5.2 Theoretical analysis

In order to understand and obviously to optimise, the performance of a small loop antenna, its equivalent electrical circuit must be analysed from the point of view of those parameters which influence its performance as a radiator and receiver of electromagnetic energy.

First of all it is necessary to define what is meant by a small loop. Small, in the sense it is used here, refers nor necessarily to physical size but rather to electrical size. In other words the physical dimensions of the antenna are small compared with the wavelength of the electromagnetic signal it is handling. Put more rigorously, a small antenna is one whose maximum dimension is less than the "radian-length", where radian-length is defined as $1/2\pi$ wavelength (Wheeler, 1947). By virtue of this fact, the current in a small antenna is of constant amplitude throughout this maximum dimension, and leads in practice, to a circular loop having a maximum diameter of about $\lambda/10$ (Kraus, 1953).

Obviously a device as small as this would exhibit very poor efficiency as a radiator of electromagnetic energy since the radiation resistance, being as it is, proportional to the ratio squared of loop area to wavelength (Kraus 1953), will be very low. As an example, to put this into perspective, the radiation resistance of the maximum sized small loop (with diameter $\frac{\lambda}{10}$) is less than 2 ohms. Now at low frequencies, even $\lambda/10$ can be an unmanageable dimension for portable applications, therefore, this figure represents the highest value of radiation resistance which could be obtained. It will be noted that reference to radiation resistance in the context of dissipative media is misleading. The above example was presented, however, merely to illustrate the magnitudes involved and by so doing to shed some light on the complexity of the antenna problem.

5.3 The equivalent circuit

In representing the small loop as an electrical network we are dealing with a device which exhibits intrinsic inductance. The imperfections in the material of the loop, and losses due to the surrounding medium, will contribute a resistive component to the equivalent circuit. The ratio between the inductive reactance and this resistive term of the loop is the well known "quality factor" of reactive components, known as "Q". In the design of the optimum small loop antenna the relationship of these quantities to each other is of the utmost importance in order to achieve maximum efficiency.

A measure of the efficiency of a small loop antenna (indeed of any small antenna) at its "moment, M". In the case of the small loop device this is referred to as the magnetic moment, and is defined by:

$$M = N\ I\ A \qquad 5 - 1$$

where N = number of turns

I = loop current

A = loop area

The relevance of this expression is obvious if we consider the expression for the various field components generated by a small loop as presented in equations 4 - 16, 4 - 17 and 4 - 18. The quantity I dA, appearing in each, is none other than this magnetic moment, where N is one turn.

For an effective antenna it is thus necessary to strive for maximum moment.

Obviously to do this, any one, or all of the constituents of M can be increased thereby increasing M, however, there are limits as to the amount by which each may be increased. Obviously A is limited by the $\lambda/10$ diameter restriction placed on it by the definition of a small antenna. In addition to this, stray capacitance exists around any loop to its surroundings, and the effect of this is to cause the loop to resonate. Since these strays are unknown quantities, this effect is undesirable, and should, therefore, be eliminated by restricting the loop area. Figure 5.1 shows the experimental variation in self resonant frequency of a single turn loop, measured on the surface of the earth.

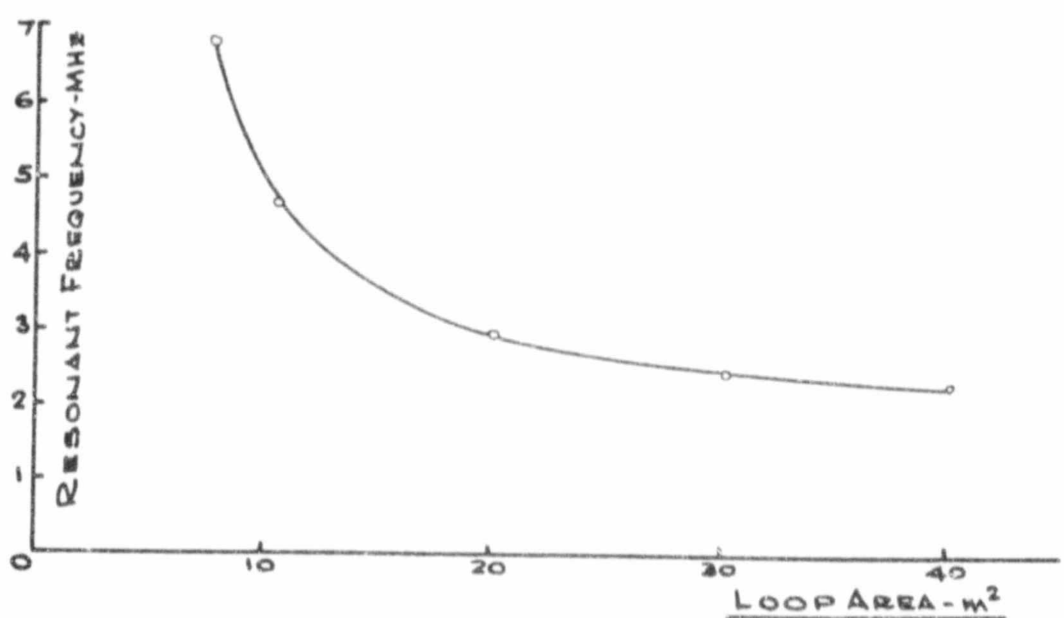


FIGURE 5.1

Variation in resonant frequency with loop area.

