BuMines OFR 134-78

PB289742

ELECTROMAGNETIC GUIDED WAVES IN MINE ENVIRONMENTS

Proceedings of a Workshop

Prepared for

United States Department of the Interior
Bureau of Mines

by

James R. Wait

Institute for Telecommunication Sciences
National Telecommunications and Information Administration
United States Department of Commerce
Boulder, Colorado 80303

Iterim Report for Period
2 February 1978 to 30 May 1978

Contract No. HO155008

31 May 1978

Bureau of Mines Open File Report 134-78

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50272-101

REPORT DOCUMENTATION PAGE	1. REPORT NO. BuMines OFR 134-78	2.	3. Recipient's Accession No. PB289742

4. Title and Subtitle ELECTROMAGNETIC GUIDED WAVES IN MINE ENVIRONMENTS Proceedings of a Workshop	5. Report Date May 31, 1978
	6.
7. Author(s) James R. Wait (Editor) and others	8. Performing Organization Rept. No.
9. Performing Organization Name and Address Institute for Telecommunication Sciences National Telecommunications and Information Administration U.S. Department of Commerce Boulder, CO 80303	10. Project/Task/Work Unit No.
	11. Contract(C) or Grant(G) No. (C) H0155008 (G)
12. Sponsoring Organization Name and Address Office of Assistant Director--Mining Bureau of Mines U.S. Department of the Interior Washington, D.C. 20241	13. Type of Report & Period Covered Contract research 2/2/78-5/30/78
	14.

15. Supplementary Notes

16. Abstract (Limit: 200 words)

This report is a compilation of papers presented and discussed at a workshop held on March 28-30, 1978, at the Institute of Telecommunication Sciences on the general subject of mine communications. Active workers from the United States, Canada, England, France, Belgium, South Africa, and Japan were invited to present their latest research results and exchange views. An edited record of the discussions that followed the presentations is included.

17. Document Analysis a. Descriptors

Mine communications
Tunnel waveguides
Radiowave propagation
Carrier trolley wire transmission
Geophysical probing
Mine rescue

b. Identifiers/Open-Ended Terms

c. COSATI Field/Group 08I

18. Availability Statement	19. Security Class (This Report)	21. No. of Pages
Unlimited release by NTIS.		333
	20. Security Class (This Page)	22. Price PC A15 MF A01

(See ANSI-Z39.18)

See Instructions on Reverse

OPTIONAL FORM 272 (4-77)
(Formerly NTIS-35)
Department of Commerce

FOREWORD

This report was prepared by the Institute for Telecommunication Sciences, National Telecommunications and Information Administration, U.S. Department of Commerce, Boulder, Colorado under USBM Contract number H0155008. It was administered under the technical direction of the Pittsburgh Mining and Safety Research Center with H.K. Sacks acting as Technical Project Officer. This report is a summary of the work recently completed as a part of this contract during the period 2 February 1978 to 31 May 1978. This report was submitted by the authors on 31 May 1978.

The document is a compilation of the papers presented and discussed at a three day informal meeting on the general subject of mine communications. Active workers from the U.S., Canada, England, France, Belgium, South Africa, and Japan were invited to present their latest research results and exchange views. These papers are included here along with an edited record of the discussions that followed the presentations. Any views expressed or alluded to in the individual papers are not necessarily endorsed by the U.S. Bureau of Mines and, in no way, do they represent the official position of the U.S. Government on such matters.

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In organizing the workshop, the undersigned received a great deal of assistance from his colleague, David A. Hill, who was responsible for most of the physical arrangements. Secretarial support was provided by Lana Hope who also assisted in the preparation of these Proceedings. Our technical Project Officer, H.K. Sacks from the U.S. Bureau of Mines, was also a continuous source of encouragement. Also, much credit must go to my colleagues at Arthur D. Little, Inc.: R.L. Lagace, R.H. Spencer, and A.G. Emslie for their vital contributions and many useful suggestions. Robert Lagace, in particular, played a major role in making this workshop so successful. Their summary and suggestions for further research are included at the end of this document.

Speaking on behalf of my North American colleagues, I would like to warmly thank the following people who travelled a great distance to attend this workshop: Brian Austin from South Africa, Jiro Chiba from Japan, David Martin from England, Paul Delogne from Belgium, and Robert Gabillard from France. Their active participation was a key ingredient. Our only regret was that Quin Davis, the organizer of the 1974 Leaky Feeder Conference in Guildford, was not able to attend but his paper was ably presented by David Martin.

Finally, I would like to thank John N. Murphy of the U.S. Bureau of Mines, not only for his general support of the workshop, but for taking the time from his busy schedule to give an excellent and vigorously presented overview lecture on mine communi-ation research. This particular event was sponsored jointly by the local chapters of the IEEE Communications Society and the IEEE Antennas and Propagation Society.

31 May 1978 James R. Wait

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EXPERIMENTAL RESULTS OBTAINED IN A MINE ON A NETWORK OF LEAKY CABLES

P. DEGAUQUE, B. DEMOULIN, M. CAUTERMAN and R. GABILLARD, Electronics dept., Lille University, 59650 VILLENEUVE D'ASCQ - FRANCE

1. INTRODUCTION

A few years ago, the "Centre d'Etudes et de Recherches des Charbonnages de France" (CERCHAR) wished to provide a radiowave communication with a leaky cable which can be easily installed in the galleries. Indeed, since this cable can be put near the coal face, its path and thus its environment, is often changed. Then, to realize such a mobile communication at a given frequency (about 7 MHz), we have proposed to use a leaky braided coaxial cable. At first, we recall the characteristics of the braid and we indicate the changes since the original design in order to obtain a tensing and crushing strength appropriate to the working conditions. With this type of leaky coaxial cable, two kinds of transmission may be used. The first one is a communication between two mobile transceivers, called MOTRANS in the following, and the second one corresponds to a communication between a MOTRANS and a control place in which an other transceiver, called COATRANS is directly connected to the coaxial cable. After a brief description of these transceivers, we give the results in situ obtained in two mines.

2. DESCRIPTION OF THE TRANSMISSION SYSTEM

2.1. Coaxial cable characteristics

From an electromagnetic point of view, one of the most important characteristic is the transfer inductance of the braid. We have shown in previous papers (1), (2) that the optimum value is about L_t =40 n H/m for a transmitting frequency of 7 MHz. The geometrical characteristics of the braid such as the number of carriers and the number of braid wires per carrier have also been determined and two kinds of cables have been built successively. They only differ from one another by the diameters d_1 of the inner conductor, d_2 of the braid and by the thickness h of the outer jacket. They are given by :

Cable 1	d_1 = 1.5 mm	d_2 = 7 mm	h = 1 mm
Cable 2	d_1 = 2.3 mm	d_2 = 6.5 mm	h = 2.8 mm

Table 1

The first communication was obtained 4 years ago with cable 1, but from this time it is appeared that the cable was often either short circuited or in open circuit due to mechanical working conditions.

Mechanical tests of tensing and crushing strength have given the following average results on cable 1 (Table 2)

. Relative elongation just before the breaking load : 35 %

. Maximum crushing strength before a short circuit inner conductor-braid :

340 bars (3400 N/cm^2)

. Maximum tensing strength (breaking load) : 720 Newton

Table 2

To improve the reliability of the communication, it appeared necessary to change these mechanical characteristics. For that, the diameter of the inner conductor and the thickness of the outer jacket have been increased.

In order to obtain a cable with a not too important diameter, the thickness of the filling dielectric has been reduced. The new characteristics have been given in Table 1 and the mechanical results are

. Maximum tensing strength (breaking load) : 1350 Newton

. Relative elongation just before the breaking load : 31 %

. Maximum crushing strength before a short circuit inner conductor- braid :

240 bars (2400 N/cm^2)

Table 3

From these values, we see that the tensing strength could be doubled while the maximum crushing strength is smaller than the one for cable 1. Indeed, in that case, it is the mechanical resistance of the filling dielectric material which plays a leading part and, for cable 2, its thickness has been reduced. Furthermore, we can remark that the presence of a thick outer jacket is useful to protect the cable against frictions. The main difference on the electrical characteristics between the two cables is their impedance : 80 Ω for cable 1, and 55 Ω for cable 2, but this parameter has only a small influence on the performances of the cable.

We have also measured the apparent conductivity σ and the relative permittivity ε_r of the outer jacket as a function of the frequency and the temperature. The results are described in Table 4.

F kHz	σ mho/m	ε_r	t° °C	σ mho/m	ε_r
1	1,9 10^{-8}	4,6	70	1,7 10^{-5}	4,6
10	1,9 10^{-7}	4,1	50	1,5 10^{-5}	4,2
100	1,8 10^{-6}	3,6	30	1,2 10^{-5}	3,8
1000	10^{-5}	3,2	10	6,6 10^{-6}	3,2
t° = 15° C			f = 500 kHz		

Table 4

Wait and Hill (3), Mahmoud and Wait (4) have extensively study the propagation characteristics along a leaky cable by taking into account the presence of such a dielectric jacket and a possible outer lossy film. At 7 MHz, this jacket has not a great influence on the propagation constants provided that the cable is not situated just near the tunnel wall.

2.2. Influence of the position of the cable

The position of the cable inside the gallery is often a difficult problem for miners and many configurations may occur. For example, in the tunnels where experiments have been made, there are two extreme cases. In the first case, the cable is well positioned, at about 20 cm from the tunnel wall and it is parallel, at 30 cm, to a rail used for the hanging trains. In the second case, the leaky cable creeps among many other cables and its relative position is not at all a constant. Furthermore, in some cases, there is a branching of two galleries, the cable coming from the main tunnel to go in a very small one with a diameter of about 2 m, which is a crossing between two galleries at various levels. The experimental results of propagation show that such sections do not create an important disturbance on the total transmission. Last, since the cable could be moved, they are no special connectors and often, when the cable is broken, miners makes only a splice.

2.3. Transceivers

As we have already indicated, there are two kinds of transmitter. The coaxial transceiver COATRANS, which is directly connected to the cable and is often put near the tunnel wall and the mobile transmitter (MOTRANS) with an inductive coupling to the cable, the antennas consisting of a coil wound around a small bar of ferrite 10 cm long. The MOTRANS can be carried by miners or put on a train. In this last case, the antennas are positioned above the vehicule to obtain a small coupling loss with the cable.

The COATRANS is characterized by an output power of 1 watt (50 Ω) and a sensitivity of 1 μV. The transmitting power of the MOTRANS could be characterized by a voltage of 1 volt which is induced on a one turn coil with a diameter of 1 m, the MOTRANS being put at the center of the coil. Thus, it gives a magnetic field of $3\ 10^{-4}$ A/m at a distance of 2 m, while, as a receiver, it can detect a magnetic field of 0,1 μA/m (x).

3. *EXPERIMENTAL RESULTS OBTAINED IN A MINE*

We describe the results obtained in two mines corresponding to various configurations. In the first case, the network has no branch but the propagation constant of the monofilar mode varies, for a good deal, along the cable while, in the second case, the communication path is established on two levels and each one with 4 or 5 branches.

3.1. *Network without branch*

Measurements have been made on a cable section 1500 m long. For the coaxial transmitter, the curve in Figure 1 represents the variation of the current amplitude along the surface of the braid as a function of the distance x transmitter-receiver. The average attenuation is 2.3 db/100 m, corresponding to the coaxial eigen mode. Then, we have used a MOTRANS as transmitter, the coaxial receiver being connected to the cable. Since the received signal depends on the relative position of the MOTRANS, this one is always moved just against the leaky cable. The curve in Figure 2 shows the variation of the voltage at the end of the cable as a function of the distance x. We see that the maximum voltage is 100 mV when x is small, i.e. when the transmitter is near the coaxial receiver, while for the 1.5 km path, this value becomes equal to 1,7 mV. Of course, we find the same attenuation per unit length as in the previous case.

Radio communication experiments between two mobile transceivers have also been tested. In that case, a MOTRANS used as a transmitter is placed at the end of the cable and the current variation along the coaxial cable is given in Figure 3. Curves 1 and 2 correspond to two positions of the transmitter, at each endof the cable as indicated in the figure. The experimental results show the two eigen modes, coaxial and monofilar. Indeed, in the communication from point A to point F the current attenuation along the section AB is 8.6 db/100 m (monofilar mode), while from F to A the attenuation along this same section AB is only 2.5 db/100m, the coaxial mode becoming dominant.

(x) These transceivers are sold in France under the trade mark of X phone for the MOTRANS and Y phone for the COATRANS. These apparatus are approved for use in fire damp environment (European specifications). Furthermore this type of leaky braided coaxial cable is called CERLIL (Trade mark).

