TABLE 2.4

LOSS TANGENT VARIATION WITH FREQUENCY FOR QUARTZITE

FREQUENCY (kHz)	LOSS TANGENT (p)
100	0,60
300	0,40
1 000	0,24
3 000	0,20

The condition $p \leq 0{,}6$ applies, and so the attenuation constant α is obtainable from

$$\alpha = \frac{1{,}64 \times 10^3 \times \sigma}{\sqrt{\epsilon_r}} \cdot \text{dB/m}$$

If the condition $p \geq 10$ were to apply then, assuming $\epsilon_r = 6$, as before, the value of σ necessary at 300 kHz would be greater than or equal to 1×10^{-3} S/m, or 25 times larger than that quoted by Wadley. Of particular interest here are the values of conductivity measured by Van Zijl (1974) from deep electrical soundings carried out near Western Deep Level gold mine in the Western Transvaal. He used d.c. sounding techniques and obtained values of conductivity from 1×10^{-3} S/m to 2×10^{-4} S/m for Witwatersrand quartzites. It is significant that these values were obtained at d.c.

The attenuation through the rock given by Wadley's (1949) laboratory value, as used above, would be 0,03 dB/m, whereas if the conductivity were 1×10^{-3} S/m, it would be 0,3 dB/m, a ten-fold increase.

A practical example quoted by Kitahara and Takagi (1973) is also of interest here. They were dealing with wet earth having the following characteristics:

$$\sigma = 4 \times 10^{-3} \text{ S/m}$$

$$\epsilon_r = 10$$

at a frequency of 100 kHz. The value of the loss tangent in this case is

$$p = \frac{\sigma}{\omega\epsilon_o\epsilon_r} = \frac{4 \times 10^{-3}}{2\pi \times 10^5 \times 8,85 \times 10^{-12} \times 10} \doteq 72$$

Therefore, since $p > 10$, the attenuation is given by:

$$\alpha = 0,545 \sqrt{100 \times 4 \times 10^{-3}} \text{ dB/m}$$

i.e. $\alpha = 0,34$ dB/m.

2.8 The antenna in a lossy medium

From the considerations discussed in the previous sections of this dissertation it would appear that a radio communication system operating underground depends very largely, for its effectiveness, on the use of low frequencies. This is obviously a relative term, however, it would be correct, at this juncture, to say that the lower the frequency the better! From the point of view of attenuation of the electromagnetic signal as it propagates through the rock, this appears reasonable. However, the practical aspects of a typical communication system, suitable for mining use, introduce other factors which modify this situation. The first, and most important of these, is equipment portability. No elaborate detail will be discussed in this regard, but suffice it to say that in a basically hostile environment, performing physically taxing work, and, in many instances, such as fire-fighting and rescue operations, a miner is already encumbered with equipment, such that the physical size of any additional apparatus affects his comfort, performance and particularly, safety, very drastically.

This size restriction imposes a very real problem on the design of the radio communication system, in as much as it involves both a power limitation in the transmitter, due to limited battery size and hence capacity, and more particularly since it necessitates the use of portable antennas (for portable apparatus). It is now clear, from the previous review of the propagation path, that frequencies possibly less than 1 MHz are envisaged in this application. In order to gain some insight as to what this means in terms of antenna size one only needs consider the free-space condition where, at that frequency, the wavelength is 300 m, and for efficient radiation of power an antenna of at least an eighth of a wavelength or 37,5 m, is required. Obviously the use of lower frequencies (for less attenuation) would require proportionately longer antennas.

2.8.1 Velocity of propagation in a lossy medium

Since the wavelength has a direct bearing on the length of an antenna, and is itself proportional to the velocity of energy propagation, it is this latter quantity which must be examined for the case of propagation via a lossy medium.

It can be shown (Ames and de Bettencourt, 1963), that the wavelength λ, in a lossy medium, characterised by a relative permittivity ϵ_r and a loss tangent p, is given by:

$$\lambda = \frac{\lambda_o}{\sqrt{\epsilon_r}\, f(p)} \qquad 2 - 14$$

where λ_o is the free space wavelength

and $f(p)$ is defined by 2 - 8.

This expression can be related (ibid) to the two conditions of loss tangent commonly encountered, viz. $p \leq 0,6$ and $p \geq 10$, as follows:

for $p \leq 0,6$ $\qquad \lambda = \frac{\lambda_o}{\sqrt{\epsilon_r}}$ $\qquad$ 2 - 15

and for $p \geq 10$ $\qquad \lambda = 2\pi\tau$, where τ is the metallic skin depth.

$$\tau = \sqrt{\frac{2}{\omega\mu_o\sigma}} = \frac{15,92}{\sqrt{f(kHz)\sigma}} \text{ m} \qquad 2 - 16$$

therefore $\qquad \lambda = \frac{100}{\sqrt{f(kHz)\sigma}}$ m $\qquad$ 2 - 17

The length of the antenna in the lossy medium will thus obviously be less than its counterpart, at the same frequency, in free space, and at 1 MHz this amounts to a reduction in length of 59 per cent, or yielding a length of about 15 m. This value was based on the condition $p \leq 0,6$ and $\epsilon_r = 6$.

If the medium conditions are such that $p \geq 10$ applies then the antenna length would be about 12,5 m for an assumed medium conductivity of 10^{-3} S/m. In neither case do these constitute antenna lengths suitable for portable operation. Hence the antenna will, of necessity have to be considerably less than an eighth of the wavelength in the medium, with attendant decrease in efficiency.

2.8.2 Antenna efficiency

All discussion, until this point, on the behaviour of an antenna operating in a lossy medium has, except for the modification in wavelength, been based on the free space considerations of the meaning of antenna efficiency. It will be shown in the text which follows that this is not entirely valid. It is, however, still perfectly true that the antenna dimensions, as derived above, are fully representative of the lossy medium condition.

The concept of antenna efficiency is fundamental to the understanding of any antenna. It is defined as the ratio of the power radiated by the antenna to the power supplied to it. From this definition a simple equivalent circuit, Figure 2,4, results, wherein the resistive components responsible for power radiation and dissipation are shown.

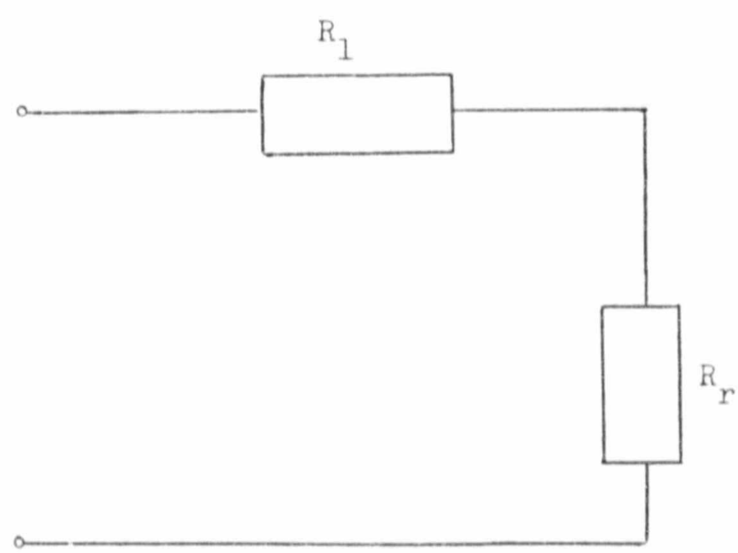

FIGURE 2.4

THE EQUIVALENT CIRCUIT OF AN ANTENNA IN FREE-SPACE

where R_1 - antenna loss resistance.

R_r - antenna radiation resistance.