It will be noted from this figure that self-resonance, in this "above ground" case, only really becomes a problem at frequencies above 2 MHz. It could, however, be relevant if the technique, described in Chapter 4, were being used to determine the medium's electrical characteristics. It should be noted, however, that the loops in question resonate above ground when their diameters are approximately equal to λ/19.

Increasing the number of turns, N, in the loop would benefit M up to a point, beyond which the inherent loss resistance in the loop material would have increased such that the current would decrease, thereby not effecting the expected increase in M. In addition to this, the effect of the close proximity of one turn to another, in this multi-turn loop, introduces further loss due to the "proximity effect" (Smith, 1972). This is shown diagrammatically in Figure 5.2.

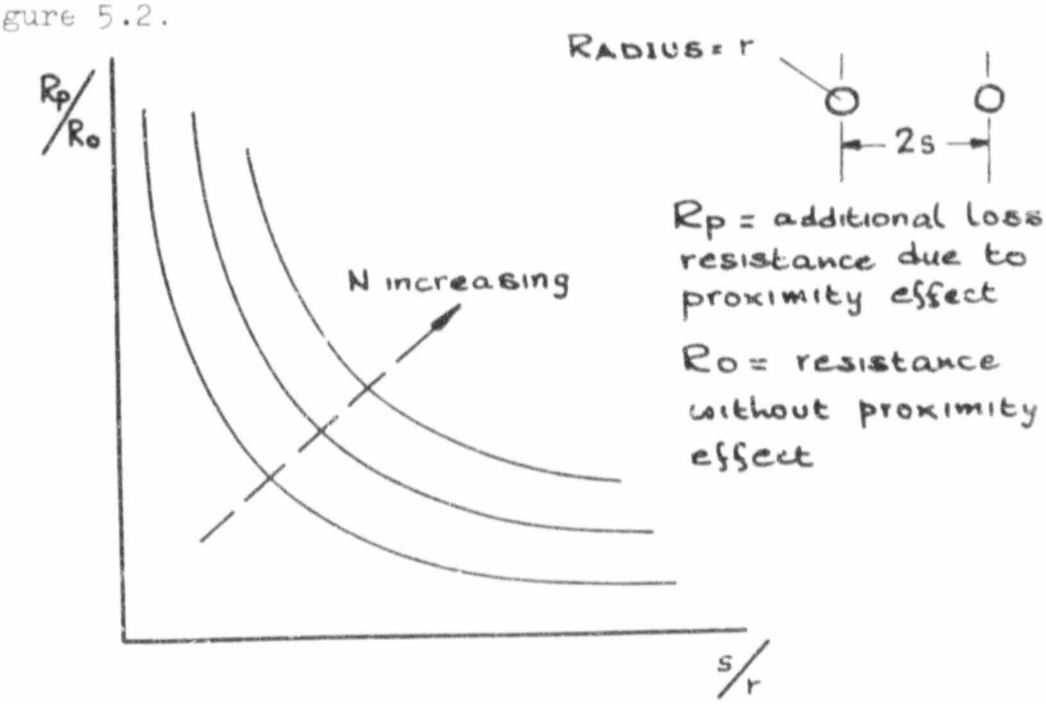


FIGURE 5.2

Effect of the "proximity effect" on the ohmic resistance of a multiturn loop (Smith, 1972).

It should thus be clear that optimum loop performance is not obtained simply by increasing either N or A, or both. A number of factors exist which are dependent on the loop parameters and these must be considered when designing a suitable antenna.

The expression for the magnetic moment M also contains a current term I, and so far no mention has been made of its derivation, whereas its effect on M is obvious. I is dependent on the active source connected to the loop which in this case would be a transmitter. From the previous discussion relating to total system loss it is to be expected that maximum power transfer between source and load would be required. This condition only obtains, of course, when source and load impedances are conjugately matched. In the case of the small loop antenna, as discussed so far, we are dealing with a lossy inductive circuit. It is thus necessary to resonate the loop and match its input impedance to the output impedance of the source.

The resonant, matched loop is shown in equivalent circuit form in Figure 5.3.