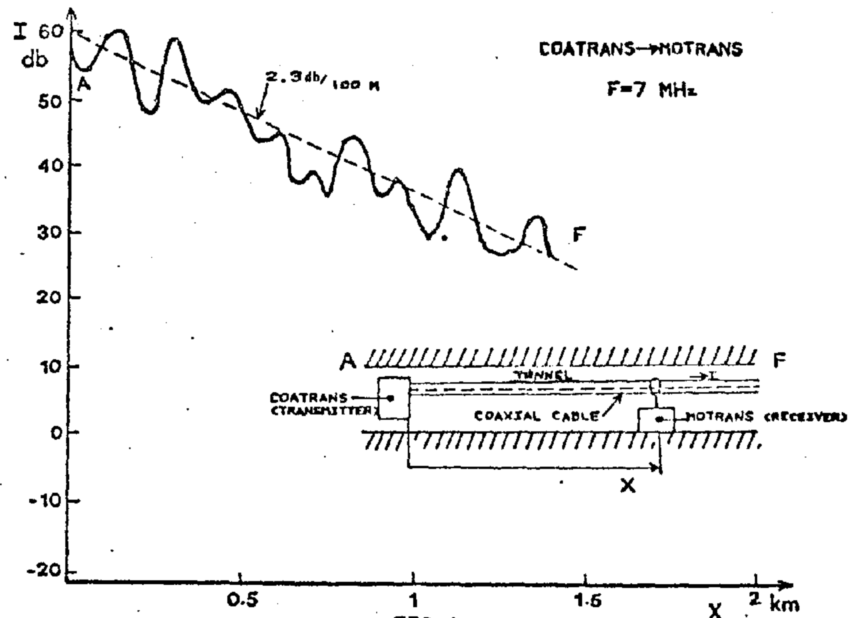


FIG.1

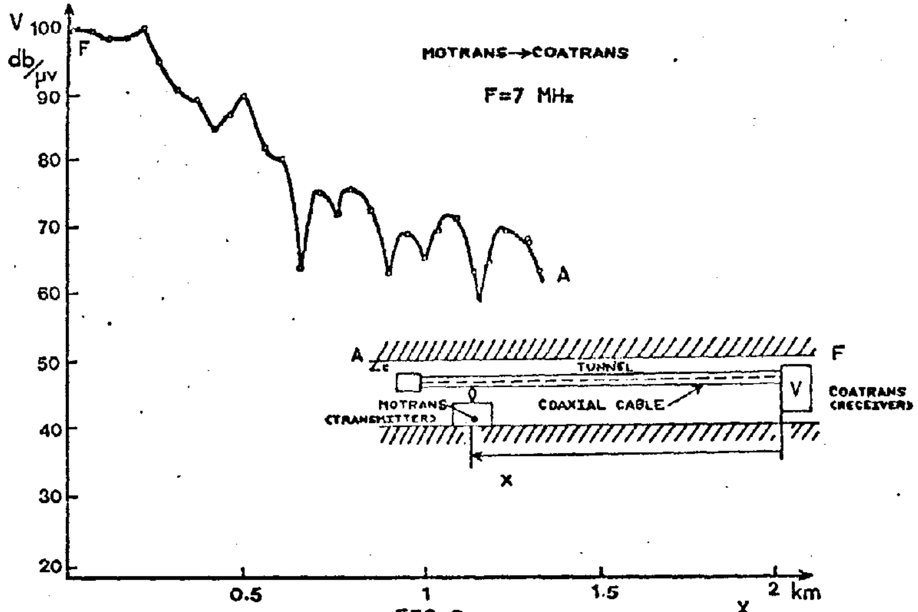


FIG.2

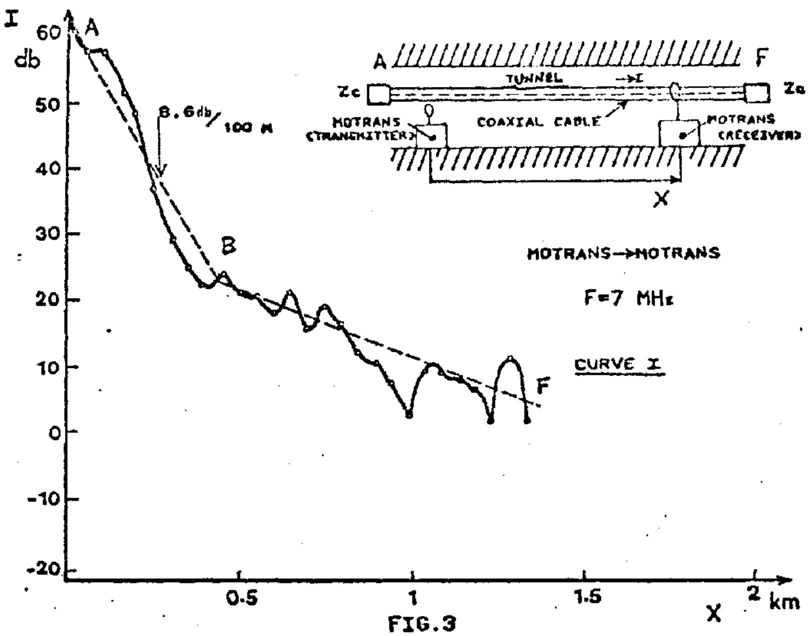


FIG.3

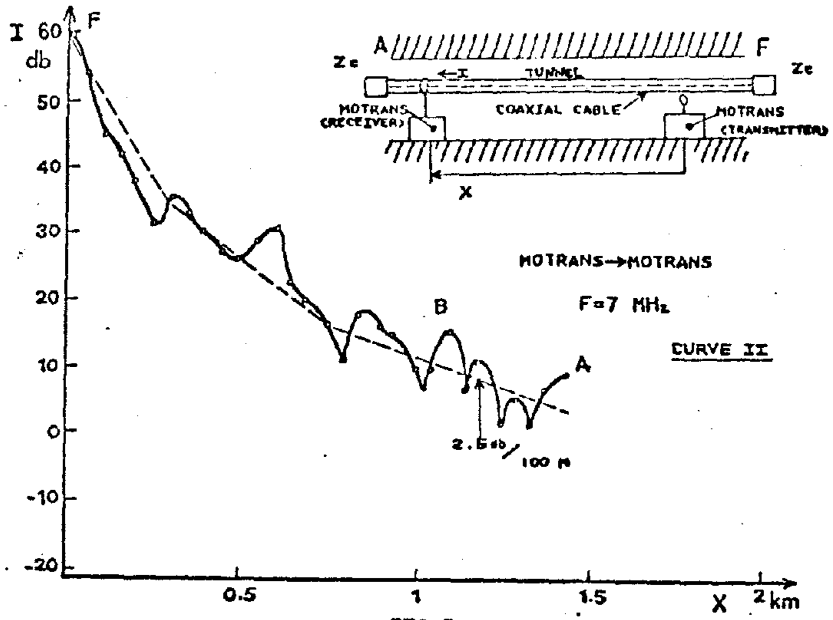


FIG.3

3.2. *Network with branches*

The network diagram and the galleries where the measurements have been made are indicated in Figure 4. The COATRANS is situated at the center of an Y connection, the total cable length is 12 km but each branch, from the COATRANS, has a maximum length of 2 km.

The curve in Figure 5 represents the variation of the current amplitude along the leaky cable when the coaxial transmitter is connected at point 0. Along this path, there is no transformer at the connection and the matching is only made by inserting resistances between the inner conductors. The various letters A, B ... F correspond to points of measurement indicated on the network diagram. On 1700 m, we obtain an attenuation of about 60 db. Then, if we suppose that the attenuation due to each branch is 6 db, the attenuation per unit length is 2.5 db/100 m, value which is nearly equal to the one obtained with a cable without branch. Then, the transmission has been made with a MOTRANS and we have measured the voltage on the coaxial receiver, connected at point 0 as a function of the distance transmitter-receiver.

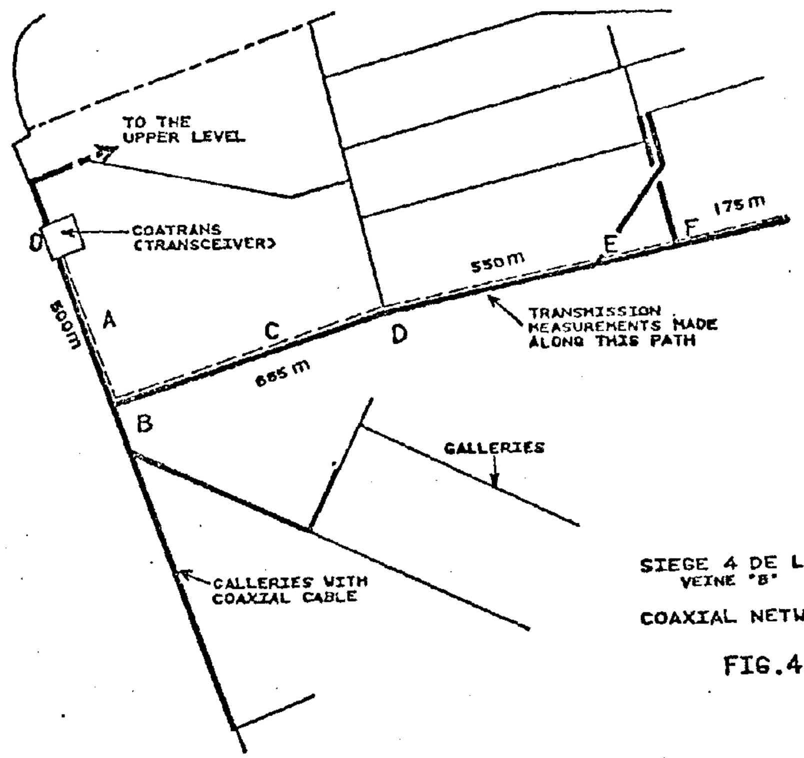
The results are shown on the curve in Figure 6. We remark that the presence of branches does not give important variations of the received signal when the transmitter is situated near a connection.

Other experiments have been made to know the coupling loss between the mobile transmitter and the cable as a function of their distance. The results are shown on the curves in Figure 7 which represent the additional loss measured on the coaxial receiver, when the MOTRANS moves from the cable to the bottom of the gallery. The reference (0 db) is the level obtained when the MOTRANS is put near the cable. At a given distance D between the MOTRANS and the cable, the MOTRANS is always oriented such that the received signal is maximum. From these two curves, we see that the coupling loss depends on the configuration of the gallery (presence of cables, rails...) but the value of 40 db can be considered as a maximum. In an operational transmission, it is of course necessary to add the loss due to an arbitrary orientation.

Then, we have represented in Figure 8 the equivalent of a "Radiation pattern". It gives the amplitude of the induced current on the braid as a function of the angle θ made by the MOTRANS, $\theta = 0$ corresponding to the optimum coupling. From these curves, we see that the orientation is not a too critical parameter.

3.3. *Noise measurements*

During operation, the machinery used in a mine creates a wide range of many types of electromagnetic interference. Two measurements techniques have been used. The first one is the direct measurement of the electric field in galleries where the leaky cable is installed. Of course, there are many sources which generate noise and the electric field noise is a function of many parameters as : orientation and location of the receiving antenna, time... Then, we have made this experiment near a power unit driving a belt conveyor in order to obtain only an idea of the frequency range in which noise is important. The receiver is a field intensity analyzer and we measure the average field intensity within a bandwidth of 9 kHz. The spectral components amplitudes, greater than 1 μV/m are shown in Figure 9. We have also indicated the noise due to asynchronous motors. This noise is easily recognizable since the amplitude of its components is modulated by a low frequency signal due to the rotor current.



SIEGE 4 DE LENS
VEINE "B"