In terms of these quantities, the radiation efficiency of an antenna is given by

$$\eta = \frac{1}{1 + Rl/R_r} \qquad 2 - 18$$

The relationship between R_1 and R_r is then vital to the success of the antenna's performance as an efficient radiator (and hence, receiver, due to reciprocity) of electromagnetic energy.

2.8.3 Radiation resistance

The radiation resistance of an antenna in free space is computed by considering the antenna to be surrounded by a sphere of radius R, and to be energised by r.m.s. current I. Dividing the average flow of power P through this sphere by I^2 yields the radiation resistance. Since the power which crosses concentric spherical surfaces is constant, any radius will yield the value of the radiation resistance.

An antenna surrounded by a dissipative medium presents a totally different picture, however, since the condition of energy storage within the near field of the antenna in a lossless medium no longer applies, and energy is in fact dissipated within the immediate vicinity of the antenna. The concept of radiation resistance, as defined for the antenna in a lossless medium, therefore, has no meaning for an antenna in a dissipative medium. (Moore, 1963).

From the arguments presented previously it is reasonable to expect that antennas, small compared with the wavelength, must be used. This being the case, the method of analysis used is that for elemental antennas. Two basic elemental antennas exist, namely the electric dipole and the magnetic dipole. It might reasonably be expected that they would perform equally well as radiators and receivers in a lossy medium. However, analysis by Moore (1963) and Wait (1969) show this not to be the case.

Considering a current element of length dl and carrying a current I, the total power crossing an enclosing sphere of Radius R is given by P_e where

$$P_e \doteq \frac{(Idl)^2}{12\pi\sigma R^3} \qquad 2 - 19$$

σ being the conductivity of the surrounding medium.

For a magnetic dipole consisting of an elemental loop of area dA carrying a current I, immersed in a medium as above, then the total power crossing the sphere is P_m

$$P_m \doteq \frac{\sigma\mu^2\omega^2(IdA)^2}{12\pi R} \qquad 2 - 20$$

(Wait, 1969)

There is a fundamental difference between these two expressions and that is that the power dissipated near the magnetic dipole increases far less rapidly than that from the electric dipole, (Moore, 1963). This is due to the fact that the impedance near the magnetic dipole, due to the radial wave, is largely imaginary, whereas that for the electric dipole is largely real. (Wait, 1969).

From the foregoing it has been shown that the analysis of antenna performance in a dissipative medium requires the application of a different technique to that in the free space case. Even the fundamental concept of radiation resistance breaks down. It should be pointed out at this stage that this applies even to slightly conducting media, in order words, where low values of loss tangent mightapply. (Moore, 1963).

On further examination of equations 2 - 19 and 2 - 20 it is apparent that as R tends to zero, an infinite power is required to produce a finite field at a distant point. It is in fact not justified to let R tend to zero (Wait, 1969) since the antenna itself is of finite size. However, the question of insulating the antenna immediately arises. If the dipole is placed in a spherical cavity of radius a, having permittivity ϵ_1, conductivity σ_1 and permeability μ_1, and the whole is immersed in a dissipative medium with properties ϵ_2, σ_2 and μ_2, it can be shown that the total power P, crossing the walls of this cavity from an elemental loop of area dA with current I is given by Wait (1952) as:

$$P \doteq \frac{(IdA)^2 \ (\mu\omega)^2 \ \sigma_2}{12\pi a} \qquad 2 - 21$$

This expression is based on the assumption that the loss tangent of the medium surrounding the cavity is high, which condition ensures that the conduction current exceeds the displacement current. This restriction means, therefore, that the outer medium is highly conducting and that the frequency is sufficiently low. Whether this condition will be satisfied in the South African environment remains to be seen. It is immediately clear, however, that if reliance is placed on laboratory measured values of rock conductivity and permittivity at frequencies of interest in this study, then it will not be satisfied.

However, considering 2 - 21 again, it may be concluded that an insulating cavity, enclosing the magnetic dipole antenna, specifies the amount of power required to maintain a given field in the exterior region. In deriving 2 - 21 (Wait, 1952) assumed the loop to be small compared to the dimensions of the cavity. In a later paper Wait (1957) removed this restriction and analysed the situation again, yielding, interestingly enough, the same result, namely:

$$P \doteq \frac{(IA)^2 (\mu\omega)^2 \sigma_2}{12\pi a}$$

where the elemental loop of area dA has been replaced by a finite loop of area A. It is thus concluded that the power radiated from the loop, for a specified current, varies approximately as the reciprocal of the cavity diameter, and is not modified to any extent by the size of the loop within that cavity.

2.8.4 <u>The impedance of the loop</u>

In order to establish the current I in the loop, a knowledge of the loop impedance Z, is required. From Wait (1957) we proceed as follows:

If the free space impedance of the loop is Z_o, then the impedance of the loop in the cavity is

$$Z = Z_o + \Delta Z \qquad 2 - 22$$

Where ΔZ is the incremental change in impedance due to the finite size of the cavity. From 2 - 21, therefore, ΔZ is given by

$$\Delta Z = \frac{(\mu\omega)^2 \sigma \, dA^2}{6\pi a} \qquad 2 - 23$$

If the loop is not considered to be small compared to the diameter of the cavity, then the above expression for ΔZ is multiplied by the factor

$$\left[1 + \frac{1}{15}\left(\frac{b}{a}\right)^2 + \frac{1}{70}\left(\frac{b}{a}\right)^4 + \ldots\right]$$

where b is the loop radius, and in the limit when b = a the loop impedance is increased by approximately 8 per cent over the value when $b \ll a$.

Kraichman (1962) considered a circular loop of thinly insulated wire, carrying a uniform current I, and immersed in a conducting medium. His approach thus differed from Wait's in that he did not assume that the loop was contained within a spherical insulating cavity, but rather is merely separated from the surrounding medium by the layer of insulation. He derived an expression for the immersion correction of the loop impedance, and compared it with that derived by Wait (1957), i.e. 2 - 23, for the case when loop radius and cavity radius were equal. The ratio of Kraichman's to Wait's result was approximately 0,86. This close agreement indicates, according to Kraichman, that the spherical insulating cavity has only a small effect on the loop impedance.

Williams (1965) considered a small insulated loop in a sphere or radome, but loaded by a permeable core with relative permeability μ_r. By so doing he was able to show that the effective area of the loop was increased by a factor

$$(1 + K_1)$$

where $$K_1 = \left[\frac{2(\mu_r - 1)}{2(\mu_r + 1) + 1}\right] \left(\frac{c}{b}\right)^3 \qquad 2 - 24$$

with c = radius of permeable core,

and b = radius of loop.

The implication is then that the loop must be tightly wound on the permeable core.

It is thus concluded that in order to radiate energy into a surrounding dissipative medium an antenna must be insulated from that medium. The thickness of this insulating layer, so necessary in establishing radiation, is inversely proportional to the amount of power radiated. A spherical insulating core within the antenna plays a minor role in influencing antenna impedance, whereas a permeable core, on which the loop must be tightly wound, increases the effective area and hence the radiated power for a given excitation current.

CHAPTER 3

MEASURED VALUES OF ROCK ELECTRICAL CHARACTERISTICS IN THE LABORATORY

3.1 Introduction

In order to gain some insight into the behaviour of rock as a medium of propagation for electromagnetic signals, the electrical characteristics of conductivity and permittivity, along with permeability, must be known in the light of the influence they have on this propagation. In Chapter 2 the limitations on the usefulness of laboratory determined values was discussed, however, it was felt useful to gain first-hand knowledge of the range of values expected from samples of rocks typical of South African mining conditions. It must be pointed out at this point that the rocks measured in this study are not restricted to those common only to gold mining areas, but are representative of types found in a variety of mineral rich areas, and hence give an overall view of rock electrical characteristics as determined in the laboratory.