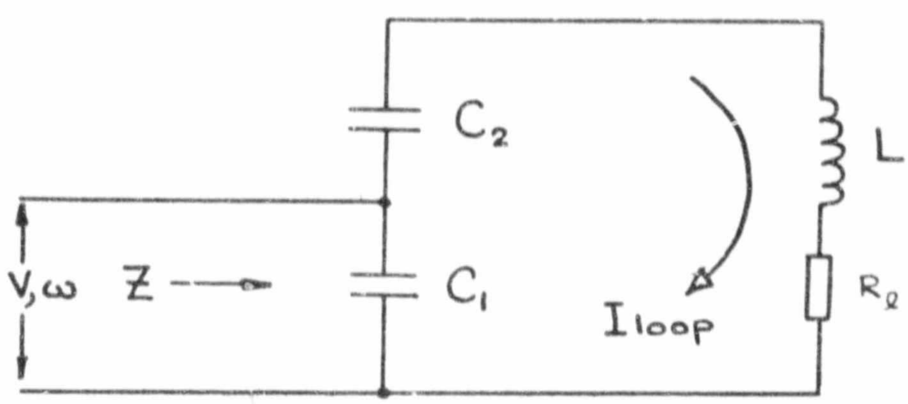


FIGURE 5.3

Equivalent circuit of the loop antenna.

Analysing this circuit (see Appendix B) for the condition of maximum current through the loop, with inductance L, and quality factor Q, we obtain

$$I_{loop} = \frac{VQ}{\omega L} \qquad 5 - $$

Obviously $Q = \frac{\omega L}{R_1}$, where R_1 is the inherent loss resistance of the loop, associated both with ohmic and skin resistance and the variety of other components, such as the previously described proximity effect, which contribute to this resistive component.

The effect of immersing the loop in a lossy medium adds a further resistive term to its impedance. This fact was covered in 2.8.4, where 2 - 23 gave the incremental change in loop impedance Δz, for the loop of radius "b" immersed in a cavity of radius "a" in a medium of conductivity σ.

$$\Delta z = \frac{(\mu\omega)^2 \ \sigma \ dA^2}{6\pi a} \qquad 5 - $$

obviously with σ being frequency dependent the loop current is affected by a change in frequency due to the effects of the medium and the parameters of the loop.

The loop is resonated basically by capacitor C_2, and is shunt matched to the required source impedance by C_1. Analysis of this circuit shows the existence of three poles and two zeroes, as shown in Figure 5.4.

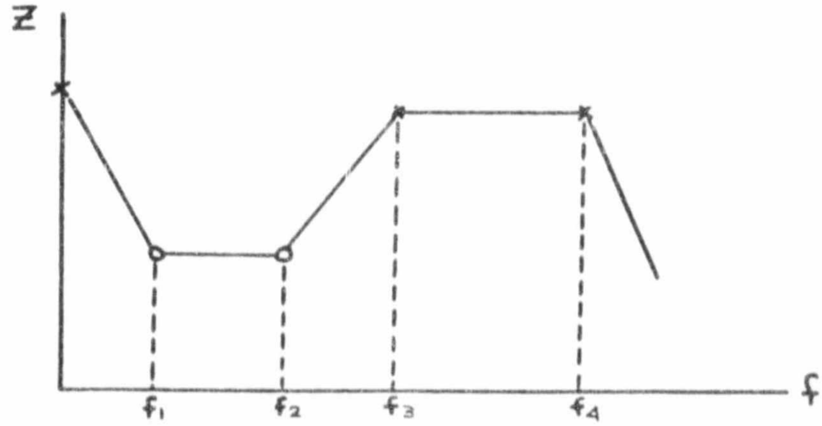


FIGURE 5.4

Pole - Zero diagram of the loop antenna

The high, "dynamic" impedance, Z_D, between f_3 and f_4 is used to obtain the resonating and matching constants as follows:

$$\left|\frac{Z}{Z_D}\right| = \left(\frac{C_2}{C_1 + C_2}\right)^2 \qquad 5 -$$

where $Z_D = \omega LQ$,

and the resonant frequency is given by

$$f_r = \frac{\sqrt{C_1 + C_2}}{2\pi\sqrt{L\,C_1 C_2}} \qquad 5 -$$

from which the resonating and matching capacitors are given by

$$C_1 = \frac{1}{4\pi^2 f_r^2 L}\sqrt{\frac{Z_D}{Z}} \qquad 5 -$$

$$C_2 = \frac{1}{4\pi^2 f_r^2 L\left(1 - \sqrt{\frac{Z}{Z_D}}\right)} \qquad 5 - 7$$

For a given loop geometry, L can be determined and Z_D measured, when this loop is resonated at the required operating frequency f_1. Thus from 5 - 6 and 5 - 7 the required matching impedance is determined by C_1.