COAXIAL NETWORK

FIG.4

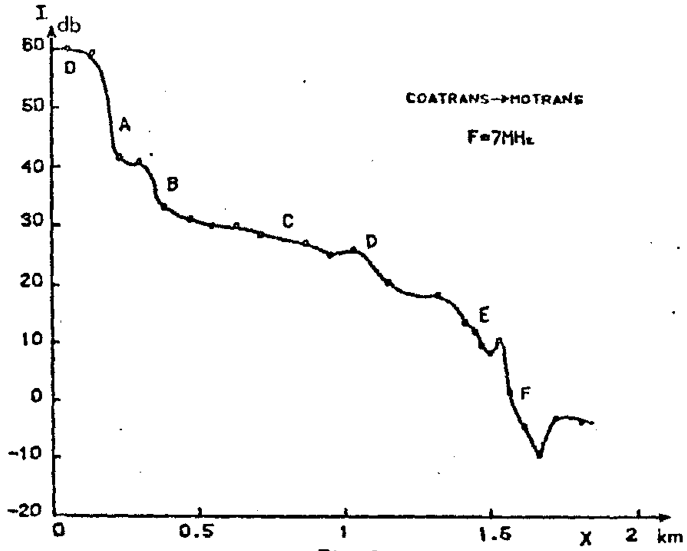


FIG.5

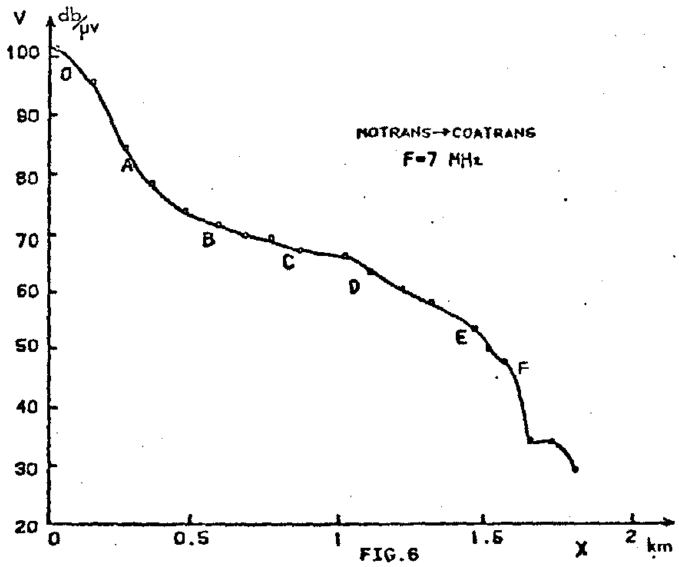


FIG.6

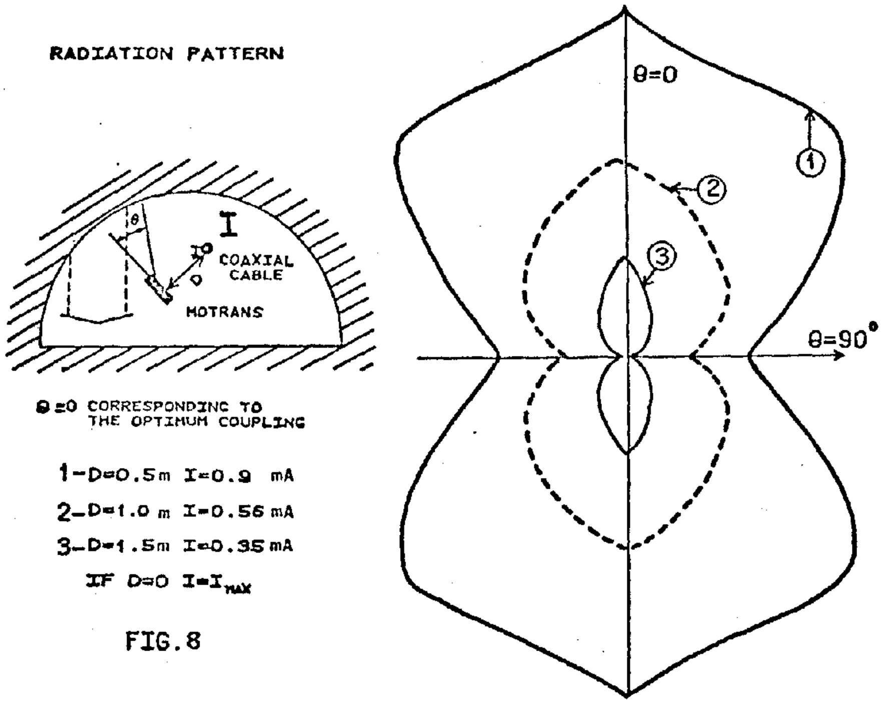


FIG.8

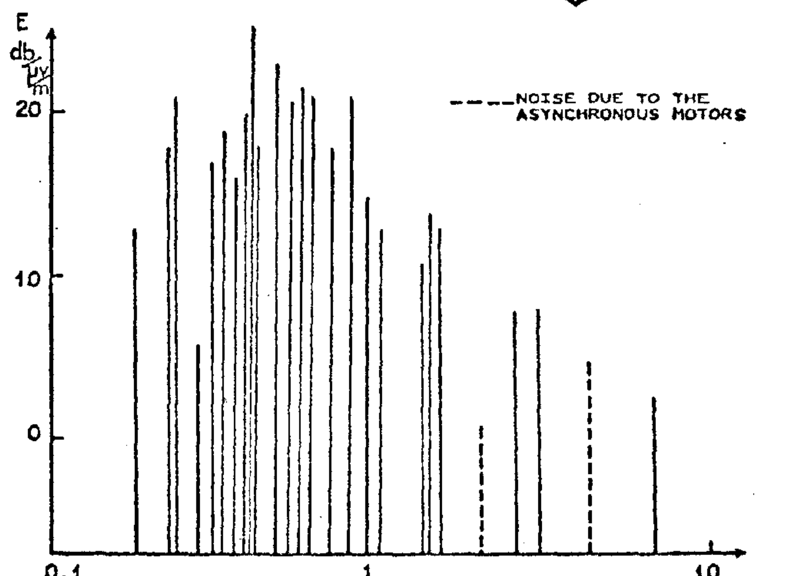


FIG.9

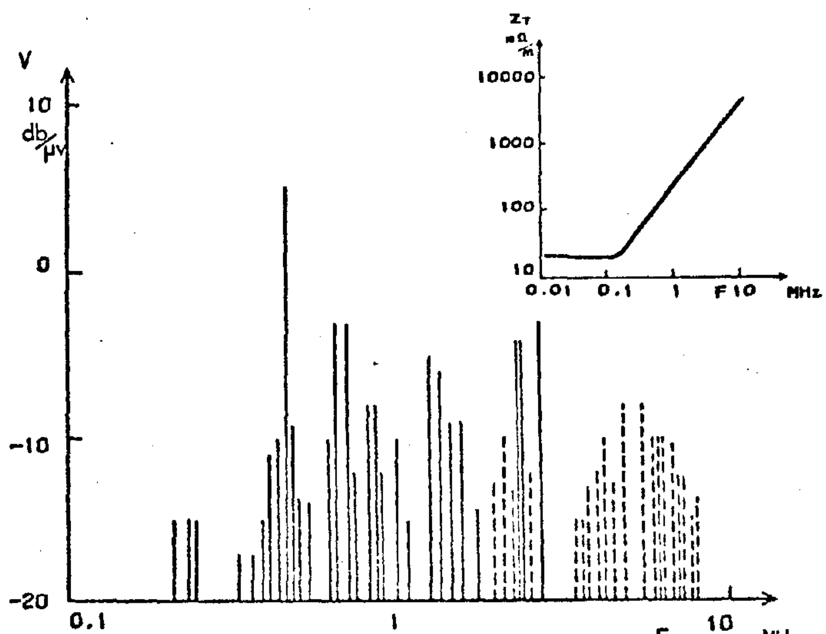


FIG.10

The second set of measurement wasmade with the same field intensity analyzer but directly connected to the cable, the total cable length for this configuration is 12 km. The results concerning the average voltage amplitude are given in Figure 10. If we compare Figures 9 and 10, we can remark that the variation of the transfer impedance Z_t as a function of the frequency (Figure 10) gives a relative attenuation of the noise in the low frequency range (100 kHz-1MHz). Last, the numerous asynchronous motors induce more spectral components but we see that, in our frequency range, the average field due to noise is smaller than 0.5 μV in the 9 kHz band width.

4. *CONCLUSION*

The experimental results show that the performances of the cable does not depend appreciably on its environment. Furthermore, branches can be added to an actual network as they are wanted. Of course, the efficiency of the radio communication can be improved by inserting various types of cables (low loss coaxial cables, leaky cables...) in order to optimize the energy radiated by the cable along its path. But such a solution was not possible in the mines where the experiments have been made. Indeed, the network of leaky cables is set near workings and its path is often subject to modifications which can be made by unskilled workers.

ACKNOWLEDGEMENTS : The work described in this paper was supported by Centre d'Etudes et de Recherches des Charbonnages de France (CERCHAR). We wish to thank the personnel of the "Bassin des Houillères de Provence, Nord- Pas de Calais et Lorraine" for their cooperation during the experiments.

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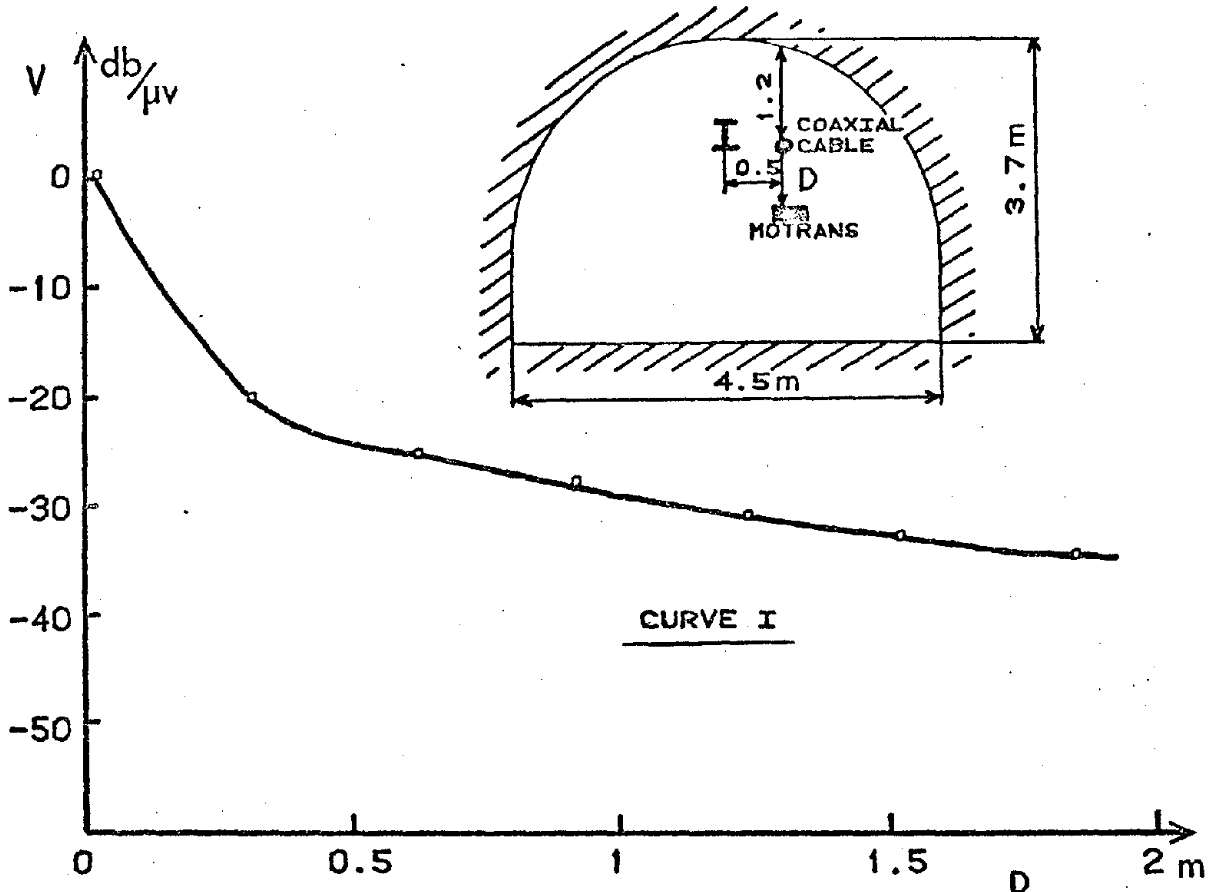
V db/μv
0
-10
-20
-30
-40
-50
0
0.5
1
1.5
2 m
D
1.2
COAXIAL
CABLE
0.5
D
MOTRANS
3.7 m
4.5m
CURVE I

FIG.7

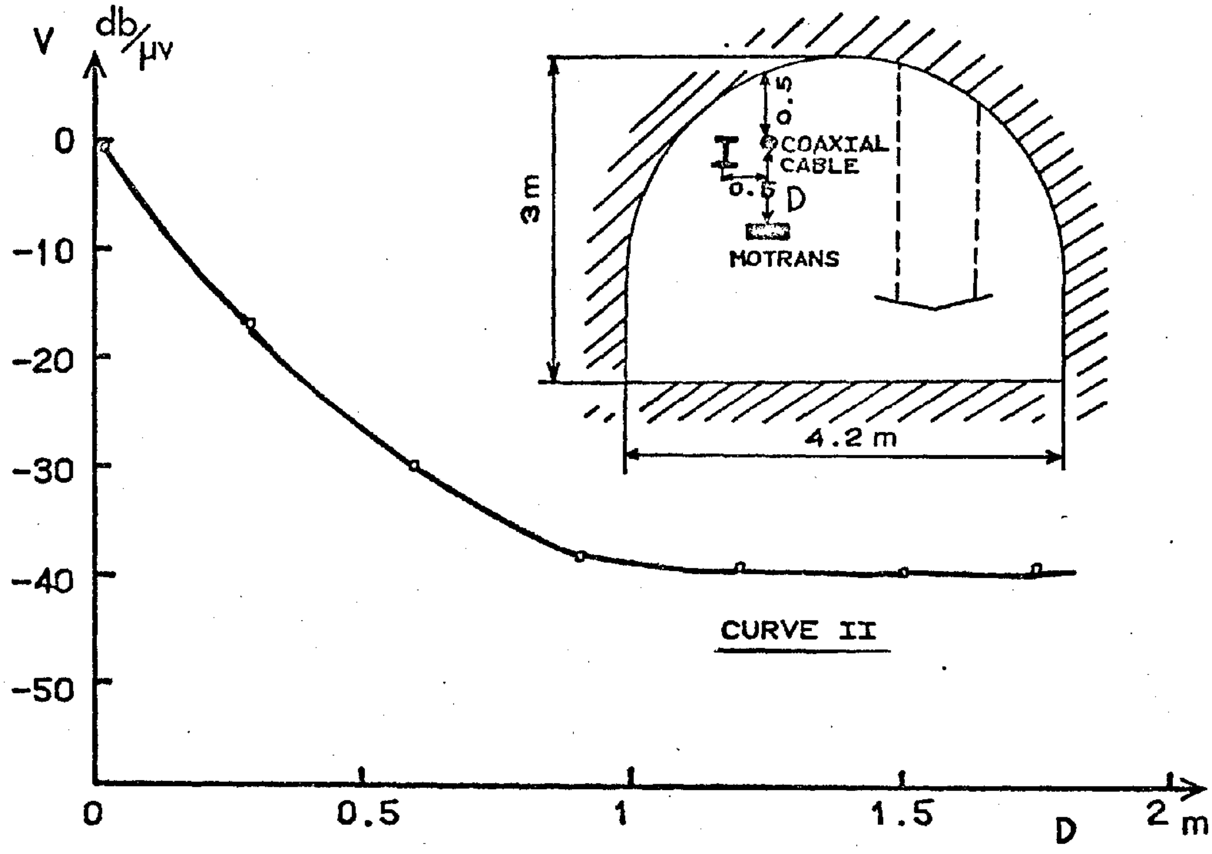
V db/μv
0
-10
-20
-30
-40
-50
0
0.5
1
1.5
2 m
D
0.5
COAXIAL
CABLE
0.5
D
MOTRANS
3 m
4.2 m
CURVE II

FIG.7

COMPARATIVE ANALYSIS OF LEAKY CABLE TECHNIQUES FOR MINE COMMUNICATIONS

R. De Keyser(1), P. Delogne(2), L. Derijck(3) and R. Liégeois(1)

1. INTRODUCTION

The propagation of e.m. waves in tunnels has been extensively studied in the past decade and it is now possible to compare different techniques which can be used for radiocommunications in the mine and to draw general conclusions. It first appears that one has to consider independently the low and the high frequency cases. For the purpose of this analysis we call low frequencies those lying under the tunnel cut-off. It is clear that one must, in this frequency range, resort to the monofilar mode and to leaky feeder techniques which excite this mode, but the analysis must also take into account that the antennas of the mobile transceivers are necessarily very short compared with the wavelength and may thus be rather inefficient. At high frequencies natural or waveguide mode propagation is possible but it is severely affected by obstacles and bends of the tunnel and leaky feeder or antenna techniques are frequently required. It is the purpose of this paper to compare the various methods and to predict obtainable ranges.

2. THE LOW FREQUENCY CASE

2.1. The monofilar mode

When a conductor is strung parallel to the tunnel axis a TEM-like mode can propagate : this is the monofilar mode. The properties of this mode depend to a large extend on the position of the wire with respect to the wall and, but less severely, on the electrical characteristics of the ground.

Experimental [1] and theoretical [2] studies show that the specific attenuation increases with the frequency and when the distance of the wire to the wall decreases. Physically this effect can easily be understood by the fact that the return current uses only a small fraction of the tunnel perimeter as shown qualitatively on Fig. 1. The attenuation curves given there can be approximated by the equation

$$\log \alpha_m = .43 - .29\, d + (.96 - .49\, d) \log f \qquad (1)$$

where the attenuation is in dB/100 m, the distance in meter and the frequency in MHz. In the absence of more accurate values this equation may be used for coal mines. It is surely not valid for extreme cases of ground conductivities like in salt mines or in some metallic mines [3].

A second very adverse effect of the cable proximity w.r.t. the wall is the eccentricity effect consisting in a distortion of the field lines as suggested by Fig. 1. This may yield a very bad coupling of the antennas of mobile transceivers with the monofilar mode. A comparison with exact calculations [4,5] shows that, provided the skin depth of the surrounding medium is not much larger that the cross-section of the tunnel, one may approximate the field distribution by that of the TEM case with a perfectly conducting wall.

(1) INIEX, rue du Chéra 200, 4000 Liége, Belgium
(2) Université Catholique de Louvain and INIEX
(3) Université de Liége and INIEX.

This approximation yields a pessimistic estimation of the eccentricity effect but it is believed that the error does not exceed a few dB in many cases. The error may however be important for very low frequencies and/or very low conductivities, like for instance in salt mines. This assumption will be used in this paper.