3.1.2 Measurement procedure

Of the quantities to be determined, the relative permeability of rock may be considered, for all practical purposes, to be equal to unity. In other words the medium is non-magnetic and exhibits the permeability of free space, equal to $4\pi \times 10^{-7}$ H/m.

The other quantities, conductivity σ and relative permittivity ϵ_r, are related to one another by the loss tangent, p, of the medium.

Thus $p = \frac{\sigma}{\omega \epsilon_0 \epsilon_r}$, which is a dimensionless quantity, expresses very succinctly, the medium's performance as conductor, quasi-conductor or insulator.

The method of measurement adopted in this study to determine σ and ϵ_r shown in Figure 3.1, was based on a technique which provides ϵ_r and ρ directly, and then, by means of the above relationship, the value of σ could be obtained. Values were obtained for a number of frequencies over the range 300 kHz to 10 MHz using a dielectric loss test jig (Marconi TJ 1558/1) in conjunction with a circuit magnification, or "Q" meter (Marconi TF1245), which was driven by a signal generator (Marconi TF 1246).

3.1.3 The samples

To minimise, as much as possible, the possibility of experimental error, the samples were cut in thin (2 - 3 mm) slices, and projected slightly beyond the edges of the electrodes of the test jig to minimise the effects of fringing.

All the samples used were dry, showing no signs of moisture content at all, and were of good quality, in that obviously cracked or fractured specimens were rejected. In addition, more than one example of each rock type was measured in case an anomaly existed in any which might have been taken to be representative of that group.

3.1.4 The measured results

Experimental results are presented for conductivity σ, permittivity ϵ_r and loss tangent for various rock types.

Figure 3.2 shows the variation in conductivity with frequency. As expected, conductivity, in every case, increases

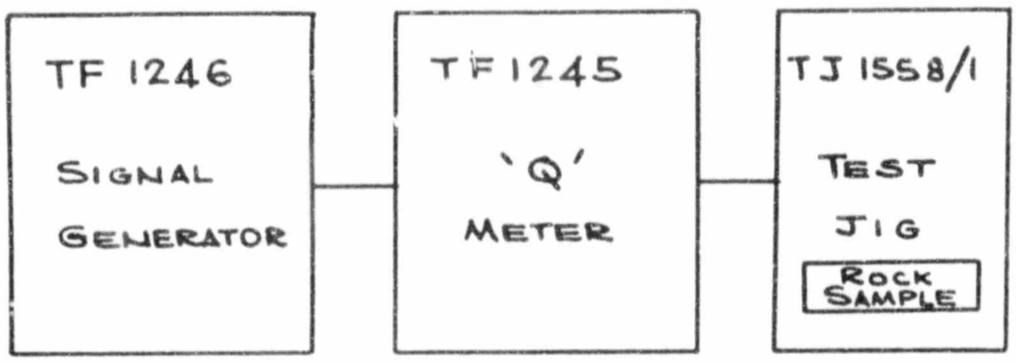

FIGURE 3.1

The laboratory method for measuring the conductivity and permittivity of rock samples.

with frequency, and the spread in values, over the frequency range, from, of the order of 6×10^{-6} S/m to 3×10^{-4} S/m is in good agreement with the values quoted in the literature for samples measured under similar conditions, (Parkhomenko, 1967).

Figure 3.3 presents the measured values of relative permittivity ϵ_r for the same samples as appeared in Figure 3.2. Noteworthy here is the fact that the values obtained hardly vary with frequency and in this respect coincide with those quoted in the literature for samples with low moisture content, (ibid).

The values of loss tangent presented in Figure 3.4 are particularly interesting since they illustrate, only too vividly, the difference between laboratpry and expected <u>in situ</u> conditions. Loss tangent decreases somewhat with frequency, in agreement with the results appearing in the literature, but of particular importance is the fact that no sample measured exhibited a value of p greater than 0,6. In other words, if propagation predictions were made, based on these values, the medium would be considered as one where displacement currents dominated and where the attenuation would be given by 2 - 11,

i.e. $$\alpha = \frac{1,64 \times 10^3 \sigma}{\sqrt{\epsilon_r}} \text{ dB/m.}$$

Practical measurements of α in typical mining environments do not support this conclusion, as will be shown later.

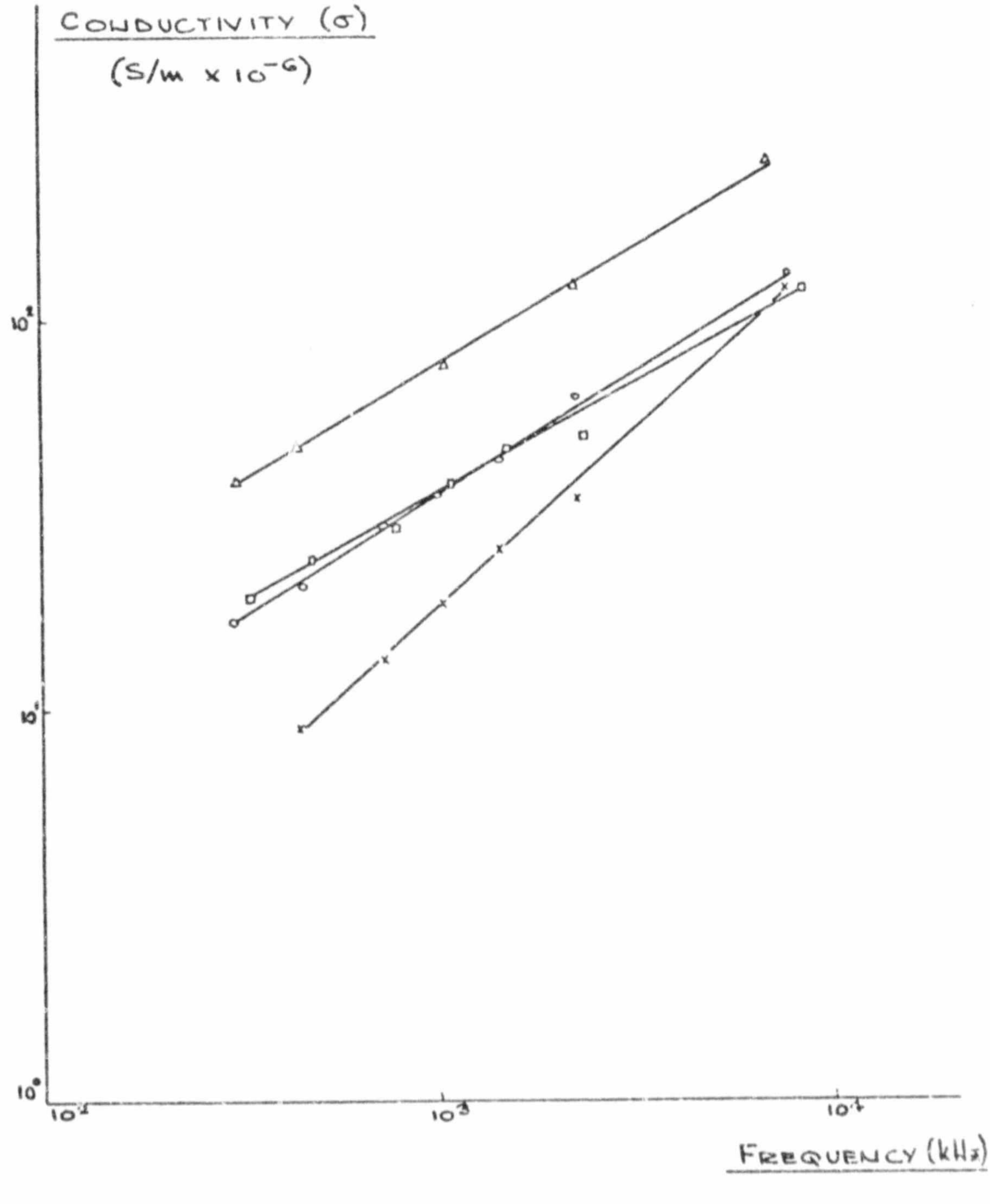

FIGURE 3.2

Variation in conductivity with frequency for pyroxemite (Δ) - Rustenburg platinum mine, quartzitic conglomerate (O,□) - Durban Deep gold mine, anorthosite (X) - Rustenburg platinum mine.