If however, a single loop is to be used over a wide range of frequencies, as was the case in this study, for ease of practical operation, two basic problems arise. These will be discussed below:

A practical, portable loop might consist of a single turn, small device, which is easily manageable. This loop will have an inductance, determined by its geometry, which is low. Typically, a 0,5 m diameter loop of 10 mm diameter copper tubing has an inductance of nominally 1,3 μH, and a "Q" which varies with frequency as shown in the experimental results plotted in Figure 5.5.

Using 5 - 5 and 5 - 7, and values of Q obtained from Figure 5.5, we find that in order to resonate and match this loop to 50 Ω at 2 MHz requires C_1 = 0,042 μF and C_2 = 5 506 pF. In order to use the same loop at 6 MHz it is necessary to change both capacitors, with C_1 now being 9 953 pF and C_2 = 572 pF.

In practice it is found that the large values required for C_1 cause problems as the frequency is raised, since, in the example quoted here, only 0,07 μH of stray inductance is required to resonate with 9 953 pF at 6 MHz. Particular care is thus required in minimising all interconnecting lead lengths and this restriction immediately limits the "broadband" application of the loop due to the problem of mechanical switching which results.

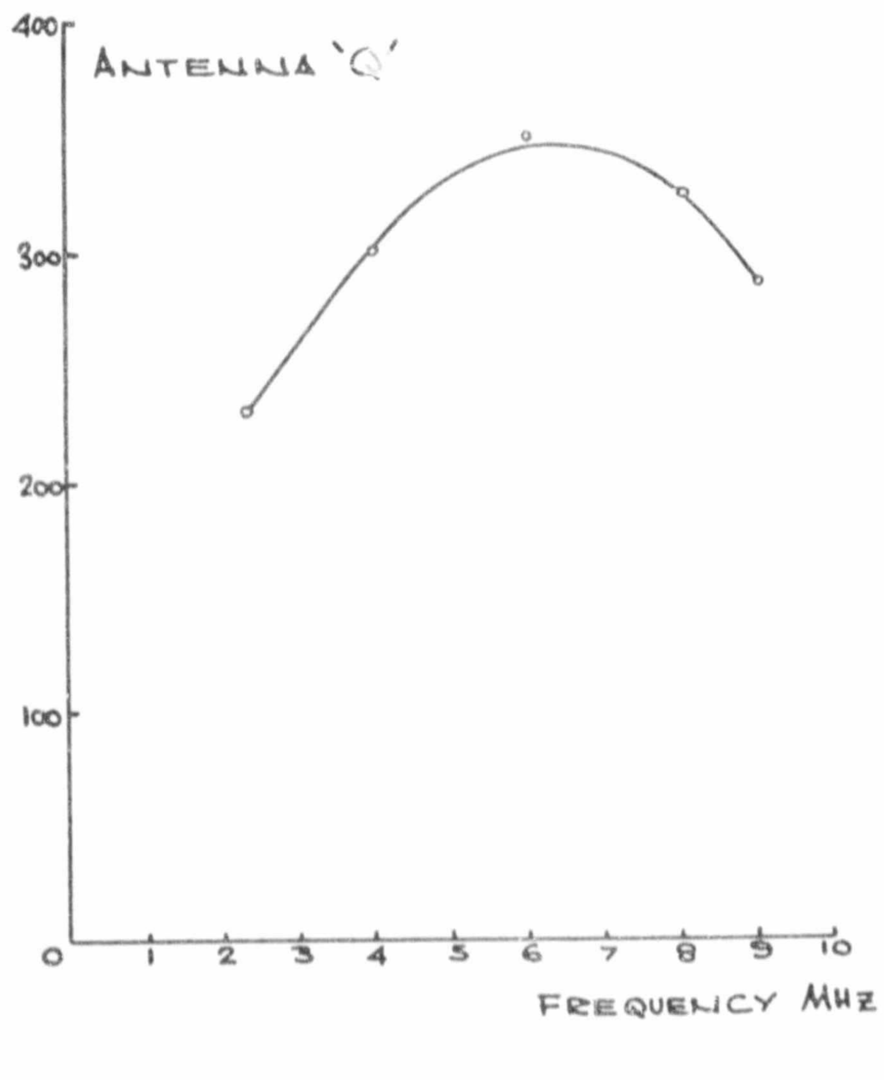


FIGURE 5.5

Variation of Q with frequency for a 0,5 m diameter loop using 10 mm copper tubing.

A single turn large loop, on the other hand, as well as being physically unmanageable, particularly for ease of orientation, suffers from the disadvantage, made apparent in Figure 5.1, that stray capacitance around the loop limits its upper frequency range and makes fine adjustment difficult since the value required for C_2 is reduced.

It was for these reasons that no single antenna was used at all frequencies discussed here, and hence that no direct comparison of antenna performance with frequency was obtained.