In order to carry out calculations it may be useful to consider a reference tunnel and apply a correction for the eccentricity effect. We use the reference of a circular tunnel with a radius of 3 meter in which the wire is located along the tunnel axis and the field quantities are considered against the wall. In order to simplify all calculations we also assume that the characteristic impedance of the monofilar mode is equal to the intrinsic impedance of free space (377 Ω), irrespective to the wire radius and position; the resulting error does not exceed 2 or 3 dB. The fields of the monofilar mode in the reference tunnel are then related to the power of this mode by

$$E_{ref} \simeq \sqrt{P_m} \tag{2}$$

$$H_{ref} \simeq 2.73 \,.\, 10^{-3} \sqrt{P_m} \tag{3}$$

The eccentricity penalty that must be applied to take account of the actual tunnel cross-section and wire position can be evaluated using electrostatic theory. For instance, in a circular tunnel with a radius R and for an observation point located against the wall opposite to the wire, it is given by

$$C_{ecc} = 20 \log \frac{R(2R - d)}{3d} \tag{4}$$

and is shown on Fig. 2. For a rectangular tunnel the penalty can be calculated using the quadruple infinity of images shown on fig. 3; it may be written

$$C_{ecc} = 10 \log(ab/9) + f(\alpha/a, \beta/b) \tag{5}$$

where the function f depends on the ratio b/a and on the position of the observation point. As an example Fig. 4 gives this functions for b/a = 1/2 and for an observation point located at (0.25a, 0.25b). Figures 1, 2 and 4 show that the position of the cable in the tunnel cross-section is an important factor in the design of an installation; a proper choice can save many dB's. This effect is particularly critical for mobile-to-mobile communications where the eccentricity penalty has to be counted twice.

Finally and in order to complete the model it is also necessary to express the transceiver characteristics in terms of the monofilar mode in the reference tunnel. One may in general consider that the effective height of the antenna is the same in the tunnel and in free space and this allows to convert the receiver sensivity into a minimum required electric or magnetic field, according to the type of antenna which is used and finally, using (2) and (3) , into a minimum required monofilar mode power. The power excited by a mobile transmitter in the reference tunnel can be obtained by a special application of the reciprocity theorem; for instance, a magnetic Hertz dipole with a magnetic moment M (ampere.m^2) excites a power given by

$$P_{m,ref} = 1.16 \quad 10^{-4}\ M^2\ f^2_{MHz} \tag{6}$$

This power is in most cases of the order of microwatts because M does not exceed a few mA.m^2. It should be stressed that manufacture specifications announcing transmitter powers of the order of 1 watt are meaningless; it is believed that they refer to the complex power supplied by the transmitter to the antenna but do not take into account that this power is nearly reactive for electrically small antennas.

As a numerical example we consider the communication from a mobile to a receiving base station connected to the cable in a tunnel with R = 2 m and d = 0.2 m. The frequency is 5 MHz. The mobile transmitter has a magnetic moment 10 mA.m^2 and the base station sensitivity is 10 μV/377 Ω, i.e. - 125.7 dBW. The calculations are as follows

- monofilar mode spec. attenuation, eq.(1) 9.4 dB/100 m
- eccentricity loss, Fig. 2 22 dB
- $P_{m,ref}$, eq.(6) -65.4 dBW
- P_m = -65.4 - 22 -87.4 dBW
- allowable monofilar attenuation : 125.7-87.4 = 38.3 dB
- maximum range : 38.3/9.4 x 100 = 407 m

2.2. Continuous leaky cables

Leaky coaxial cables or bifilar lines are used to obviate the high attenuation suffered by the monofilar mode. They support two eigenmodes which are frequently called the coaxial and the monofilar eigenmodes |4,6|. The monofilar eigenmode is not directly useful, excepted in the vicinity of a transmitter, because it suffers a high specific attenuation. For practical applications one may consider that the mobile transceivers are connected to the leakage field of the coaxial eigenmode.

Excepted in the immediate vicinity of the cable, the leakage field has the same spatial distribution as that of the monofilar mode analyzed in the previous paragraph [4] and the eccentricity loss can be calculated in the same way. The only difference is that a higher power is required from the coaxial mode to produce a given field. For a leaky coaxial cable this ratio can be expressed by a coupling coefficient given by

$$C_1 = \left| m/(\ell_m \ell_c)^{1/2} \right| / [2(\varepsilon^{1/4} - \varepsilon^{-1/4})] \qquad (7)$$

where m is the transfer inductance, ℓ_m and ℓ_c are the specific inductance of the coaxial and monofilar coupled modes resp., and ε is the dielectric constant of the cable insulation. Assuming characteristic impedances of 50 and 377 Ω resp. for theses modes, the coupling coefficient can be put in the practical form

$$C_{dB} = -20 \log C_1 = 59{,}2 - 20 \log \frac{m_{nH/m}}{\varepsilon^{1/2} - 1} \qquad (8)$$

This coupling loss can be reduced by increasing the transfer inductance and reducing the dielectric constant. The basic idea in the use of leaky feeders compared to a simple monofilar wire is that the lower specific attenuation of the coaxial mode will allow to compensate the coupling loss and still gain a certain number of decibels. The range improvement is given by the equation

$$\alpha_c R_c = \alpha_m R_m - C \qquad (9)$$

which obviously recommends a low specific attenuation of the coaxial mode and a low coupling loss.

However, if the coupling is increased beyond a certain limit, this will increase the specific attenuation of the coaxial eigenmode because of the current induced in the tunnel wall by the leakage current. This effect can be calculated in a very simple way. Consider an elementary section dx of a leaky cable carrying a power P_0; the power loss in this section is partly due to the cable itself and partly to the wall; remembering that the leakage field has the same value as the field of a monofilar mode with power $C_1^2 P_o$, the power loss will be given by

$$dP = (2\ \alpha_{co} + 2\ \alpha_m\ C_1^2)\ P_o\ dx$$
$$= 2\ \alpha_c\ P_o\ dx \qquad (11)$$

where α_{co} is the specific attenuation of non-leaky coaxial cable. Thus

$$\alpha_c = \alpha_{co} + \alpha_m\ C_1^2 \qquad (12)$$

Equation (9) can now be written

$$R_c = \frac{\alpha_m\ R_m + 20\ \log C_1}{\alpha_{co} + \alpha_m\ C_1^2} \qquad (13)$$

The optimum design of a leaky cable, i.e. the choice of the coupling coefficient C_1, depends strongly on the total allowable propagation attenuation ($\alpha_m R_m$) for the case of a simple monofilar wire and on the monofilar mode specific attenuation α_m as shown on Fig. 5(a). The first of these parameters ($\alpha_m R_m$) depends on the radio equipment performance and both parameters ($\alpha_m R_m$) and α_m vary much with the cable position with respect to the wall. It is thus not sure that a leaky cable can be design for a good performance in all cases. Fig. 5(b) shows in continuous lines the range obtained if the coupling coefficient is optimum and in broken lines that obtained with a coupling coefficient which is optimum for $\alpha_m R_m$ = 40 dB and α_m = 10 dB/100 m. It is seen that a serious range reduction can sometimes result form an inadequate choice of C_1.

The drawback of a continuous leaky cable is its lack of flexibility. In practice, the cable distance to the wall may vary significantly along a path. If the distance can be small on a short length somewhere on the path, one should be tempted to increase the coupling coefficient C_1 in order to improve the signal at that place, but the penalty will be an increase of the coaxial mode attenuation over the whole length as shown by eq.(12). Ideally the coupling C_1 should vary along the path, but this is of course impraticable.

2.3. Mode converters

Mode converters have been designed to obviate the poor flexibility of continuous leaky cables. Various types of mode converters have been developped by INIEX. They convert an arbitrary fraction of the power carried by a non-leaky cable into the monofilar mode supported by the external surface of this cable. Historically the first mode converter (INIEX/Delogne system) was based on the principle of an annular slot of the outer coaxial cable [1,8,9]. A similar INIEX/Derijck system was developped for a bifilar line [10]. More recently a system in which mode converters consist in leaky sections inserted in a non-leaky cable has been designed [7] and a U.S. Patent is pending for.

All these mode converters can be used at any frequency. They have very wide bandwidths and offer a very high flexibility resulting from the fact that the spacing of the converters and the conversion coefficients can be varied in function of the distance to the base station and of the local conditions, namely the cable position. This is particularly the case for the annular slot converter. Fig. 6 shows the most extensive system which has been realized up to now; it is installed in a French potash mine and provides communications at 7 MHz in a starred network starting from a base station located at point A ; the total range covered is about 18 km with 48 mode converters (mean spacing 372 m).

It should be stressed that continuous leaky cables and mode converter systems are strictly equivalent for what regards the eccentricity loss in tunnels like haulageways. However, the mode converters have also a radial radiation which can be useful at low frequencies in some applications wherein the cable has to be laid against the wall. Of course, the converter spacing must then be very small. This property is currently used in long wall faces with converter spacings of about 30 m, the cable being laid along the belt conveyor with the power cables and fluid tubes.

3. THE HIGH FREQUENCY CASE

3.1. Natural propagation

When the frequency increases above the tunnel cut-off, a increasing number of waveguide modes can propagate. In the first octave above the cut-off, they still suffer a very high attenuation. At higher frequencies, the number of propagating modes is so great that a geometrical optical method is a more useful approach than the addition of waveguide modes [11]. Both methods agree to predict important standing waves, which can be considered the result of different waveguide mode velocities or of reflections on the wall.

Field measurements along a path parallel to the tunnel axis show a global exponential decrease superimposed on an irregular standing wave pattern. The attenuation between two dipole antennas separated by a distance d in a straight-line tunnel can be written in the form

$$a = a_o + s + \alpha d \qquad \text{(dB)} \qquad (14)$$

where a_o is the extrapolation to d = o of the standing wave maximums envelope, s is a margin for standing waves with some probability of occurence (i.e. 95 % not exceeded), and α is a specific attenuation. Some loss has to be added for each bend of corner.

Table 1 gives experimental results obtained in an iron mine with 7 x 8 m tunnels. They should not be extrapolated to other cross-sections without much care. The ranges given in this table are for a total loss of 138 dB corresponding to a 1 watt transmitter and a 1 μV receiver. The general evolution of the parameters is as expected. The small value of the bend loss at 68 MHz could be explained by the fact that the tunnel width was close to 1.5 λ. It was also observed that the straight-line crossing of a tunnel intersection had no effect.

f	a_o	s	α	corner	ranges (m)		
MHz	dB	dB	dB/100 m	dB	straight	1 corner	2 corners
36	17	5	60	6	193	183	173
68	20	10	40	2	270	265	260
150	21	15	35	15	291	249	206
450	24	22	15	25	613	447	280

Table 1. Natural propagation in an iron mine

It is seen that the range obtained with natural propagation is very limited excepted for straight-line paths.

3.2. Continuous leaky feeders

The working principle of continuous leaky feeders at high frequencies is now well understood [7]. First it is clear that the monofilar eigen mode does not play any role because its specific attenuation is very high. The radiation is due to the scattering of the leakage fields of the coaxial eigenmode by various discontinuities among which the cable fastening devices and the wall irregularities. The random character of this process has two drawbacks. In the first place, the radiation is strongly dependent on the local conditions, which requires to take a safety margin for the calculations; as a demonstration of this, it has been observed that a leaky coaxial cable suspended by thin nylon strings at 30 cm of the tunnel wall had a negligible radiation at 450 MHz. In the second place, a strong radiation requires an important scattering of the coaxial mode leakage field and this yields an increase of the specific attenuation of this mode; this effect increases with the frequency and it is frequent that the specific attenuation at 450 MHz is doubled when the cable is fixed against or very near the wall.

The coupling loss, which is defined as the ratio of the power carried by the coaxial eigenmode to the power received by a dipole antenna, may be dependent on the local conditions. It does not appear to depend strongly on the position of the receiving antenna in the tunnel cross-section in which the leaky cable is strung. Important standing waves are observed; they are due to the large number of radiation sources with random phases and this yields a Rayleigh distribution of the field intensities. A standing wave margin of 25 dB has to be included to obtain coverage at 99 % of places. With this definition, the coupling loss of leaky coaxial cables is frequently near 100 dB. A 1/2 inch diameter coaxial cable has typical specific attenuations of

2 dB/100 m at 70 MHz
3.2 dB/100 m at 150 MHz
6 dB/100 m at 450 MHz

and if we allow a 30 % increase due to radiation, the obtainable ranges will be 1460, 913 and 587 m respectively, for a 1 W transmitter and 1 μV receiver.

It is seen that the cable attenuation has a dramatic influence on the range. Of course the cable radiation can diffuse in lateral tunnels with a specific attenuation varying as shown in Table 1; this is in favor of the higher frequencies. The lateral range is nevertheless limited by the fact that the initial coupling loss is always high.

3.3. Discrete radiators

Discrete antennas connected to a non-leaky coaxial cable are in principle a good solution for the coverage of an area. Compared to continuous leaky cables they have the advantage to avoid a useless increase of the coaxial mode attenuation by allowing a better control of the radiation process. They also allow to reach a higher radiation level, which allows to cover a larger lateral range. Compared to natural propagation they offer the advantages of the lower specific attenuation of the coaxial cable and in particular to obviate the effects of tunnel bends and corners. In fact, they are potentially able to combine the advantages of both solutions.

For cost and reliability reasons however, it should not be very intersting to connect a large number of discrete antennas to a main coaxial cable by means of couplers and secundary cables. An alternative solution is offered by the mode converters described in paragraph 2.3. At high frequencies these devices act as antennas connected to the cable and they can be designed to radiate up to 20 % of the power carried by the cable. The annular slot radiator has a typical radiation pattern shown by curve 1 of Fig. 7. This is also the radiation pattern of an elementary leaky section. By properly chosing the length of the section this pattern can be modified; it can for instance be multiplied by cardioid 2 to yield curve 3. The length of the leaky sections is of the order of a wavelength while spacing between them is of the order of 100 m. They offer a bandwidth of 1.6 octave; it is thus possible to incorporate the leaky sections in the cable at the manufacturing stage in order to provide a system which will work satisfactorily at very different frequencies (i.e. simultaneously at 150 and 450 MHz). Careful evaluation showed that such a cable, compared to a continuous leaky cable, would increase the linear range by 30 % and the lateral range by 100 % or more.

4. GENERAL CONCLUSIONS

Various techniques are now available for mine communications. They have different properties which are summarized in Table 2. The calculated ranges take into account of electrical noise and antenna performance. It should be stressed that these techniques can frequently be combined and allow the simultaneous use of very different frequency bands : one may insert mode converters for low and high frequencies in the same cable which may be continuously leaky or comprise leaky sections designed for higher frequencies.