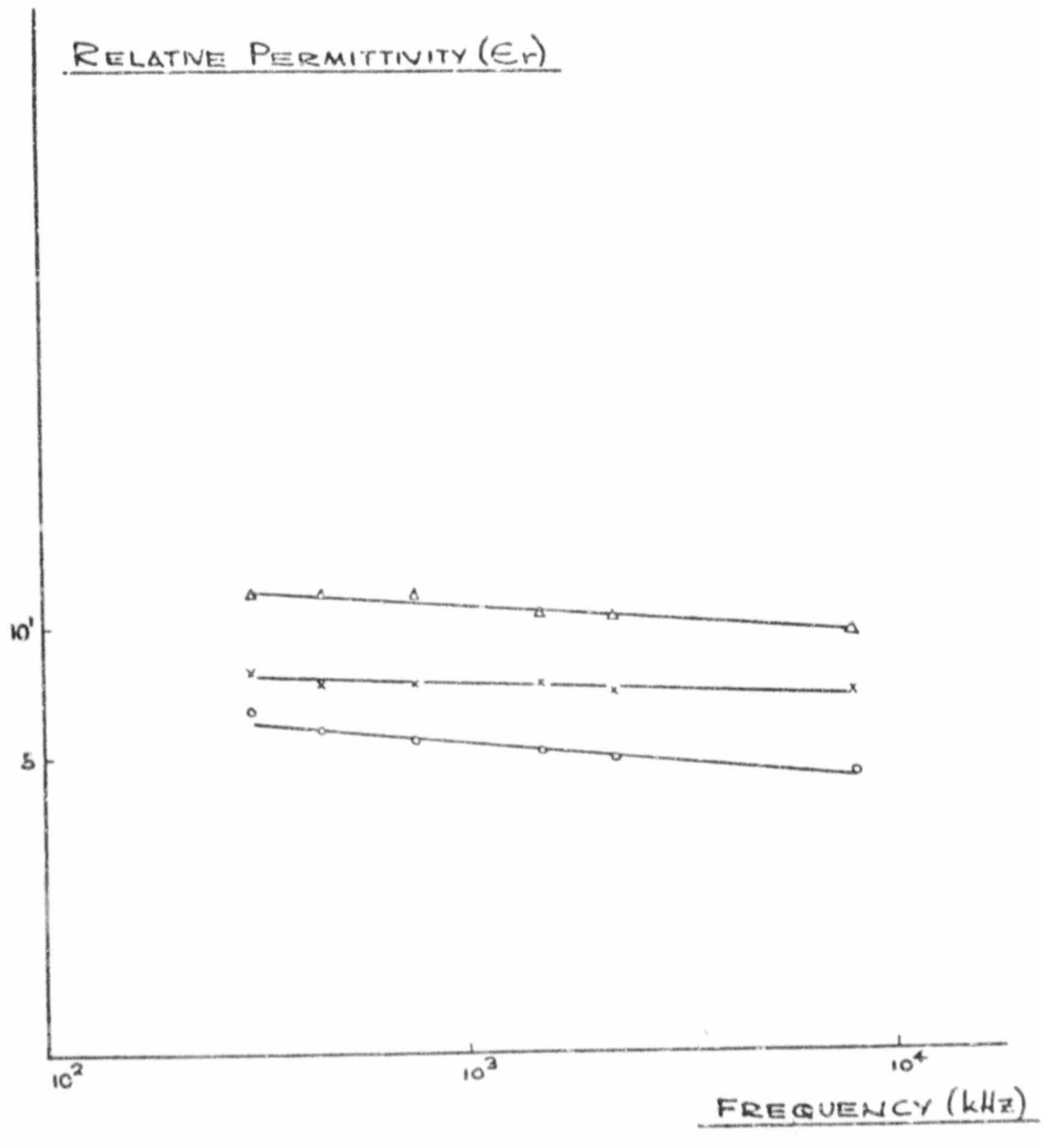

FIGURE 3.3

Variation in relative permittivity with frequency for the same samples as in Figure 3.2.

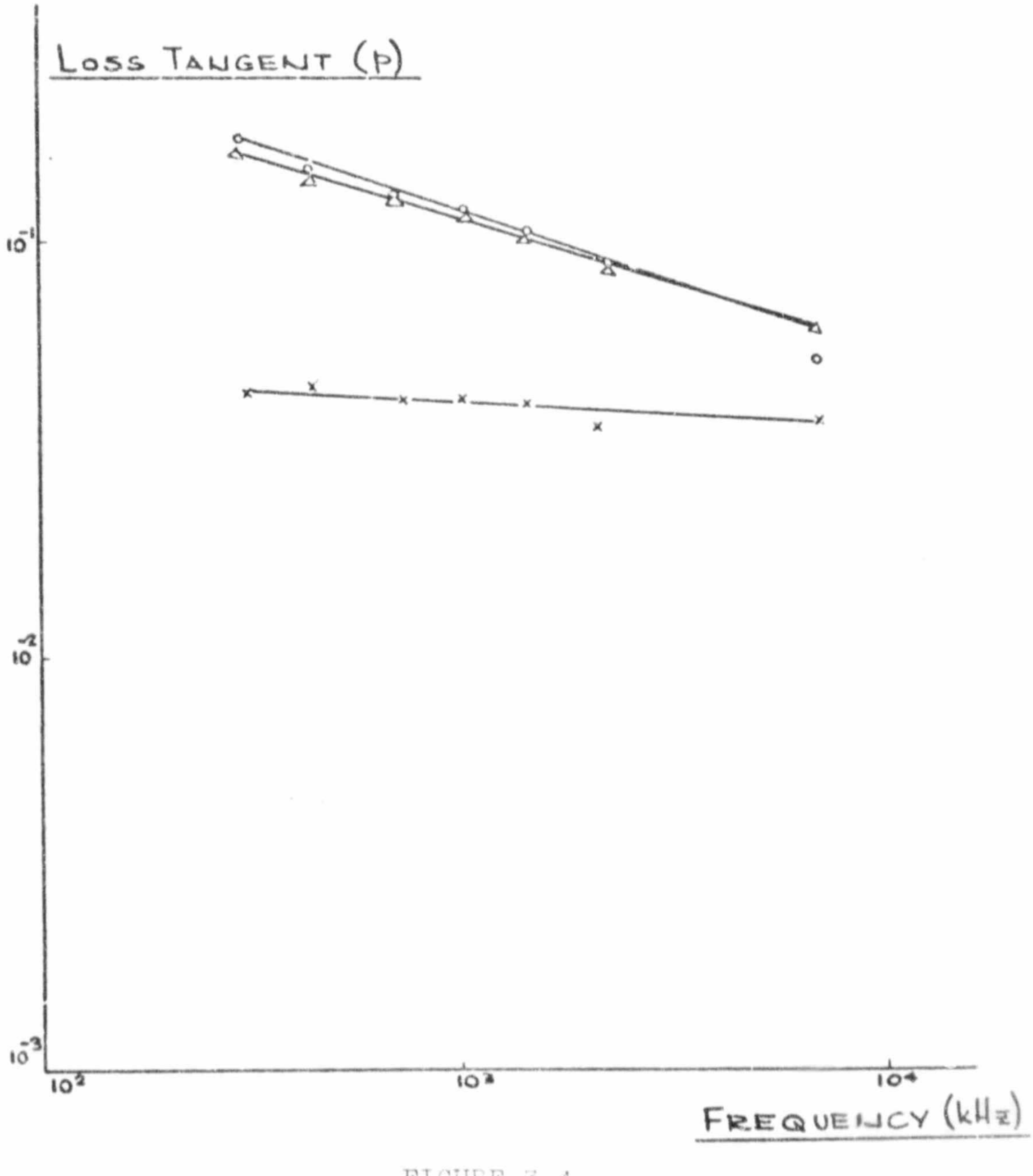

FIGURE 3.4

Variation of loss tangent with frequency for the same samples as in Figure 3.2.