From the discussion in Chapter 2 on the effect of immersing an antenna in a dissipative medium it was made clear that the concept of radiation resistance of an antenna had no meaning in this context, unless one considered, as a basis for comparison, that all measurements were being made at the same distance from an elemental source in order that the effect of attenuation via the medium might be due only to α, the respective values of medium attenuation constant prevailing at each frequency and not also to variable "spreading losses." Bearing in mind the fact that the wavelengths within the medium were affected very definitely by the medium characteristics, this restriction is hardly practicable.

The definition of antenna efficiency applicable in free space, namely:

$$\eta = \frac{R_r}{R_r + R_l} \qquad 5 - 8$$

does not apply here. Since the surrounding dissipative medium has such a profound effect on the performance of the underground antenna due to the antenna's dependence on the characteristics of that medium, which have been shown to be variable with frequency, more work is required in order to be able to determine the variation in efficiency of a given antenna in this environment. Thus it is not possible here to predict an optimum frequency for communication since this requires a knowledge of this antenna efficiency which is not a simple function of frequency.

This latter point is of particular interest since it directly contradicts the statement made by Wadley (1949) that "the efficiency of the aerial is proportional to the cube of frequency." This was based on the classical result 5 - 8, where he considered in the first instance, a long wire antenna to be used, which, in free space, has a radiation resistance proportional to the square of the frequency, and that the loss resistance would decrease approximately inversely proportional to frequency, for rock of constant Q, where $Q = \frac{1}{p}$, which is reasonable. However, the free space definition of radiation resistance does not apply here, thereby invalidating this statement of Wadley's.

CHAPTER 6

CONCLUSIONS

Consideration has been given to the problem of using radio transmission techniques in order to communicate between points underground in a mine, which are separated by a mass of rock, and which have no metallic objects intervening either between them, or anywhere else within the area of interest.

This necessitated a study being made of the mine environment from the point of view of possible propagation paths, and the conclusion was soon reached that, in order to be of use under all circumstances, both in normal day to day activities and in emergencies, a mode of propagation, independent of air spaces such as tunnels, shafts, working places and the like, and of artificial transmission lines of any type, was necessary. The requirement was thus for "direct-through-rock" propagation.

In order to achive this, the characteristics of the propagation medium must be known and a study of these was made which brought to light the fact that the electrical properties of the intervening rock which affected the electromagnetic signal could not be obtained from "handbook" values, which were typical of laboratory measured quantities but rather required a knowledge of the in situ values.

It was also evident from a study of the relevant literature, that the method of launching electromagnetic energy into a dissipative medium was different from that used in the "free space" case, since the medium's electrical characteristics affected the antenna.

A large number of experimental results were obtained over a range of frequencies which, were felt, would provide the most likely typical operating area for the portable communication systems envisaged. This frequency range extended over four octaves and brought to light a number of interesting factors, namely the importance of correct antenna orientation and the significance of coaxial and coplanar orientations as regards inductive "power" versus radiated power; the fact that attenuation of the signal increased with frequency, basically as predicted, with some evidence being presented that intrusions of geological materials, having higher electrical conductivity than the host rock, such as a dyke, affect the propagation of electromagnetic energy by increasing attenuation, but also of causing this attenuation to be more constant over a range of frequencies, than is the case for the sedimentary rock.

Values for the attenuation constant were calculated from the measured signal attenuation results, and from these an estimate of host rock electrical conductivity was made by assuming a value for the dielectric constant or permittivity, since this parameter is far less frequency dependent than is conductivity, and the spread of typical values is considerably less. The values so obtained were an order of magnitude, or so, higher than those measured using typical samples in the laboratory.

The results obtained in this work basically confirmed those of other workers in this field, with a spread in values being expected from site to site due to the non-homogeneity of geological media, and the variation in stratification and faulting which occurs throughout the earth's crust. However, the accumulated evidence does confirm that effective propagation, via this lossy medium, will only occur at the lower end of the electromagnetic spectrum.

Considerably more work remains to be done on such areas as the effect of the medium on antenna performance, the effect of layering and stratification on signal propagation, and the effect of metal objects such as cables, pipes and rails as artificial transmission lines, which, many qualitative tests have shown, have highly beneficial effects on signal propagation.

The weak link in any system based on the results of this work will be at the portable radio transceiver end, and not at any fixed station which may be erected to act as a control, or as a base station in a multi-station system. This weakness is due to the fact that maximum portability must be ensured, meaning that antenna area is [illegible] This restriction on area, and that on the nu[illegible] of turns which may be employed in a given loop, due to stray capacitance and proximity effects, means that the loop current must be maximised. Peak power limitations exist in portable voice-modulated transmitters, hence the design criterion presented in this dissertation, relating loop Q and inductance must be adhered to in order to obtain maximum communication range.

No electrical noise, emanating from any external source whatsoever, was found over the frequency range investigated here. Receiver sensitivity based on this finding would be limited only by the noise generated internally by the equipment itself. This ideal condition will not apply in all areas, particularly in those served by power lines, hence consideration would have to be given to the selection of operating frequencies based on the limitations set by attenuation, antenna performance and noise, and a modulation technique which would provide optimum signal-to-noise ratio under these conditions is a necessity.