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System	Frequency	Linear range (1)	(2)	(3)	Area coverage (1)	(2)	(3)	Availability of equipment
Monofilar	500 kHz 1 MHz 5 MHz 10 MHz	3000 3000 1200 700	1800 1300 800 700	600 700 700 500	No			.1-1 MHz available for mine use Cerchar X-Y phones at 7 MHz See companion paper for 5-10 MHz band
Leaky cable	500 kHz 1 MHz 5 MHz 10 MHz	3600 7600 6500 4400	1800 4000 7000 5200	Not useful 2500 2000				
Mode converters	same as above	increase leaky cable values by 30 %						
Leaky cable	27 MHz 70 MHz	2000 1500	2000 1500	1500 500	No			Surface equipment
Mode converters	27 MHz	3000	3000	2000				
Antennas	150 MHz 450 MHz	200 to 300 300 to 600			200 x 300 600 x 400			
Leaky cable	150 MHz 450 MHz	900 600	900 600	Not useful	900 x 100 600 x 200	Not useful		
Cable and discrete radiators	150 MHz 450 MHz	1500 1000	1500 1000	Not useful	1500 x 200 1000 x 300	Not useful		

Table 2. Available systems and expected range in meter.
(1) base station to mobile (2) mobile to base station
(3) mobile to mobile

[9] U.S. Patent n° 3.829.767, 13th August 1974.

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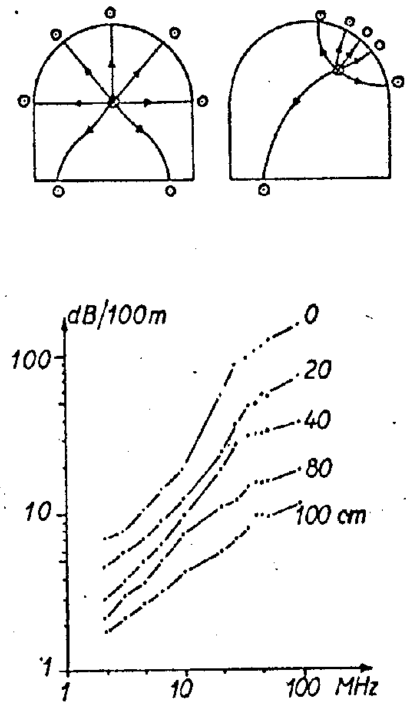


Fig. 1. Electric field lines, current distribution and experimental attenuation of the monofilar mode as a function of the distance between the cable and the wall.

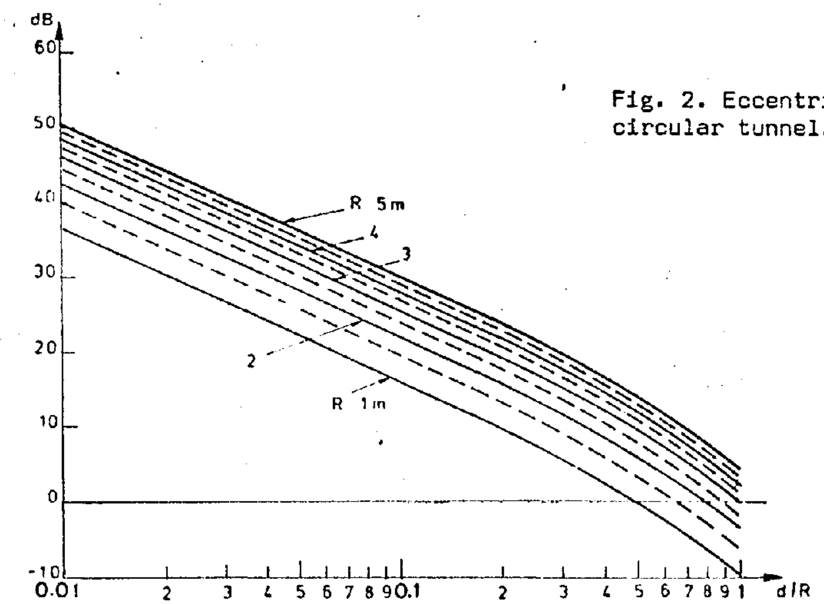


Fig. 2. Eccentricity loss for circular tunnel.

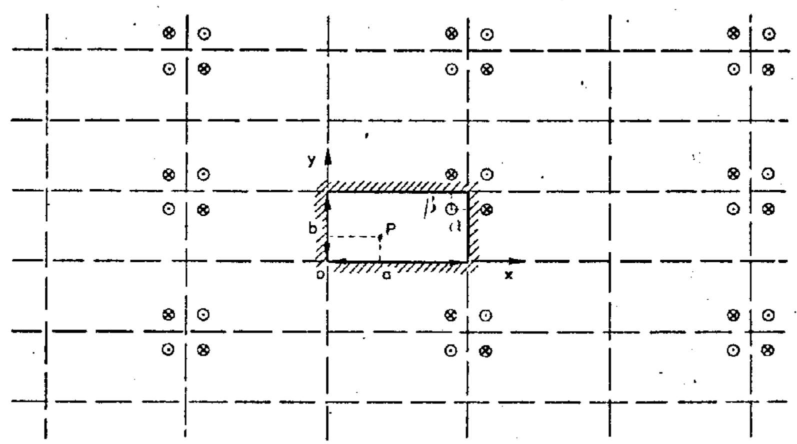


Fig. 3. Double infinity of images for a rectangular tunnel.

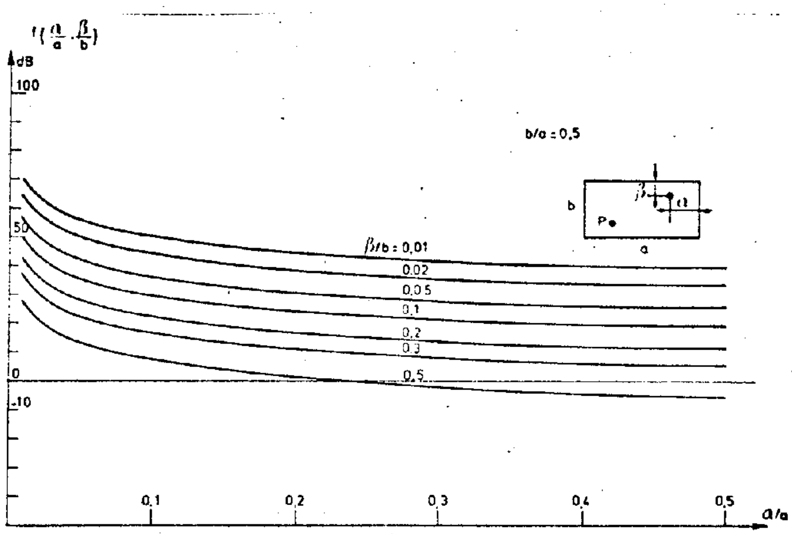


Fig. 4. Eccentricity loss for rectangular tunnel.

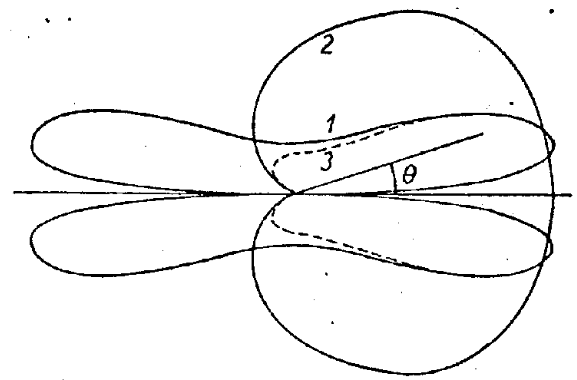


Fig. 7. Radiation pattern of a leaky stub inserted in a nonleaky cable.

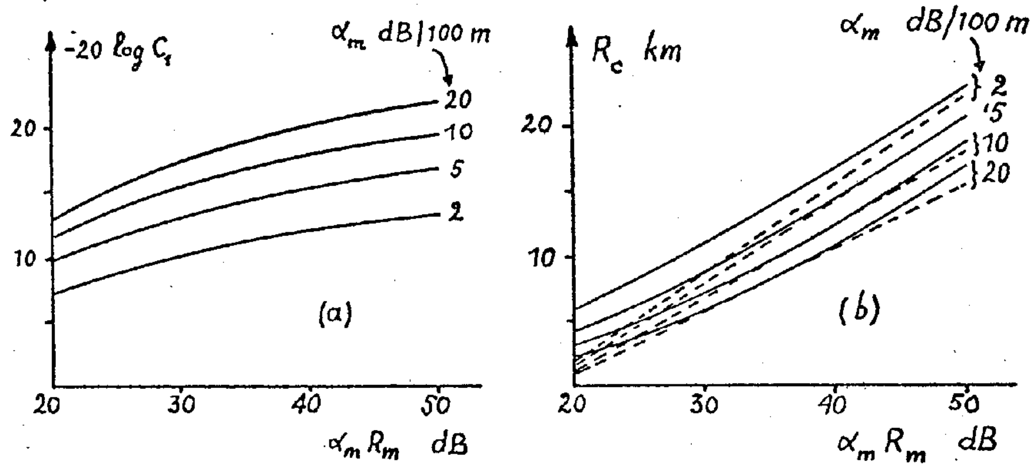


Fig. 5. Optimum coupling loss (a) of a continuous leaky feeder and maximum range (b) for optimum coupling (continuous lines) and 17 dB coupling (dotted lines).

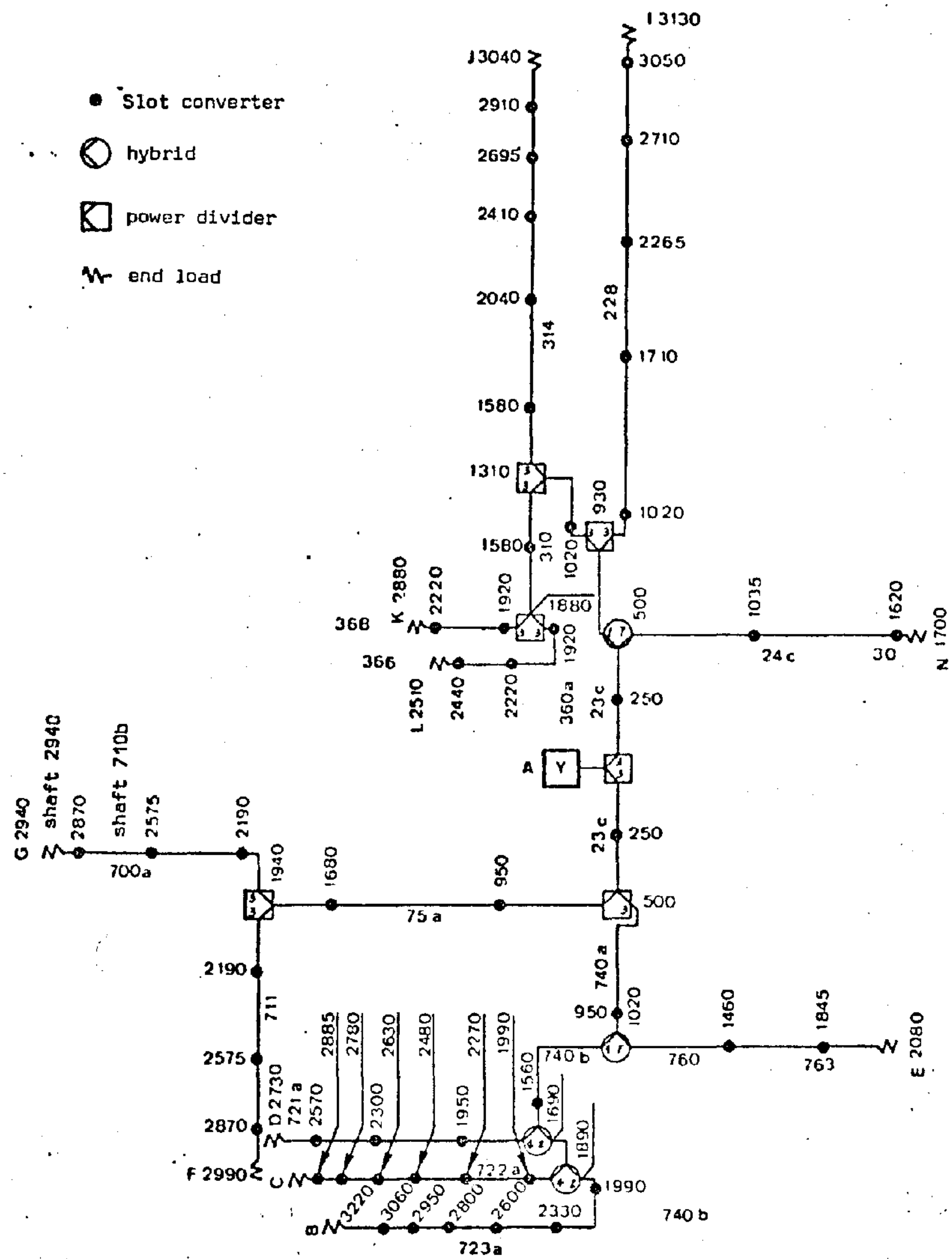


Fig. 6. Network in French potash mine with X-Y phones

QUESTIONS BY D.C. CHANG:

1. In what way is the optimization procedure confirmed? Is the polarization of the transmitter, as well as the difference between the monofilar and bifilar modal fields in the channel important in this calculation?

2. When the concept of mode converter is used, what is the trade-off between the maximum conversion and minimum mismatch (reflection of the bifilar mode)?

REPLIES BY P. DELOGNE:

1. The calculation procedure has been used successfully for about 5 years and the validity of the approximations if thus fully confirmed by experiments in about 100 km of tunnels. The fact that the leakage field of the coaxial (or bifilar) mode and the main field of the monofilar mode have nearly the same distribution and polarisation is confirmed by experiments and also by comparison with exact theoretical solutions published by Wait and Hill. The polarisation of the transmitter is thus important. We suppose, in our calculation that it has the correct orientation.

2. A mode converter is to be considered as a four-port network (with two ports on each side, one for each mode). Mode converters like, for instance, the annular slot converter use series loading and this imposes some constraints on the scattering matrix of the four-ports. In particular, it is impossible to have mode conversion without having some mismatch (see references 1 and 10 of my paper). For a conversion of less than 10 percent of the energy, like we do in practice, the mismatch is negligible (VSWR of the bifilar mode less than 1.2).

MODES IN BRAIDED COAXIAL CABLES IN CIRCULAR AND ELLIPTICAL TUNNELS*

D.B. SEIDEL AND J.R. WAIT**

Cooperative Institute for Research in Environmental Sciences
University of Colorado/NOAA
Boulder, CO 80309

Abstract-Radio frequency transmission in circular and semicircular tunnels each containing a braided coaxial cable is considered. The general formulation accounts for both the ohmic losses in the tunnel wall and a thin lossy film layer on the outer surface of the dielectric jacket of the cable. Using a quasistatic approximation, it is found that the propagation constants of the low-frequency transmission line modes are obtained through the solution of a cubic equation. However, for the special case when the conductivity thickness product of the lossy film layer

*Research supported by U.S. Bureau of Mines under contract No. H0155008.