3.1.5 Conclusion

It is thus concluded that values for the electrical properties of rocks, determined from dry laboratory samples, need to be treated with a fair degree of caution since the findings here agree with those quoted in the literature - hence the restrictions on application quoted therein must be heeded.

CHAPTER 4

THE TRANSMISSION SYSTEM

4.1 The transmission equation

The transmission of information through this lossy medium has been shown to involve a large number of factors to which consideration must be given. These factors, namely the propagation medium, the antennas, the separation distance between transmitter and receiver and the power output of the transmitter and receiver and the sensitivity of the receiver, may be conveniently expressed in an equation (Ames et al, 1963):

$$\frac{P_r}{P_t} = \left(\frac{\lambda}{4\pi R}\right)^2 e^{-2\alpha R} G_t G_r \qquad 4 - 1$$

where P_r = Power available at receiver antenna terminals

P_t = Power into transmitter antenna terminals

λ = Wavelength in the medium

R = Separation distance between transmitter and receiver

α = Attenuation constant of the medium

G_t, G_r = Gain of transmitter and receiver antenna respectively

In a practical situation it might well be reasonable to expect the gain (and in fact, all characteristics) of the transmitter and receiver antennas to be similar, therefore, we may write $G_t = G_r = G$.

The term $\left(\frac{4\pi R}{\lambda}\right)^2$ is known as the spreading loss, and accounts for the "spreading" of the electromagnetic energy, as if it were radiated from an isotropic source.

This equation, relating P_r and P_t is valid, subject to the following c[illegible]ions, namely, that far field conditions apply, given by $|\beta R|^2 >> 1$, that the maximum linear dimension of the antenna is considerably less than the separation distance R,

i.e. $\left(\frac{h}{R}\right)^2 << 1$

where β = complex phase constant,

and h = maximum linear dimension of the antenna and

that the medium is homogeneous and large enough such that no interfaces exist which would scatter electromagnetic waves.

In addition to these restrictions the characteristics of a portable, or at least small compared with a wavelength, antenna system will introduce a modifying term into 4 - 1, due to the inefficiency of the antenna as a radiator (and hence receiver, by reciprocity) of electromagnetic energy. Thus η_t and η_r must be included as multiplying factors, and in the case where the antennas are similar, we may write $\eta_t = \eta_r = \eta$ thus 4 - 1 becomes

$$\frac{P_r}{P_t} = G^2 \eta^2 \left(\frac{\lambda}{4\pi R}\right)^2 e^{-2\alpha R} \qquad 4 - 2$$

From a practical point of view the measurement of lossy media propagation characteristics would most likely involve the measurement of voltages related to some characteristics of transmitter and receiver. These voltages would, in turn, be related to the transmitter output power and receiver input power respecti ely. In the case where the transmitter output impedance and receiver input impedance are equal, equation 4 - 2 may be expressed in the form:

$$\frac{V_r}{V_t} = G_r \frac{\lambda}{4\pi R} e^{-\alpha R} \qquad 4 - 3$$

4.2 The measurement problem

In order to determine the characteristics of the medium in order that valid propagation predictions may be made, it is necessary to know accurately the values of conductivity and permittivity for that medium, and also how these vary with frequency. A discussion of these factors constituted Chapter 2 of this dissertation.

It was also discussed therein that no reliance could be placed on laboratory determined values of σ and ϵ_r. Even though a variety of techniques exist for the in situ determination of σ and ϵ_r, many of these have been developed for geophysical purposes and require the deployment of large arrays of measuring electrodes on surface (IEEE Std. 356-1974). A measurement technique based on the transmission of electromagnetic energy between two suitably sited points would seem best suited to the determination of these characteristics in this application. Any method of measurement, employing this technique, would involve, directly, the application of the transmission equation 4 - 3.

The immediate object would be the determination of the attenuation constant α, from which the medium characteristics could be obtained once it were known what value of loss tangent existed. This latter condition is vital to the whole operation since it is the loss tangent p, of the medium, which basically determines its attenuation characteristics with frequency.

However, loss tangent is dependent on both conductivity and permittivity, hence a knowledge of these is required in order that p may be used effectively. In other words the determination of the medium characteristics is dependent on quantities which are themselves inter-related.

4.3 Experimental techniques

Two basic approaches for the determination of the propagation characteristics of the medium exist, which rely on the transmission of electromagnetic energy via that medium. These are described below:

From equation 4 - 3, we may write

$$V_r = V_t \, G \, \eta \, \frac{\lambda}{4\pi R} \, e^{-\alpha R} \qquad 4 - 4$$

Now for a particular measurement system at a fixed frequency in a given area, V_t, G, λ and η are all constants, therefore

$$\frac{V_t \, G \, \eta \, \lambda}{4\pi} = K \text{ (a constant)} \qquad 4 - 5$$

Thus $$V_r = \frac{Ke^{-\alpha R}}{R} \qquad 4 - 6$$

and $$V_r \, R = Ke^{-\alpha R} \qquad 4 - 7$$

therefore $$\ln V_r \, R = \ln K - \alpha R \qquad 4 - 8$$

Equation 4 - 8 may be expressed graphically by plotting $\ln V_r R$ against R, where the slope of the curve yields the attenuation constant α, of the medium and the intersection on the vertical, $\ln V_r R$, axis is

$$\ln K = \ln \frac{V_t\ G\ \eta\ \lambda}{4\pi}$$

Now V_t would be known, but G, η and λ are not quite so readily determined. It will be appreciated that λ is the wavelength in the medium and is given by equations 2 - 14, 2 - 15 and 2 - 17 for various values of loss tangent. This value is thus not simply determined.

Considering now G and η. G is the antenna gain and is related to the antenna efficiency η by the directivity D,

where $\quad G = \eta D$ (Kraus, 1953) $\quad$ 4 - 9

Now $\quad \eta < 1$, therefore $G \neq D$ $\quad$ 4 - 10

D is independent of η and is based entirely on the shape of the far field radiation pattern.

Since it has been repeatedly assumed that the antenna used in this application would be small, its treatment as an elemental device was justified and hence its directivity D would be that of such an antenna where D = 1,5 (Kraus, 1953).

Now from Moore (1963) we find that the maximum gain of a small loop immersed in a lossy medium is given by

$$G = \frac{3,7d}{\lambda} \qquad 4 - 11$$

where $\quad$ d = diameter of the loop,

and $\quad \lambda$ = wavelength in the medium.

Of interest here are the values of gain obtaining for the electric dipole antenna. An insulated wire, with its ends short circuited to the conducting medium (in order to establish more uniform current flow) has a value of gain given by

$$G = \frac{4l}{\lambda} \qquad 4 - 12$$

where l = length of the wire, whereas a completely insulated wire has a gain of

$$G = \frac{3l}{\lambda} \qquad 4 - 13$$

Thus small antennas in a conducting medium have almost equal gains if their dimensions are similar.