APPENDIX A

THE FIELDS DUE TO AN ELEMENTAL LOOP ANTENNA IMMERSED IN A DISSIPATIVE MEDIUM (WAIT, 1969)

The medium has characteristics given by the conductivity σ, the permeability μ, and the permittivity ϵ. The magnetic dipole source is considered to vary as $e^{j\omega t}$.

From Maxwell's fundamental equations we may write:

$$\text{curl } \underset{\sim}{E} = -j\mu\omega\underset{\sim}{H} \qquad \text{A - 1}$$

$$\text{curl } \underset{\sim}{H} = (\sigma + j\omega\varepsilon)\,\underset{\sim}{E} + \underset{\sim}{J} \qquad \text{A - 2}$$

where $\underset{\sim}{E}$ and $\underset{\sim}{H}$ are the vectors of electric and magnetic field intensity respectively, and $\underset{\sim}{J}$ is the current density vector, and $e^{j\omega t}$ is implied.

Now, introducing the magnetic vector potential B,

$$\text{where} \quad B = \frac{1}{4\pi j\mu\omega}\int_V \frac{\underset{\sim}{J}_m\, e^{-\gamma R}}{R}\, dV' \qquad \text{A - 3}$$

where $\underset{\sim}{J}_m$ is the magnetic current density vector and is due to the relationship between an assumed magnetic current J_m flowing through a linear element dl being related to the loop current I flowing around an elemental area dA, by

$$J_m\, dl = j\mu\omega I dA \qquad \text{A - 4}$$

The field components are

$$\underset{\sim}{E} = -j\mu\omega \text{ curl } B \qquad \text{A - 5}$$

$$\text{and} \quad \underset{\sim}{H} = \gamma^2 B + \text{grad div } B \qquad \text{A - 6}$$

the Hertz vector for this magnetic dipole is evaluated from A - 3 and A - 4, and is given by

$$B = \frac{J_m \, dl}{4\pi j\mu\omega} \frac{e^{-\gamma r}}{r} \hat{Z} \qquad A - 7$$

where the dipole axis is orientated in the Z direction and $\hat{Z}$ is the unit vector.

From A - 5, A - 6 and A - 7, then

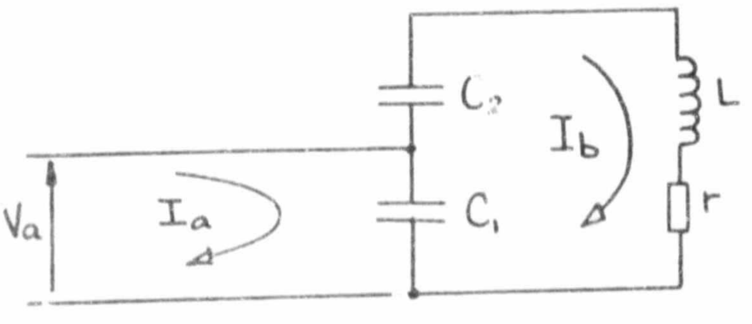
$$E_\phi = \frac{-j\mu\omega I dA}{4\pi r^2} (1 + \gamma r) e^{-\gamma r} \sin\theta$$

$$H_r = \frac{IdA}{2\pi r^3} (1 + \gamma r) e^{-\gamma} \cos\theta$$

$$H_\theta = \frac{IdA}{4\pi r^3} (1 + \gamma r + \gamma^2 r^2) e^{-\gamma r} \sin\theta$$

APPENDIX B

THE RELATIONSHIP BETWEEN LOOP CURRENT AND THE CHARACTERISTICS OF THE LOOP ANTENNA



$$V_a = \frac{1}{j\omega C_1}(I_a - I_b)$$

$$o = I_b\left(j\omega L + r + \frac{1}{j\omega C_1} + \frac{1}{j\omega C_2}\right) - \frac{I_a}{j\omega C_1}$$

$$\text{therefore } I_a = j\omega C_1\left(\frac{-\omega^2 LC_1C_2 + j\omega C_1C_2 r + C_1 + C_2}{j\omega C_1C_2}\right) I_b$$

substituting (2) into (1)

$$V_a = \frac{1}{j\omega C_1}\left\{\frac{j\omega C_1(-\omega^2 LC_1C_2 + j\omega C_1C_2 r + C_1 + C_2)\, I_b}{j\omega C_1C_2}\right.$$

$$= \frac{I_b}{j\omega C_1}\left\{\frac{C_1 + C_2 + j\omega C_1C_2 r - \omega^2 LC_1C_2}{C_2} - 1\right\}$$