**Consultant to the Institute for Telecommunication Sciences, U.S. Department of Commerce, Boulder, CO.

vanishes, this cubic equation reduces to a quadratic. The spatially dispersive form of the braid transfer impedance is also accounted for. It is shown that the quasi-static theory is well justified for frequencies as high as 100 MHz for typical tunnel geometries. Finally, these techniques are extended to a tunnel of elliptical cross-section. We find that, for frequencies up to 20 MHz, that the attenuation rate is relatively insensitive to the ellipticity if the cable-to-wall separation and the cross-sectional area are kept constant.

QUESTION BY R. GABILLARD:

You demonstrate rigorously that the eccentricity of an elliptical tunnel has little or no effect on the propagation constant of a leaky cable. This is an important and interesting result that we had found by experiment. But do you think it is possible to generalize this result to a tunnel of any cross-sectional shape? What are your views on this matter?

REPLY BY D.B. SEIDEL:

Although we have only specifically shown this correlation between cross-sectional area and propagation constant for an elliptical cross-section, we feel that at low frequency, this is most likely a general result. In support of this generalization, Ed Kuester has informed us that a similar low frequency cross-sectional dependence for dielectric waveguides of arbitrary cross-section can be shown.

REPLY BY J.R. WAIT:

Earlier calculations by Samir Mahmoud for rectangular shaped tunnels are also consistent with this conclusion.

USE OF AUXILIARY DEDICATED WIRE AS A MEANS OF AIDING CARRIER CURRENT PROPAGATION ON A TROLLEY WIRE/RAIL TRANSMISSION LINE

A.G. Emslie, R.L. Lagace, R.H. Spencer, and P.F. Strong
Arthur D. Little, Inc.
Cambridge, MA 02140

ABSTRACT

Underground coal mines with electric rail haulage almost always use carrier frequency communication to facilitate dispatch type communications with the rail vehicles. The transmission line formed by the trolley wire/rail is not well suited to this usage because of the many bridging loads, branches and unterminated line ends. As a result, poor transmission is a common complaint about such systems. A "dedicated" auxiliary wire deployed in the same tunnel and connected to the dispatcher's fixed transceiver can overcome much of the present deficiency by forming a low loss dedicated wire/rail transmission line which can carry the dispatcher's signal to vehicles in the far reaches of a mine. The natural electromagnetic coupling between the trolley wire/rail and the dedicated wire/rail is shown to be a good means of maintaining a useable received signal for the mobile transceiver units mounted on the mine vehicles and connected to the trolley wire/rail.

Theory is developed and shows that signal attenuation on the dedicated wire is largely dependent on its properties alone and is not markedly increased by losses due to its coupling to the lossy trolley wire/rail.

Supporting measurements made in an actual coal mine are discussed.

INTRODUCTION

The theory of radio frequency carrier current communication along an unloaded trolley wire/rail system used in a coal mine tunnel and operating at frequencies in the range of 50-800 kHz has recently been published. The authors,[1,2,3] using a wave theory approach, concluded that attenuation rates of the order of 1-2 dB/km are to be expected, and are due partly to the series resistance in the two conductors and partly to currents induced in the surrounding rock. Actual attenuation rates are much higher than these predicted values, owing to the shunting effect of low-resistance bridging loads including rectifiers, mine haulage vehicles, lights, pumps, and personnel heaters which are distributed at intervals along the trolley-wire/rail system. Experience shows that these discrete shunt loads result in an effective attenuation rate that is at least 10 dB/km and perhaps as high as 20 dB/km over long distances comparable to a wavelength.

A concept for greatly reducing the attenuation rate is predicated on the use of a thin, "dedicated" wire attached near the roof or on the wall of the tunnel which, along with the rail, provides a low-loss transmission line coupled weakly to the trolley-wire/rail transmission line. The idea is that the "dedicated" wire/rail transmission line, once it is excited either directly or by coupling to the trolley-wire/rail line, will act as a distributed power source and feed power back into the trolley-wire/rail line at great distances. This drastically reduces the signal attenuation rate along the trolley-wire/rail line, thereby significantly extending the range of communication between a fixed station dispatcher and operators of mobile rail haulage vehicles connected to the trolley wire via a trolley pole.

The purpose of this paper is to develop the theory of these two coupled transmission lines, with allowance both for the shunt losses in the trolley-wire/rail line and for the losses due to currents induced in the surrounding coal and rock, to determine the degree to which the loss rate of a high loss line can be reduced by mutual coupling to a nearby low loss line. Since the discrete shunt losses would be difficult to include in a wave theory analysis, we treat the problem by means of multi-conductor transmission line theory. The losses due to the rock are represented by appropriate resistances in series with the trolley-wire and the dedicated wire. These resistances are deduced from the loss rates given by Wait and Hill.[2] The discrete shunt loads placed across the trolley-wire/rail, typically at intervals short compared to wavelength, are approximated by a continuously distributed shunt loss across the trolley-wire/rail line that will produce the total loss equivalent to that produced by the discrete shunt loads over a long length of the line.

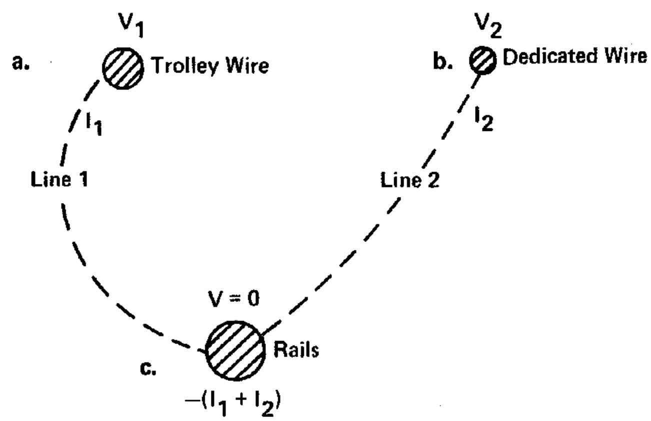
This work was supported by the U.S. Department of the Interior, Bureau of Mines, Pittsburgh Mining and Safety Research Center, under USBM Contract HO346045.

The results of the calculations show that ranges of the order of 20 km should be possible with the help of the dedicated wire, in contrast to ranges on the order of 4 km in the absence of the dedicated wire or other means of extending range.

MODE ANALYSIS

Figure 1 is a schematic diagram of the three-conductor transmission system, which can be regarded as two coupled transmission lines labeled 1 and 2. Line 1, with voltage V_1 and current I_1, is the trolley-wire/rail transmission line. Line 2, with voltage V_2 and current I_2, is the dedicated wire/rail transmission line. The electrically interconnected rails are represented by a single cylindrical conductor which also carries the return current $-(I_1 + I_2)$.

FIGURE 1
THE THREE-CONDUCTOR TRANSMISSION SYSTEM



The voltages and currents are related by the following first-order differential equations:

$$\frac{dV_1}{dz} = -Z_{11}I_1 - Z_{12}I_2 \qquad (1)$$

$$\frac{dV_2}{dz} = -Z_{21}I_1 - Z_{22}I_2 \qquad (2)$$

$$\frac{dI_1}{dz} = -Y_{11}V_1 - Y_{12}V_2 \qquad (3)$$

$$\frac{dI_2}{dz} = -Y_{21}V_1 - Y_{22}V_2 \qquad (4)$$

where z is the distance along the transmission lines and the Z's and Y's are impedances and admittances per unit length that include (1) effective series resistances per unit length due to currents induced in the nearby conducting coal and rock, and (2) the average shunt conductance per unit length due to bridging loads connected across the trolley-wire/rail transmission line. The solution of equations (1) – (4) leads to the mode equation for the propagation constant γ:

$$\begin{vmatrix} Z_{11}Y_{11} + Z_{12}Y_{12} - \gamma^2 & Z_{11}Y_{12} + Z_{12}Y_{22} \\ Z_{22}Y_{12} + Z_{12}Y_{11} & Z_{22}Y_{22} + Z_{12}Y_{12} - \gamma^2 \end{vmatrix} = 0 \tag{5}$$

The expressions for the currents in the two lines can be shown to be:

$$I_1 = \frac{1}{Z_{11}Z_{22} - Z_{12}^2}\left[(Z_{22} - r_+Z_{12})\gamma_+Ae^{-\gamma_+z} + (Z_{22} - r_-Z_{12})\gamma_-Be^{-\gamma_-z}\right] \tag{6}$$

and

$$I_2 = \frac{1}{Z_{11}Z_{22} - Z_{12}^2}\left[(r_+Z_{11} - Z_{12})\gamma_+Ae^{-\gamma_+z} + (r_-Z_{11} - Z_{12})\gamma_-Be^{-\gamma_-z}\right] \tag{7}$$

where γ_+ and γ_- are the roots of Eq. (5), and A, B are arbitrary constants.

Two cases are of interest:

(1) where the transmitter applies a voltage V_o at $z = 0$ between the trolley-wire and the rail, and

(2) where V_o is applied between the "dedicated" wire and the rail.

For the transmitter across line 1, the boundary conditions at $z = 0$ are:

$$V_1 = V_o \tag{8}$$

and

$$I_2 = 0. \tag{9}$$

On applying these boundary conditions one finds for the voltages and currents on the two lines:

$$\frac{V_1}{V_0} = \frac{\gamma_-(r_-Z_{11} - Z_{12})e^{-\gamma_+z} - \gamma_+(r_+Z_{11} - Z_{12})e^{-\gamma_-z}}{\gamma_-(r_-Z_{11} - Z_{12}) - \gamma_+(r_+Z_{11} - Z_{12})} \tag{10}$$

$$\frac{V_2}{V_0} = \frac{\gamma_- r_+ (r_- Z_{11} - Z_{12}) e^{-\gamma_+ z} - \gamma_+ r_- (r_+ Z_{11} - Z_{12}) e^{-\gamma_- z}}{\gamma_- (r_- Z_{11} - Z_{12}) - \gamma_+ (r_+ Z_{11} - Z_{12})} \tag{11}$$

$$\frac{I_1}{V_0} = \frac{\gamma_+ \gamma_-}{\left(Z_{11} Z_{22} - Z_{12}^2\right)} \left[\frac{(Z_{22} - r_+ Z_{12})(r_- Z_{11} - Z_{12}) e^{-\gamma_+ z} - (Z_{22} - r_- Z_{12})(r_+ Z_{11} - Z_{12}) e^{-\gamma_- z}}{\gamma_- (r_- Z_{11} - Z_{12}) - \gamma_+ (r_+ Z_{11} - Z_{12})} \right] \tag{12}$$

$$\frac{I_2}{V_0} = \frac{\gamma_+ \gamma_- \left(r_+ Z_{11} - Z_{12}\right)\left(r_- Z_{11} - Z_{12}\right)\left(e^{-\gamma_+ z} - e^{-\gamma_- z}\right)}{\left(Z_{11} Z_{22} - Z_{12}^2\right)\left[\gamma_- \left(r_- Z_{11} - Z_{12}\right) - \gamma_+ \left(r_+ Z_{11} - Z_{12}\right)\right]} \tag{13}$$

where

$$r_\pm = \frac{\gamma_\pm^2 - Z_{11} Y_{11} - Z_{12} Y_{12}}{Z_{11} Y_{12} + Z_{12} Y_{22}} \tag{14}$$

Similar equations are found for case 2.

CALCULATION OF THE IMPEDANCES AND ADMITTANCES

Inductance per Unit Length

We denote the radii of the trolley-wire, the dedicated-wire, and the rail by a, b, and c, respectively, and also use the same letters as subscripts to indicate these conductors. Equations (1) – (4) define the Z's and Y's. For example, Z_{11} is given by the formula

$$Z_{11} = -\frac{1}{I_1}\left(\frac{dV_1}{dz}\right)_{I_2=0} = R_a + i\omega L_{11} \tag{15}$$

where R_a is the effective resistance per unit length of the trolley-wire and L_{11} is the magnetic flux that links unit length of line 1 for currents of +1 ampere in the trolley-wire and −1 ampere in the rail. Since the radii a and c are small compared with the distance S_{ac} between the axis of the two conductors, we can write

$$L_{11} = \frac{\mu_o}{2\pi} \ln\left(\frac{S_{ac}^2}{ac}\right) \tag{16}$$

where H_a and H_c are the magnetic fields due to the unit currents in wires a and c and r, r′ are distances from the axes of wires a and c. Likewise:

$$Z_{22} = R_b + i\omega L_{22} \tag{17}$$

where

$$L_{22} = \frac{\mu_o}{2\pi} \ell n\left(\frac{S_{bc}^2}{bc}\right) \qquad (18)$$

From Equation (1), the mutual impedance Z_{12} is defined as

$$Z_{12} = -\frac{1}{I_2}\left(\frac{dV_1}{dz}\right)_{I_1=0} = i\omega L_{12} \qquad (19)$$

where L_{12} is the magnetic flux linking line 1 due to +1 ampere in wire b and −1 ampere in wire c, which can be shown to be:

$$L_{12} \cong \frac{\mu_o}{2\pi} \ell n\left(\frac{S_{bc}}{S_{ba}}\right) + \frac{\mu_o}{2\pi} \ell n\left(\frac{S_{ac}}{c}\right) = \frac{\mu_o}{2\pi} \ell n\left(\frac{S_{bc}S_{ac}}{cS_{ab}}\right) . \qquad (20)$$

Capacitance per Unit Length

From Equations (3) and (4), we find that

$$Y_{11} = -\frac{1}{V_1}\left(\frac{dI_1}{dz}\right)_{V_2=0} = G_{ac} + i\omega C_{11} \qquad (21)$$

$$Y_{21} = -\frac{1}{V_1}\left(\frac{dI_2}{dz}\right)_{V_2=0} = i\omega C_{21} \qquad (22)$$

where G_{ac} is the shunt conductance per unit length across line 1, and C_{11} and C_{21} are the charges per unit length induced on wires a and b by unit potential on wire a, while wires b and c are held at zero potential. The capacitance can be shown to be:

$$C_{11} = \frac{q_1}{V_1} = \frac{2\pi\epsilon_o}{\Delta} \ell n\left(\frac{S_{bc}^2}{bc}\right) \qquad (23)$$

$$C_{21} = \frac{q_2}{V_1} = -\frac{2\pi\epsilon_o}{\Delta} \ell n\left(\frac{S_{ac}S_{bc}}{S_{ab}c}\right) \qquad \text{where } \Delta = \ell n\left(\frac{S_{ac}^2}{ac}\right)\ell n\left(\frac{S_{bc}^2}{bc}\right) - \left(\ell n \frac{S_{ac}S_{bc}}{S_{ab}c}\right)^2 \qquad (24)$$

By reciprocity, Equation (24) also gives the value of C_{12}. By symmetry we find from Equation (23) that

$$Y_{22} = i\omega C_{22} \tag{25}$$

where

$$C_{22} = \frac{2\pi\epsilon_o}{\Delta} \ln\left(\frac{S_{ac}^2}{ac}\right). \tag{26}$$