Considering again the graphical expression 4 - 8,

$$\ln V_r R = \ln K - \alpha R$$

the intersection on the vertical axis is ln K,

where $$K = \frac{V_t \; G \; \eta \; \lambda}{4\pi} \qquad 4 - 14$$

from which one would imagine that since G can now be approximated,

$$K = \frac{V_t \; 3,7 d \eta}{4\pi} \qquad 4 - 15$$

In a practical system all terms, except η, would be known, therefore, it should be a simple matter to calculate η, and having done so, to substitute it in 4 - 14, from which λ, the elusive wavelength in the medium, could be obtained. What appears to be a very elegant solution unfortunately cannot be used, because it is based on an expression, 4 - 1, which considers the far field zone only.

In other words, since the field pattern in the immediate vicinity of the antenna is considerably more complex than this being considered here, the analysis breaks down.

A further major assumption here is that the characteristics of the medium and the characteristics of the antenna remain constant. This assumes a homogeneous medium, which condition is unlikely to be satisfied in practice. However, the alternative poses many problems, not least of which is that of deciding of what a homogeneous medium consists. On a more macroscopic scale it is possible to evaluate the propagation characteristics of the medium if it is considered to be stratified. This problem has been analysed theoretically by Wait (1962), but will not be considered here.

4.4 Antenna fields

As stated above, the method of measurement proposed is dependent on far field radiation conditions. To investigate these, and in particular, their relationship to any other fields produced by an elemental antenna we proceed as follows:

Consider an infinitesimal loop of wire of area dA carrying a current I. (This restriction on loop area ensures that this current will be of constant amplitude around the loop, and also satisfies the portable aspect in this application where physically small antennas will be used).

The loop is situated at the centre of a spherical cavity of radius "a" which has the following electrical properties:

ϵ_1, μ_1 and σ_1, which can later be allowed to vanish. The spherical cavity is immersed in a dissipative medium characterised by ϵ_2, μ_2 and σ_2.

The time factor $\epsilon^{j\omega t}$ is assumed.

The dipole produces field components given by:

$$H_{r_1} = \frac{IdA}{2\pi R^3} (1 + \gamma_1 R)\, e^{-\gamma_1 R} \cos\theta \qquad 4 - 16$$

$$H_{\theta_1} = \frac{IdA}{4\pi R^3} (1 + \gamma_1 R + \gamma_1^2 R^2)\, e^{-\gamma_1 R} \sin\theta \qquad 4 - 17$$

and

$$E_{\phi_1} = \frac{-j\mu_1 \omega IdA}{4\pi R^2} (1 + \gamma_1 R)\, e^{-\gamma_1 R} \sin\theta \qquad 4 - 18$$

(APPENDIX A)

where the subscript 1 refers to the primary field of the dipole, within the cavity.

A similar set of expressions may be obtained for the field existing within the exterior region, and differ from those above, only in so far as the subscript 2, instead of 1 now applies, and that IdA, known as the magnetic moment of the antenna is replaced by (IdA)e, being the moment of the equivalent magnetic dipole within the spherical, insulating cavity.

Now if $\sigma_1 = o$, $|\alpha| = (\epsilon_1 \mu_1 \omega^2)^{\frac{1}{2}}\, a \ll 1$ and $\mu_1 = \mu_2 = \mu_o$, 4 - 19

then

$$\frac{(IdA)e}{IdA} = \frac{3e^{\beta}}{\beta^2 + 3\beta + 3} \qquad 4 - 20$$

where

$$\beta = \gamma_2\, a \qquad 4 - 21$$

and

$$\alpha = \gamma_1\, a \qquad 4 - 22$$

γ_1 and γ_2 being the propagation constants of the media within and surrounding the spherical cavity respectively, where, in general

$$\gamma^2 = j\sigma\mu\omega - \epsilon\mu\omega^2$$

It is clear that 4 - 20 approaches unity as β approaches zero. In other words, at either sufficiently low frequencies or for sufficiently small cavity radii, the presence of the cavity may be ignored.

On examination of equations 4 - 16, 4 - 17 and 4 - 18 it is evident that three fields with differing rates of decay with distance R from the source exist. The radial magnetic component H_r is made up of two fields, varying as $\frac{1}{R^3}$ and $\frac{1}{R^2}$. The former is the electrostatic field, the latter the induction field, and both are collectively referred to as constituting the "near field".

The angular magnetic components, H_θ, contains fields which vary as $\frac{1}{R^3}$, $\frac{1}{R^2}$ and $\frac{1}{R}$. The $\frac{1}{R}$ component is the "radiation" field, and the only one of the three to involve the flow of real power, hence its relevance in the communication context. The $\frac{1}{R^2}$ and $\frac{1}{R^3}$ fields are related to stored energy considerations (in an antenna in free space) and do not contribute to power flow away from the antenna. For an antenna surrounded by a lossy medium all three components also suffer attenuation due to the dissipative characteristics of the medium.

The above expressions isolate the various field components of the antenna, and also express very succinctly the effect of the medium characteristics and those of the insulating sphere on the radiation performance of the antenna. Of particular importance here is the effect of this insulating cavity, since in any practical mining situation the antenna will always be electrically (and physically) separated from the surrounding rock medium. Even though a spherical insulating cavity was considered above, this was purely with ease of analysis in mind. It is surely reasonable to expect that an irregularly shaped mine tunnel would modify the antenna's performance in a similar manner. The requirement that both α and β in expressions 4 - 21 and 4 - 22 should be very much less than unity involves either the characteristics of the relevant medium ensuring this (through γ), or that the cavity radius "a" should tend to zero. This latter requirement obviously necessitates minimum insulation (air) thickness between antenna and rock, and would seem to agree with that obtained by Kraichman (1962), where the insulating cavity within the antenna has the lesser effect on its performance capabilities.

4.5 <u>Antenna orientation</u>

Equations 4 - 16, 4 - 17 and 4 - 18 provide an insight into the expected orientation of the transmitting and receiving antennas when operating under near field and far field conditions. Figure 4.1 illustrates the orientation of the loop, with the respective field components.

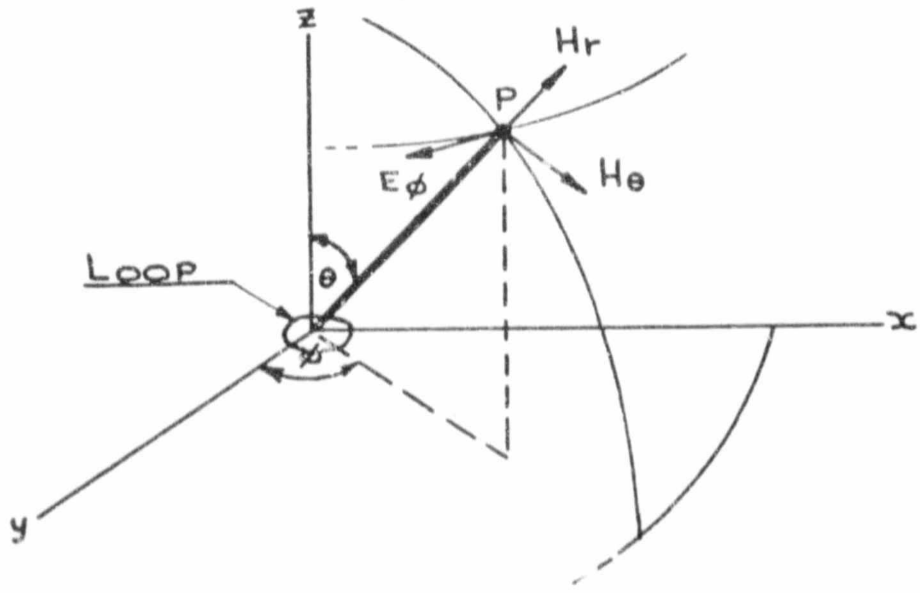

FIGURE 4.1

CO-ORDINATE SYSTEM FOR A CIRCULAR LOOP ANTENNA

The H_r components have their maxima occurring when $\theta = 0°$, or in other words for radiation in the radial direction along the axis of the antenna. This orientation of two antennas to effect maximum signal transfer is referred to later in this dissertation as the coaxial condition. Since no $\frac{1}{R}$ term is present when this orientation is optimum, it is conclusive proof that the two antennas are within the "near field" range.

