequency. However, the test remains valid at low frequencies, and in this case we could say the ratio of H-field to E-field sensitivity exceeded 25 dB at 4 MHz. There was also no measurable difference in apparent antenna factor when swapping between ‘H-field-only’ and ‘E- and H-fields both present’.

Polarisation Illustrations of the antenna usually show it mounted upright; that is, with its axis parallel to the ground. This is fine for the reception of LF and MF transmissions, which are vertically polarised, but most HF transmissions are horizontally polarised. In theory, if the direct signal is being received from an HF transmitter, there will be no output from the antenna! In practice, the situation is unlikely to be as bad as that, since distant transmissions arrive courtesy of disorderly reflections from the ionosphere. However, it is worth bearing in mind.

Receiving Set-up This report has gone to some lengths to emphasise the need for a careful set-up when measuring antennas factors, if high-frequency ‘anomalies’ are to be avoided. In particular, the antenna needs a good earth connection, using a cable appreciably shorter than a quarter-wavelength at the highest frequency of interest. Similar precautions are needed at the receiving site if the measured antenna factors are to be valid.

10. References

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[6] ITU-R, 2022. 'Radio Noise.' Recommendation ITU-R P.372-16. (Figure 8.1 of this White Paper is taken from Figure 2 of ITU-R P.372-16.)

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[8] STANFORD UNIVERSITY, 2001. 'Noise in Antennas.' Noise in Antennas Stanford University https://web.stanford.edu › class › antenna_noise

[9] KRAUSS, J D, 1988. 'Antennas.' Mc Graw-Hill, ISBN 0-07-463219-1.

11. Glossary of Terms

AF	Antenna factor. The ratio of electromagnetic field strength (in volts per metre) to the output of the antenna (in volts). The unit is hence metre^{-1}. It is often expressed in dB/m ($20 \times \log_{10}$ of the value in metre^{-1}).
AM	Amplitude modulation.
LF	Low frequency (the radio spectrum existing between 30 kHz and 300 kHz).
MF	Medium frequency (the radio spectrum existing between 300 kHz and 3 MHz).
HF	High frequency (the radio spectrum existing between 3 MHz and 30 MHz).
RF	Radio frequency (any frequency suitable for radio transmissions).
$dB_{uV/m}$	The signal level expressed in dB relative to 1 uV per metre.
DRM	Digital Radio Mondiale.
EUT	Equipment under test.
EMF	Electromotive force. The output voltage of a generator in the absence of a load.
GTEM-cell	'TEM' stands for 'transverse electromagnetic'. A TEM-cell provides an electromagnetic field equivalent to the far-field of a transmitter, but within a small space. The 'G' stands for 'gigahertz', meaning that the cell is usable up to a number of gigahertz.
LNA	Low-noise amplifier.
PTEM-cell	'Pseudo-TEM-cell.' This is a poor man's TEM-cell — something that can be rigged up in any screened room. Its calibration is more problematic, and it is only likely to be usable up to the bottom end of the VHF band (below 100 MHz).
WPT	Wireless power transfer.