Series Resistance per Unit Length

The series resistance R_b of transmission line 2, acting alone in the tunnel, is chosen so that the attenuation constant a_2 lies in the range calculated by Wait and Hill.[2] The connection between R_b and a_2 is given by the transmission line formula:

$$a_2 + i\beta_2 = \left\{\left(R_b + i\omega L_{bc}\right) i\omega C_{bc}\right\}^{1/2} \tag{27}$$

where

$$L_{bc} = \frac{\mu_o}{2\pi} \ln\left(\frac{S_{bc}^2}{bc}\right) \tag{28}$$

and

$$C_{bc} = 2\pi\epsilon_o / \ln\left(\frac{S_{bc}^2}{bc}\right). \tag{29}$$

On separating the real and imaginary parts of Equation (27) and eliminating β_2, we find that:

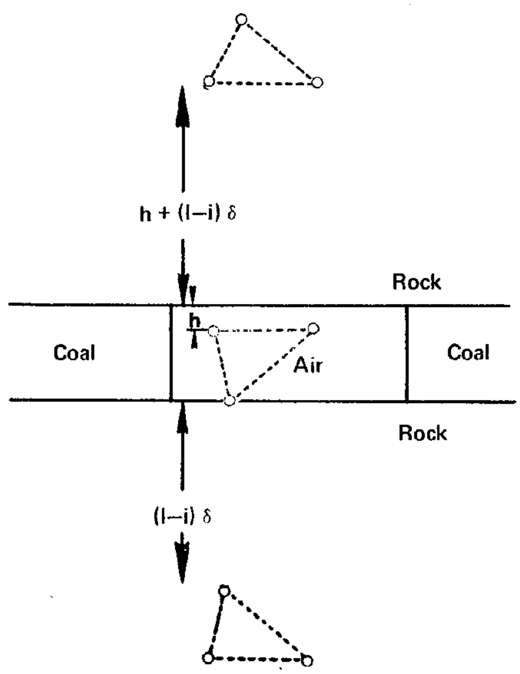
$$R_b = \frac{2a_2\left(a_2^2 + \beta_o^2\right)^{1/2}}{\omega C_{bc}} \tag{30}$$

where $\quad \beta_o = \omega / c_o \qquad (31)$

and c_o is the speed of light in free space. Since transmission lines 1 and 2 have comparable geometry relative to the rock, we assume that $R_a = R_b$.

In the foregoing analysis we have assumed that the only significant effect of the rock surrounding the three conductors is to induce series resistances in the conductors. This assumption is based on the relatively distant locations of the electrical images of the conductors. Figure 2 shows the positions of the nearest electrical images[5] in the rock for the case $\omega\epsilon << \sigma$ that applies in the present situation.

FIGURE 2
ELECTRIC IMAGES IN THE ROCK



Since the skin depth δ in the rock at 100 kHz for a typical rock conductivity of 0.1 Mho/m is about 5 m, the absolute value of the distance from any conductor to the closest image is 7 m. This is sufficiently large compared with the tunnel height of about 2 m that the free-space calculations of the inductances and capacitances should be approximately valid. However, since the actual resistances of the conductors are very low, the induced resistances caused by the images in the rock cannot be neglected.

Since the conductivity of coal is several orders of magnitude less than that of the rock, the images in the coal seam are even farther away, making them less important than the images in the rock.

Shunt Conductance per Unit Length

The effective shunt conductance G_{ac} per unit length of line 1 may be calculated from the transmission line equation

$$\alpha_1 + i\beta_1 = \left\{\left(R_a + i\omega L_{ac}\right)\left(G_{ac} + i\omega C_{ac}\right)\right\}^{1/2} \qquad (32)$$

where

$$L_{ac} = \frac{\mu_o}{2\pi} \ln\left(\frac{S_{ac}^2}{ac}\right) \tag{33}$$

and

$$C_{ac} = 2\pi\epsilon_o / \ln\left(\frac{S_{ac}^2}{ac}\right) \tag{34}$$

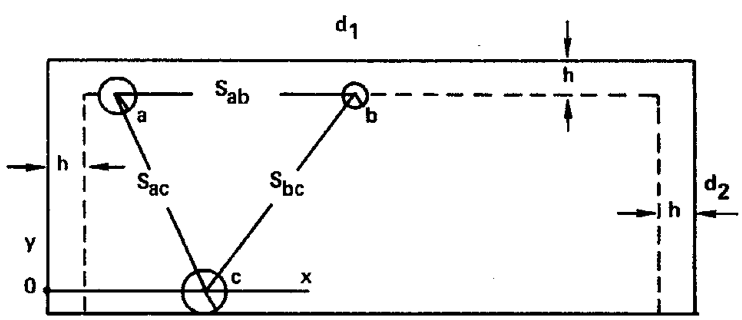
On separating real and imaginary parts of Equation (32) and eliminating β_1, we obtain the expression

$$G_{ac} = \left[\left\{R_a\left(\frac{1}{c_o^2} + \frac{2\alpha_1^2}{\omega^2}\right)\Big/L_{ac}^2\right\}^2 + \left\{\frac{4\alpha_1^2\left(\alpha_1^2 + \beta_o^2\right)}{\omega^2} - R_a^2 C_{ac}^2\right\}\Big/L_{ac}^2\right]^{1/2} - R_a\left(\frac{1}{c_o^2} + \frac{2\alpha_1^2}{\omega^2}\right)\Big/L_{ac}^2 \tag{35}$$

for the determination of G_{ac} for any given value of the attenuation constant α_1 of transmission line 1 acting alone. Since the dedicated wire is assumed to be insulated from the rail, we assume that $G_{bc} = 0$.

COMPUTATION AND DISCUSSION OF RESULTS

FIGURE 3
CROSS SECTION OF TUNNEL, OF DIMENSIONS $d_1 \times d_2$



Computation

Figure 3 shows the locations of the three conductors relative to the tunnel cross section. We assume the following dimensions in meters:

$d_1 = 5$, $d_2 = 2.5$, $h = 0.3$, $a = 0.015$, $b = 0.0015$, $c = 0.1$, $x_a = 0.5$, $y_a = 2.2$, $x_c = 1.25$, $y_c = 0$.

The "dedicated" wire is placed at x = 2.2, y = 2.2 for calculations of signal strength versus z, and placed at various points on the dotted contour in the tunnel cross section in Figure 3 for calculations of signal strength at fixed distance z and various locations of the wire on the contour. A time-sharing computer system (On-line Systems, Inc.) was used to evaluate the currents and voltages as functions of z for a variety of values of the parameters and two excitation conditions. The results obtained are discussed below.

DISCUSSION OF RESULTS

Results of our investigation[5] of the potential improvements to be gained by the introduction of a "dedicated" wire parallel to a high-loss trolley-wire/rail transmission line are shown in Figures 4 through 10. These figures illustrate the performance capabilities to be expected by employing the dedicated wire to extend the range of trolley-wire carrier phone communication systems in mines. Figure 4 illustrates the way in which voltages decrease with distance from transmitters in such a system. It should be noted that when the transmitter voltage is applied between the trolley-wire and the rail, severe attenuation of the signal occurs with distance due to the multitude of loss mechanisms present across the trolley-wire/rail transmission line. For the example shown in Figure 4, a net attenuation rate of 10 dB/km is assumed for the trolley-wire/rail transmission path. The distances covered in this plot run to 20 km. Thus, the total signal loss would be approximately 200 dB on the trolley-wire/rail in the absence of the dedicated wire. For most trolley carrier phones a practical loss of between 60 and 70 dB can be tolerated depending on ambient noise conditions on the trolley-wire/rail.

FIGURE 4
VOLTAGE LEVELS VERSUS DISTANCE
α_T = 10 dB/km
α_D = 1 dB/km

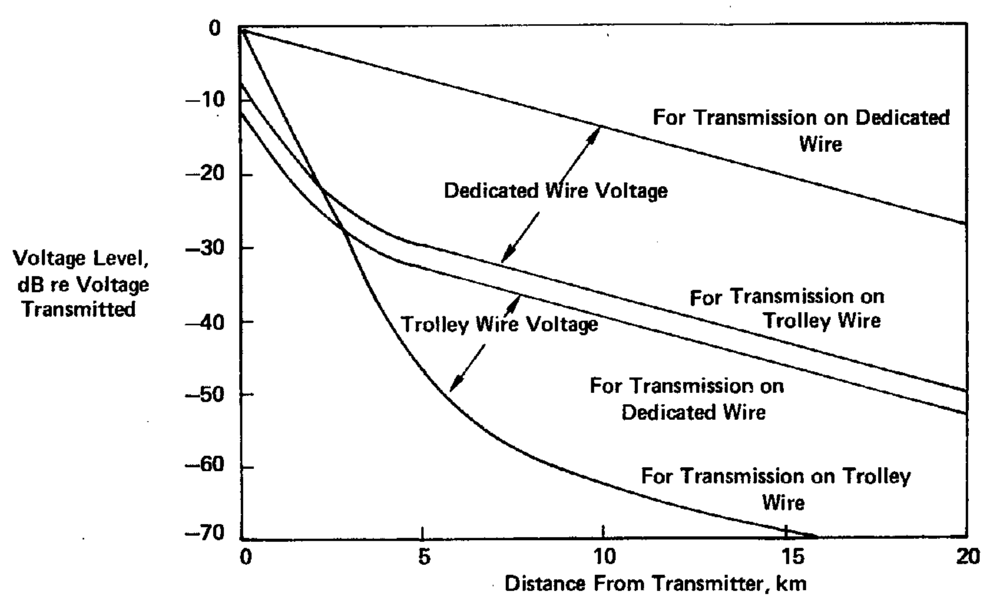


Figure 4 presents signal levels on both transmission lines as a function of distance, for transmitters feeding the trolley-wire/rail line such as would be the case for tracked vehicles in the mine, and for transmitters feeding the dedicated wire/rail line which could represent the feed arrangement for a dispatcher's transmitter. Figure 4 shows that, at the end of 20 km, the loss from the dispatcher to a vehicle and the loss from vehicle to dispatcher are 53 dB and 49.5 dB, respectively. Thus, an improvement of 150 dB in signal level is achieved over conventional trolley-wire/rail transmission line systems in which both vehicle and dispatcher's carrier phones are connected across the trolley-wire/rail and no parallel wire or other signal-extending means is present. This figure also illustrates the fact that when beyond the "initial condition" region caused by the discrete source transmitter, signals on both lines attenuate at a uniform and low rate closer in value to the rate of the dedicated wire/rail line alone. In addition, Figure 4 shows that even conventional trolley-line communication, i.e., transmission and reception on the trolley-wire/rail line between two vehicles or a vehicle and the dispatcher, will be markedly improved just by the presence of a dedicated wire even though no signals are directly impressed onto the dedicated wire.

FIGURE 5
VOLTAGE LEVELS VERSUS DISTANCE
$\alpha_T = 20$ dB/km
$\alpha_D = 1.0$ dB/km

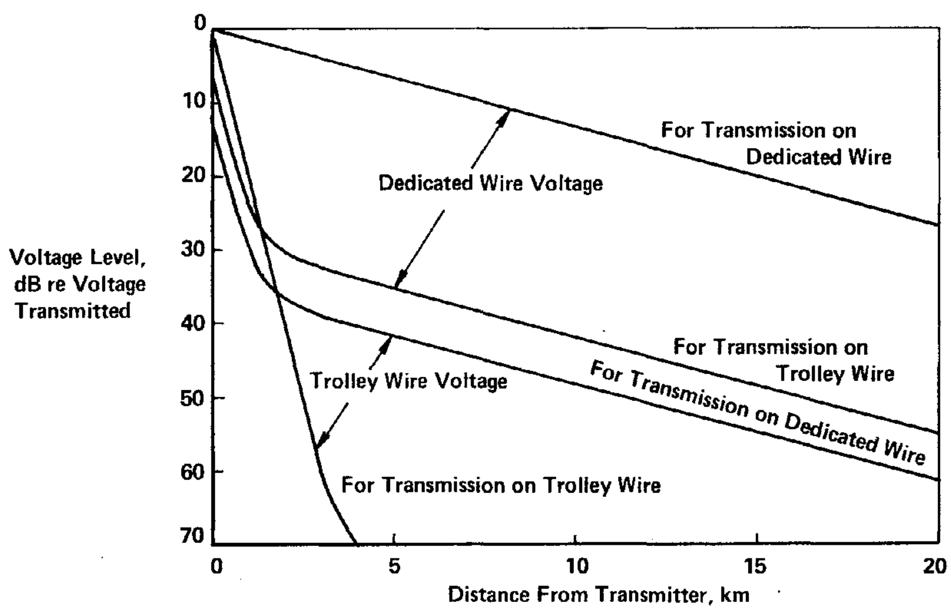


Figure 5 presents results for the same conditions of operation as shown in Figure 4, except for the fact that a higher attenuation rate of 20 dB/km is assumed for the trolley-wire/rail line. In this instance it can be seen that the signal levels resulting from transmission on the dedicated wire/rail line and reception on the trolley-wire/rail line, and vice versa, show only modestly increased losses over those for the 10-dB loss rate of Figure 4. This illustrates the fact that the dominant feature in controlling the final attenuation rate is the quality of or loss associated with the dedicated line. The signal levels at the end of the 20-km test section are found to exhibit between 55 and 62 dB of loss, completely practical values for useful carrier phone systems.

Figures 6 and 7 show plots of the current levels in the trolley-wire and in the "dedicated" wire versus distance for the same conditions as those of Figures 4 and 5, respectively. Little comment is needed concerning the characteristics of these except to note that at zero distance one expects to find zero current in the line to which the transmitter is not connected; e.g., if the transmission is on the "dedicated" line, zero current will be found on the trolley-wire at the transmitter position. It is seen that the current on the line without the transmitter builds up to a maximum value within a relatively short distance and then decays at the uniform attenuation rate of the coupled transmission line system.