Both the H_θ and E_ϕ components contain "near" and "far" field terms making their isolation impossible. However, it is interesting to note that only the H_θ component includes the $\frac{1}{R^3}$ term, and this may be significant if propagation differences arise when using either of these coplanar orientations (as they have been referred to in a later section of this work).

4.6 Predominance of the field terms

Since the three field components contain more than just a single term which influence their amplitude with distance, and since the coefficients of these are all different, their relative amplitudes must vary with distance.

To examine this effect consider initially the fields due to an elemental dipole in free space (Jordan,1964). The expression for the E_θ field contains all three terms, as follows:

$$E_\theta = \frac{I\,dl\,\sin\theta}{4\pi\epsilon}\left(\frac{-\omega\sin\omega t'}{Rv^2} + \frac{\cos\omega t'}{R^2 v} + \frac{\sin\omega t'}{\omega R^3}\right) \qquad 4 - 23$$

where dl is the elemental length of an electric dipole, and

$t' = t - \frac{R}{v}$ relates the time of an event to the distance R and velocity of propagation v.

Obviously the three terms will be of equal amplitude when

$$\frac{\omega}{Rv^2} = \frac{1}{R^2 v} = \frac{1}{\omega R^3} \qquad 4 - 24$$

solving 4 - 24 for R we find that

$$R = \frac{\lambda}{2\pi} \qquad 4 - 25$$

This result is particularly significant since for

$R < \frac{\lambda}{2\pi}$ the near field components $^1/_{R^2}$ and $^1/_{R^3}$ predominate, whereas for

$R > \frac{\lambda}{2\pi}$ it is the far field component $^1/_R$ which predominates.

If we consider now the case of fields propagating through a lossy medium we need only examine 4 - 19.

$$H_0 = \frac{IdA}{4\pi R^3}\left(1 + \gamma R + \gamma^2 R^2\right) e^{-\gamma R} \sin\theta$$

The field terms are equal in amplitude when

$$\left|{}^{1}/_{R^3}\right| = \left|{}^{\gamma}/_{R^2}\right| = \left|{}^{\gamma^2}/_{R}\right| \qquad 4 - 26$$

Thus $R = \frac{1}{|\gamma|}$ 4 - 27

Now for the case where the loss tangent of the medium

$p \leq 0{,}6$ we have

$$\lambda = \frac{\lambda_o}{\sqrt{\epsilon_r}} \qquad \text{(see equation 2 - 15)}$$

where $\lambda_o = \frac{1}{\sqrt{\mu_o \epsilon_o}\, f}$

therefore

$$\lambda = \frac{1}{\sqrt{\mu_o \epsilon}\, f} \qquad 4 - 28$$

since $\epsilon = \epsilon_o \epsilon_r$

and since we are considering the low loss medium condition we may write the propagation constant as

$$|\gamma| = \omega\sqrt{\mu_o \epsilon} \qquad 4 - 29$$

and since

$$R = \frac{1}{|\gamma|}$$

we have

$$R = \frac{1}{2\pi\, f\sqrt{\mu_o \epsilon}} = \frac{\lambda}{2\pi} \qquad 4 - 30$$

For the condition of high loss tangent, $p \geq 10$, we have

$$|\gamma| \doteq \sqrt{\mu_o \omega \sigma} \qquad 4 - 31$$

and from equation 2 - 16

$\lambda = 2\pi\ \tau$, where τ is the skin depth in the medium

$$\lambda = \frac{2\pi}{\sqrt{f\ \pi\ \mu_o\ \sigma}} = \frac{2\pi\ \sqrt{2}}{|\gamma|}$$

therefore

$$R = \frac{\lambda}{2\sqrt{2}\ \pi} \qquad 4 - 32$$

This result is very significant, illustrating, as it does, the difference between low and high loss media. If the wavelength in the medium is known it presents a ready method of determining the conductivity of that medium since

$$\sigma = \frac{4\ \pi}{\lambda^2\ f\ \mu_o} \qquad 4 - 33$$

In theory then, it would appear that the above approach is a straight forward method of obtaining the characteristics of the medium. This is unfortunately not the case, however, since the point at which all three field terms are equal in amplitude, for the field component in question, is not apparent. This is due to the fact that from a measurement point of view these terms are not separable.

Examination of 4 - 16, 4 - 17 and 4 - 18, however, yields an interesting result when the H_r and H_θ components are compared. H_r contains only near field components and is a maximum when measurement is made along the axis of the antenna. In other words maximum signal transfer between a transmitting and a receiving antenna occurs when they are aligned coaxially. This characteristic of a loop antenna was measured in an underground test and the result is presented in Figure 4.2. As predicted by 4 - 17, maximum signal occurs at $\theta = 0°$ and $180°$.

The H_θ component, on the other hand, produces maximum field strength along the $\theta = 90°$ and $270°$ directions, due to the sinusoidal angular dependence. Again this characteristic was confirmed by underground measurement with results presented in Figure 4.3. It should be pointed out at this stage that in fact the E_ϕ component is also present in this measurement since the sensing antenna used, did not discriminate between E and H fields. The orientation of the two antennas favouring maximum signal transfer here is now obviously coplanar. If it were poss eliminate the E_ϕ component from this latter measurem then an interesting comparison can be made between the remaining H_θ and the H_r fields. This is examined below.