at resonance $C_1 + C_2 = \omega^2 IC_1C_2$

$$\text{therefore } V_a = \frac{I_b}{j\omega C_1}\{j\omega C_1 r\}$$

$$= I_b r$$

where $r = \dfrac{\omega L}{Q}$

thus $V_a = \dfrac{I_b\,\omega L}{Q}$

giving $I_b = \dfrac{QV_a}{\omega L}$

APPENDIX C

THE EFFECT OF A CHANGE IN TRANSMITTER POWER ON COMMUNICATION RANGE

Consider the case where increased range is required, from R_1 to R_2, in a certain area of the mine, changing transmitter power output from P_{t_1} to P_{t_2} will bring this about, assuming that the required signal at the receiver remains constant. The operating frequency (hence λ) and antenna characteristics η and G are unaltered.

Therefore,

$$\frac{P_{t_2}}{P_{t_1}} = \left(\frac{R_2}{R_1}\right)^2 e^{-2\alpha(R_1 - R_2)} \qquad C - 1$$

now, if

$$P_{t_2} = n\, P_{t_1} \qquad C - 2$$

then, solving C - 1 for R_2 we find that

$$R_2 = \sqrt{n}\, R_1\, e^{-\alpha\, (R_2 - R_1)} \qquad C - 3$$

APPENDIX D

LEAST SQUARES FIT OF DATA AT 335 kHz

Examination of Figure 4.9 suggests a straight line variation in the coplanar horizontal component beyond about 45 m from the source. Thus we may write

$$y = a + bx \qquad \text{D - 1}$$

for $\ln V_r R = \ln K - \alpha R$, as given by 4 - 8. In the above straight line equation it is b which is of particular interest since it is the slope of the curve, and hence, from 4 - 8, is the attenuation of the medium.

Therefore, applying the least squares technique to the experimental data, we have

x	y	x^2	xy
49,7	3,04	2 470	151
57,3	3,01	3 283	172
64,9	2,93	4 212	190
72,6	2,94	5 271	213
80,2	2,81	6 432	225
87,8	2,85	7 709	250
95,4	2,82	9 101	269
103,1	2,79	1 063	288
$\Sigma x = 611$	$\Sigma y - 23,2$	$\Sigma x^2 = 49\ 108$	$\Sigma xy = 1\ 758$

$$\Delta_o = \begin{vmatrix} 8 & 611 \\ 611 & 49\ 108 \end{vmatrix} = 19\ 543$$

$$\Delta_1 = \begin{vmatrix} 611 & -23,2 \\ 49\ 108 & -1\ 758 \end{vmatrix} = 65\ 168$$

$$\Delta_2 = \begin{vmatrix} 8 & -23,2 \\ 611 & -1\ 758 \end{vmatrix} = 111,2$$

$$a = \frac{\Delta_1}{\Delta_0} = \frac{65\ 168}{19\ 543} = 3,33$$

$$b = \frac{-\Delta_2}{\Delta_0} = \frac{-111,2}{19\ 543} = -0,006$$

Now the above data was obtained from log VR, whereas it is here required at ln Vr, therefore, since

$$\log_e VR = \ln VR = \frac{\log VR}{0,4343}$$

we obtain

$$b = 0,014 \text{ nepers/m}$$

which is 0,12 dB/m.

REFERENCES

1. AMES, L.A., DE BETTENCOURT, J.T., FRAZIER, J.W. and ORANGE, A.S.: Radio Communication via Rock Strata. IEEE Trans. Communications, CS11: pp 159-169 (1963).

2. AMES, L.A., FRAZIER, J.W. and ORANGE, A.S.: Geological and Geophysical conditions in radio propagation through the earth's crust. IEEE Trans. Ant. and Prop. AP-11: pp 369-371 (1963).

3. BRACE, W.F., ORANGE, A.S. and MADDEN, T.R.: The effects of pressure on the electrical resistivity of water saturated crystalline rocks. J. Geophysical Res. 70,22: pp 5669-5678 (1965).

4. BRACE, W.F. and ORANGE, A.S.: Further studies of pressure on electrical resistivity of rocks. J. Geophysical Res. 73,16: pp 5407-5420 (1968).

5. COLLETT, L.S. and KATSUBE, T.J. Electrical parameters of rock in developing geophysical techniques. Geophysics 38,1: pp 76-91 (1973).

6. COOK, J.C.: R.F. Electrical Properties of bituminous coal samples. Geophysics 35,6: pp 1079-1085 (1970).

7. DELOGNE, P.: Basic mechanisms of tunnel propagation. Radio Science, 11: pp 295-303 (1976).

8. EMSLIE, A.G. and LAGACE, R.L.: Propagation of low and medium frequency radio waves in a coal seam. Radio Science, 11,4: pp 253-261 (1976).

9. FRITSCH, V.: Propagation of radio frequency electromagnetic fields in geological conductors. J. Research Nat. Bur. Stds. - D. Radio propagation 67 D,2: pp 161-178 (1963).