4.7 <u>Comparison between H_θ and H_r fields</u>

From 4 - 27 the near field and far field terms of all components are equal when

$$R = \frac{1}{|\gamma|}$$

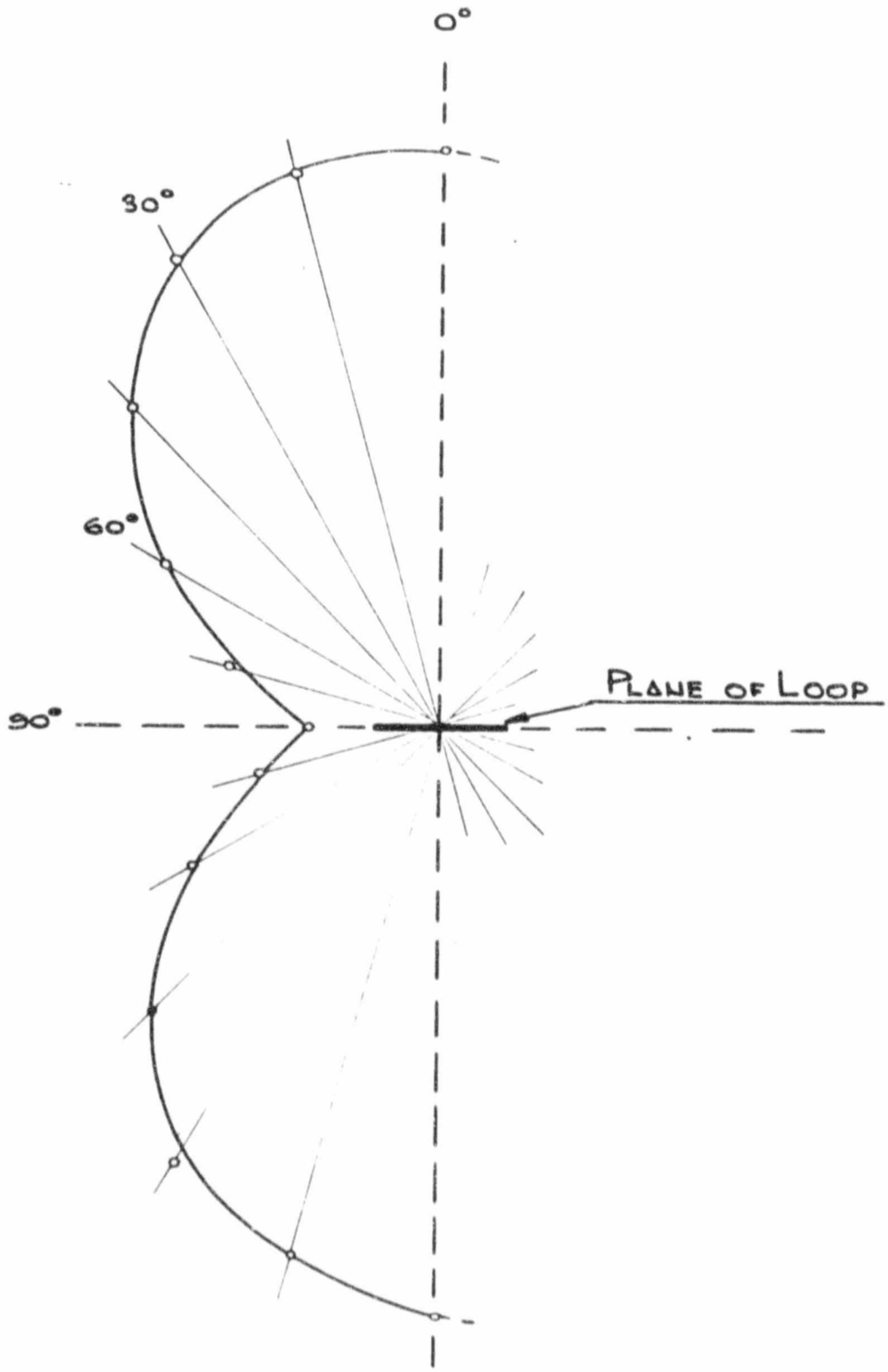

FIGURE 4.2

Measured field strength of the component with angular variation.

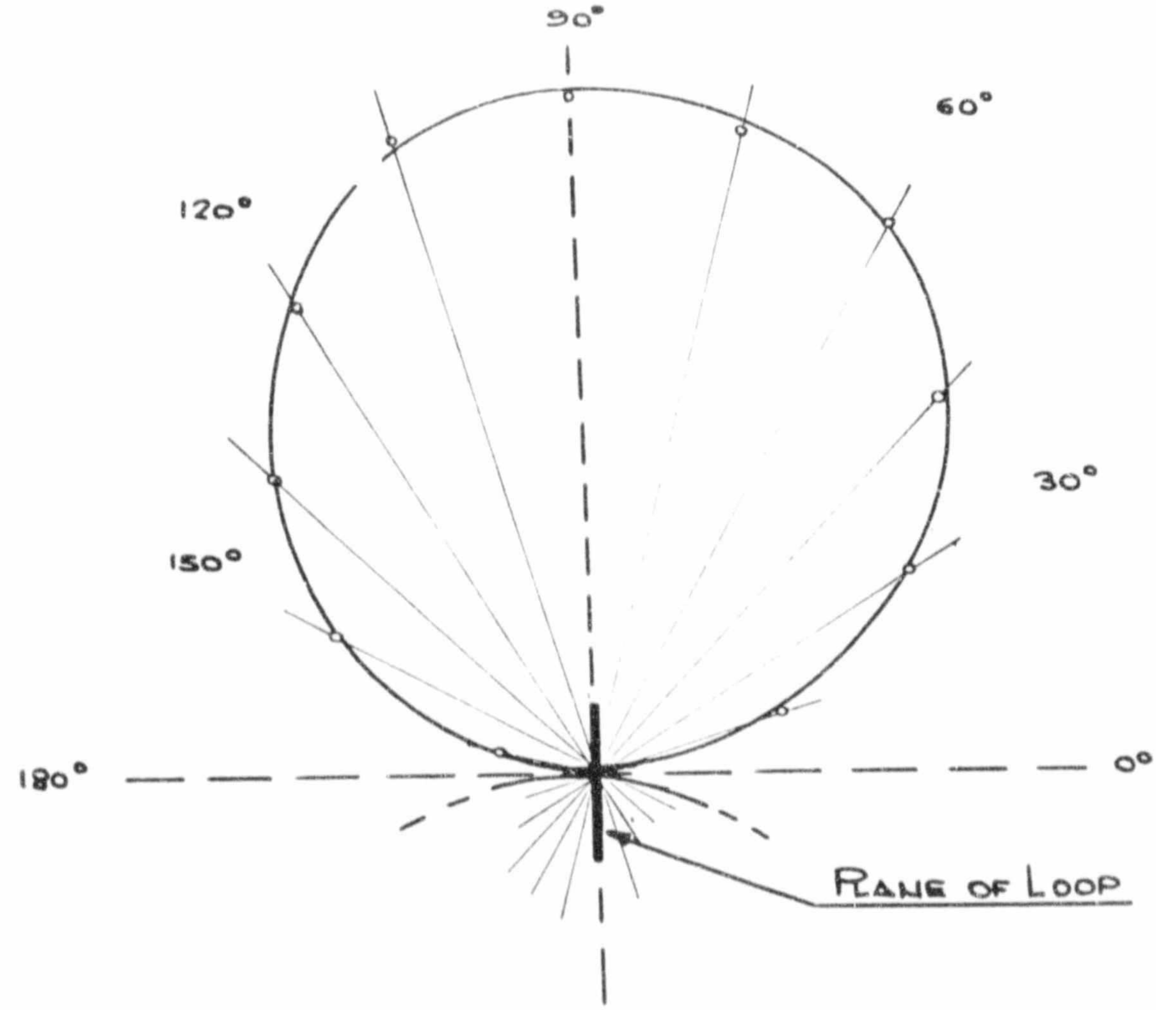

FIGURE 4.3

Measured field strength of H_θ component with angular variation.

For this case, from 4 - 17, when $\theta = 0^\circ$, we have

$$|H_r| = \frac{IdA}{2\pi} \left(\frac{1 + 1}{1/|\gamma|^3}\right) e^{-1} \qquad 4 - 34$$

thus $$|H_r| = \frac{IdA \; |\gamma|^3 \; e^{-1}}{\pi} \qquad 4 - 35$$

from 4 - 18, when $\theta = 90^\circ$, we have

$$|H_\theta| = \frac{IdA}{4\pi} \left(\frac{1 + 1 + 1}{1/|\gamma|^3}\right) e^{-1} \qquad 4 - 36$$

thus $$H_\theta = \frac{IdA}{4\pi} \; 3|\gamma|^3 \; e^{-1} \qquad 4 - 37$$

therefore, the ratio of $|H_\theta|$ to $|H_r|$, yields

$$\frac{|H_\theta|}{|H_r|} = \frac{3}{4} \qquad 4 - 38$$

Again, this is a very interesting result since it makes possible the determination of the distance, R, from the source, at which point the far field and near field components are equal in amplitude. This information is invaluable since it immediately provides a means of determining the electrical characteristics of the medium at the frequency at which the measurement was made. Also it indicates very definitely whether communication would be within the zone of near field or far field predominance.

H_r is readily determined on its own since the antenna orientation which maximises this component is 90° from that which is optimum for the far field components. However, in the far field, both H_θ and E_ϕ must be present