10. GABILLARD, R., Degaque P. and WAIT, J.R.: Subsurface Electromagnetic Telecommunication - A Review. IEEE Transactions on Communication Technology COM 19,6: p 1217-1228 (1971).

11. GEYER, R.G. and KELLER, G.V.: Constraints affecting through-the-earth electromagnetic signalling and location techniques. Radio Science, 11,4: pp 323-342 (1976).

12. HOWELL, B.F. and LICASTRO, P.H.: Dielectric behaviour of rocks and minerals. The American Mineralogist 46: pp 269-288 (1961).

13. IEEE Std. 356-1974. IEEE Guide for Radio Methods of Measuring Earth Conductivity, AP-22,2: pp 375-400 (1974).

14. JORDAN, E.C.: Electromagnetic Waves and Radiating Systems. Prentice-Hall Inc. New Jersey, (1964).

15. KELLER, G.V. and LICASTRO, P.H.: Dielectric constant and electrical resistivity of natural-state cores. U.S. Geological Survey Bulletin 1052 (H): pp 257-285 (1959).

16. KELLER, G.V.: Electrical properties in the deep crust. IEEE Trans. Ant. and Prop. AP-11: pp 344-357 (1963).

17. KITAHARA, R. and TAKAGI, H.: Propagation of radio frequency electromagnetic waves in rock strata. Mining and Safety (Japan) 19,7: SMRE Translation 6336, (1973).

18. KRAICHMAN, M.B.: The impedance of a circular loop in an infinite conducting medium. J. Res. National Bureau of Standards - D Radio Propagation 66D: pp 499-503 (1962).

19. KRAUS, J.D.: Electromagnetics McGraw-Hill, (1953).

20. LAGACE, R.L. and PARKINSON, H.E.: Two-way Communications with roving miners. U.S. Bureau of Mines Information Circular IC 8365, (1972).

21. MARTIN, D.J.R.: Radio Communication in Mines and Tunnels. Electronic Letters, 6:18: pp 563-564 (1970).

22. MOORE, R.K.: Effects of a surrounding conducting medium on antenna analysis. IEEE Trans. Ant. and Prop. AP-11: pp 216-225 (1963).

23. PARKHOMENKO, E.I.: Electrical Properties of Rocks. Plenum Press, (1967).

24. RYAYANTSEV, A.M. and SHABEL'NIKOV, A.V. Propagation of radio waves through the earth's crust. Radio Engineering 11: pp 1643-1657 (1965).

25. SCOTT, J.H., CARROLL, R.D. and CUNNINGHAM, D.R.: Dielectric constant and electrical conductivity measurements of moist rock: A New Laboratory Method. J. Geophysical Res.72,20: pp 5101-5115 (1967).

26. SMITH, G.S.: Electrically small and insulated loop antennas. Ph.D thesis, Harvard University, Cambridge, Massachussetts, (1972).

27. SMITH, G.S. and KING, R.W.P.: The resonant linear antenna as a probe for measuring the in situ electrical properties of geological media. J. Geophysical Res, 79, 17: pp 2623-2628 (1974).

28. TERMAN, F.E.: Electronic and Radic Engineering. 4th ed. McGraw-Hill (1955).

29. VAN ZIJL, J.S.V.: Head, Geophysics Division, S.A. C.S.I.R. Private personal communication. (1974).

30. WADLEY, T.L.: Radio Communication through rock on the Witwatersrand mines. NITR, ETR-4: pp 1-16, Aug. (1949).

31. WAIT, J.R.: The magnetic dipole antenna immersed in a conducting medium. Proc IRE 40: pp 1244-1245 (1952).

32. WAIT, J.R.: Insulated loop antenna immersed in a conducting medium. J. Res. National Bureau of Standards, 59: pp 133-137 (1957).

33. WAIT, J.R. Electromagnetic Waves in Stratified Media, Pergamon, New York (1962).

34. WAIT, J.R. Electromagnetic fields of sources in lossy media, in Antenna theory Pt II, edited by Collin R.E. and Zucker, F.J., McGraw-Hill pp 438-514 (1969).

35. WHEELER, H.A.: Fundamental Limitations of Small Antennas. Proc. I.R.E. 35: pp 1479-1484, (1947).

36. WILLIAMS, R.H.: Insulated and loaded loop antennas immersed in conducting media. J. Res. National Bureau of Standards - USNC/URSI 69D: pp 287-289, (1965).

37. WUNDT, R.M. and BOOTS, D.A. Underground Communications. IEEE Trans. Comm. and Electronics 71: pp 137-142 (1964).

38. ZIGGH, M.E. Modes in lossy stratified media with application to Underground Propagation of Radio Waves. IEEE Trans. on Ant. and Prop. AP-11: pp 318-323 (1963).

/MS