FIGURE 6
CURRENT LEVEL VERSUS DISTANCE
α_T = 10 dB/km
α_D = 1 dB/km

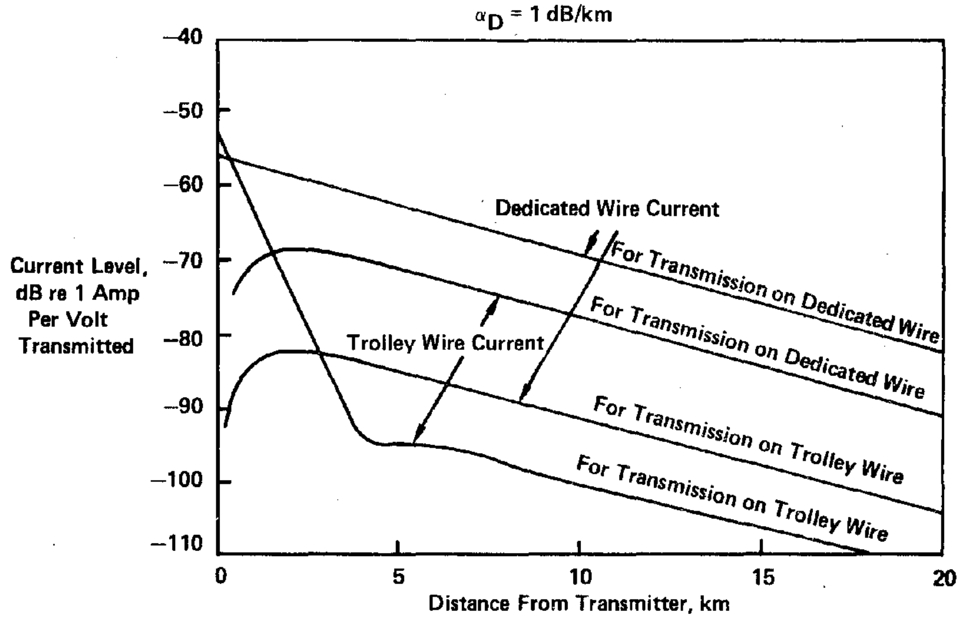
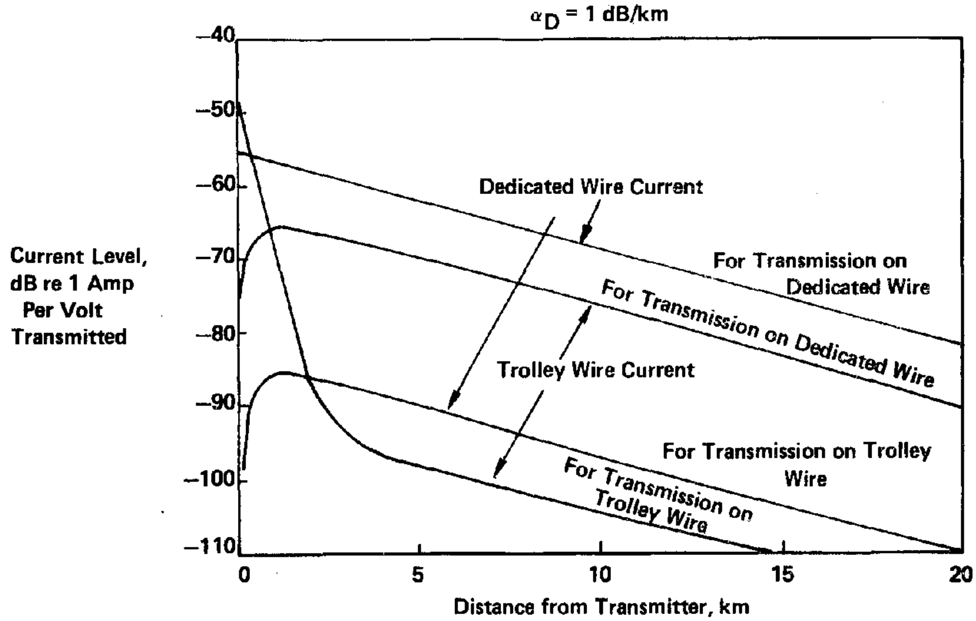


FIGURE 7
CURRENT LEVELS VERSUS DISTANCE
α_T = 20 dB/km
α_D = 1 dB/km



One of the key questions we had when we undertook this investigation was, "Where is the optimum location for a 'dedicated' wire to obtain maximum extension of communication range?" Figures 8 and 9 illustrate what happens when the dedicated wire is moved along a grid representing the possible locations within the cross section of a typical rail haulage way tunnel in a mine. We define this set of possible locations to be those at 30 cm spacing from the ribs and from the roof as illustrated in Figure 3. Figure 8 shows the values of voltages found at the far end of the transmission line system, 20 km away, as a function of the location of the dedicated wire on this grid. Figure 8 assumes that the transmitter is connected across the trolley-wire/rail line and looks at trolley-wire/rail voltages and "dedicated" wire/rail voltages for two conditions of trolley-wire/rail attenuation rate. As can be seen, there is no sharp maximum of signal level. Figure 8 also shows that one should stay away from locations near the trolley-wire itself, and suggests that the wide side of the tunnel be used for location of the dedicated wire. In fact, it will be necessary to locate the dedicated wire on the wide side of the tunnel because of existing mine safety laws. Figure 9 presents similar plots for the case where the transmitter is connected across the "dedicated" wire/rail line. One noteworthy feature is that when the transmitter is connected across the "dedicated" wire/rail, the "dedicated" wire/rail voltage 20 km away is hardly influenced by the value of trolley-wire/rail loss, and the two curves that correspond to 10 dB and 20 dB/km are virtually coincident.

FIGURE 8
VOLTAGE LEVELS 20 KM FROM TRANSMITTER ON TROLLEY WIRE AS A FUNCTION OF DEDICATED WIRE POSITION (SEE FIG. 3)

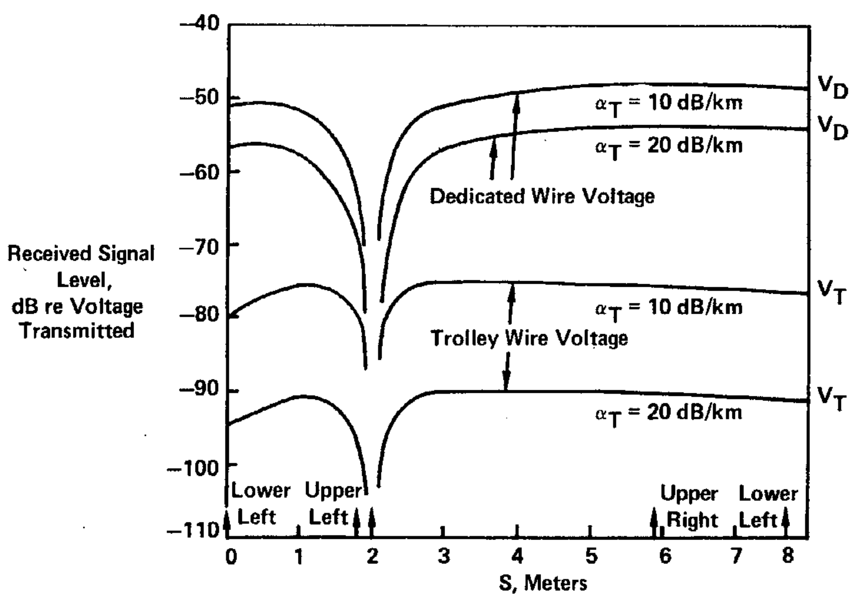


FIGURE 9
VOLTAGE LEVELS 20 KM FROM TRANSMITTER ON DEDICATED WIRE AS FUNCTION OF DEDICATED WIRE POSITION (SEE FIG. 3)

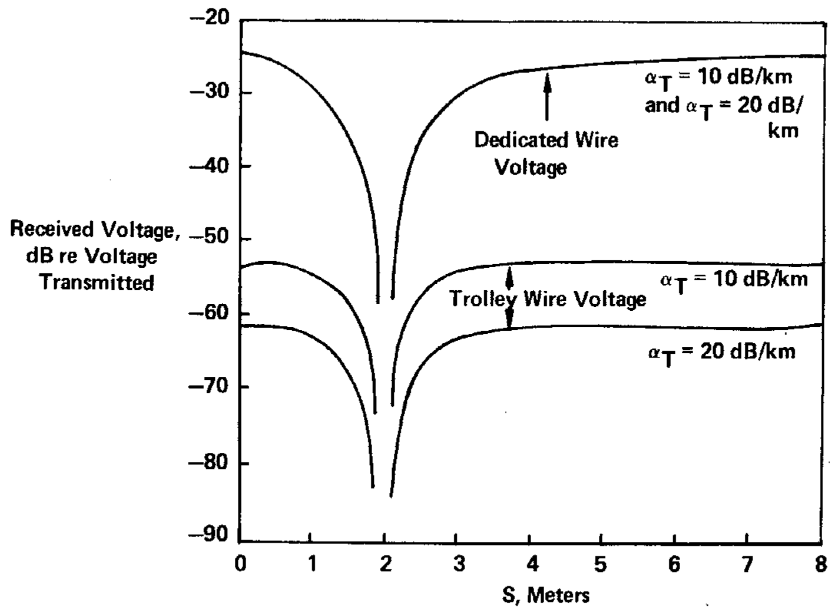
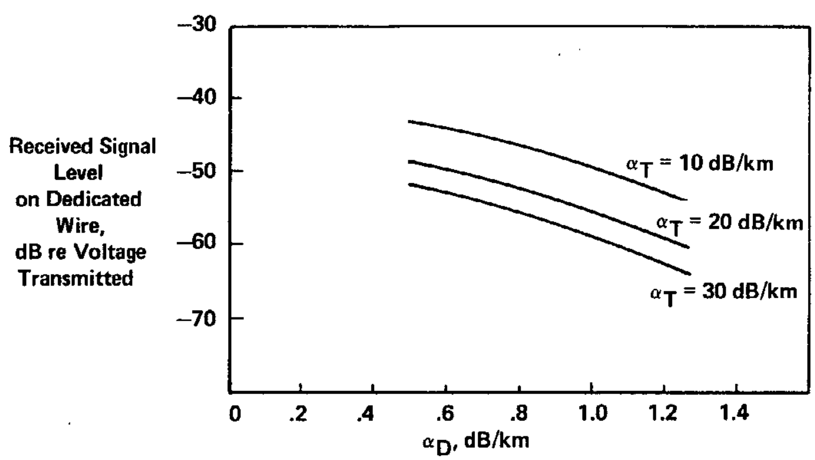


Figure 10 illustrates how the received voltage level 20 km from a transmitter on the trolley-wire/rail depends on the attenuation rate in dB/km of the "dedicated" wire/rail line, and on the attenuation rate of the trolley-wire/rail line. It further illustrates the minimal return one obtains by trying to improve signal level by decreasing the shunt loading on the trolley-wire/rail. Also illustrated is the way in which signal levels vary when the series attenuation rate of the "dedicated" wire/rail is varied. It shows that there is little utility to using an extremely low-loss "dedicated" wire line, since asymptotic values of 40 and 50 dB of loss are present even for dedicated wire lines having no loss.

FIGURE 10
VOLTAGE SIGNAL LEVEL 20 KM FROM TRANSMITTER ON TROLLEY WIRE



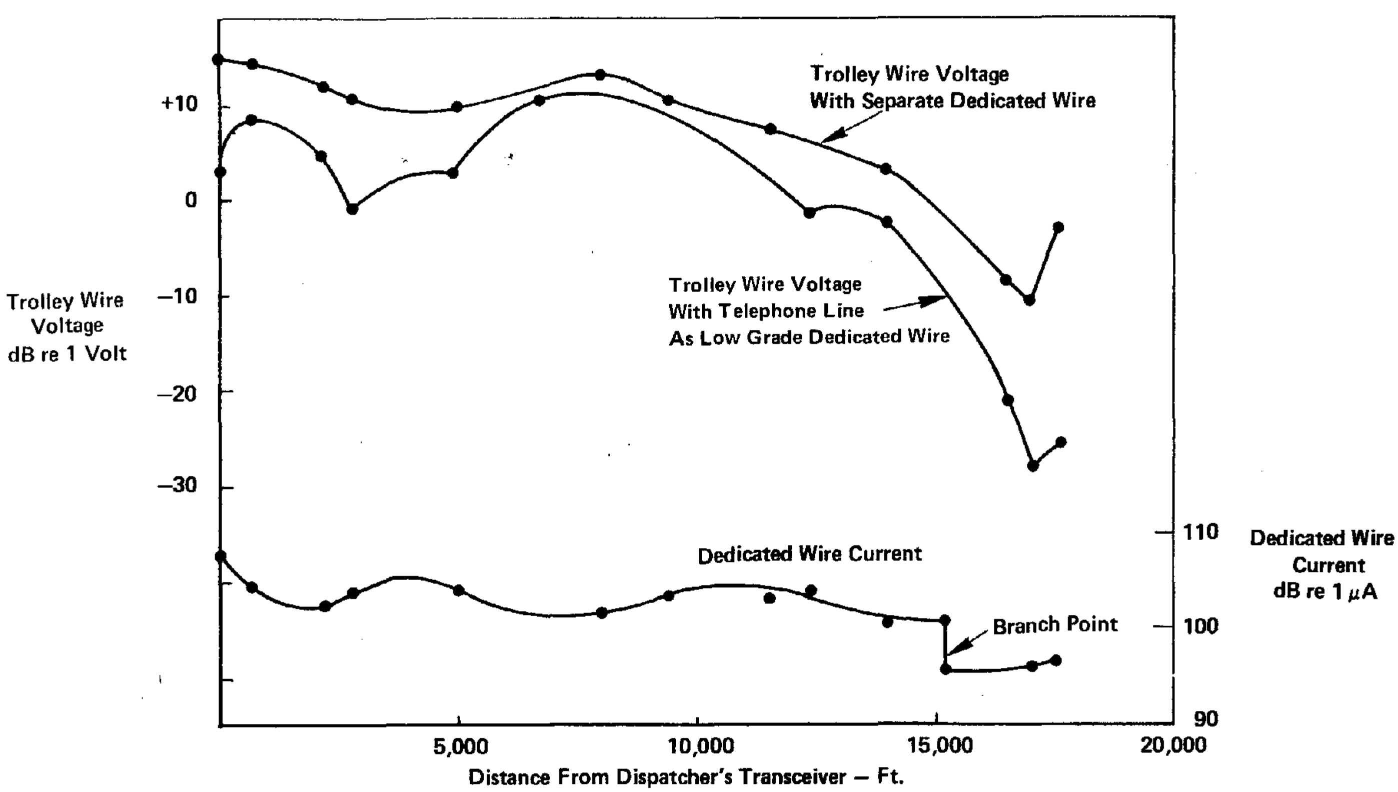
EXPERIMENTAL RESULTS

A dedicated wire was installed at the Montour No. 4 coal mine to test the theory of the dedicated wire. It was found that low loss signal transmission occurred on the dedicated wire and that matched signal splits on the dedicated wire could be used to provide coverage on branches. The length of the dedicated wire was 5.5 km (18,000 ft.). At the end of a branch where vehicle to dispatcher communication had been extremely poor the branch dedicated wire provided operational coverage. The enhancement of signal levels from the dispatcher's transceiver to the vehicles improved over the 5.5 km section of rail haulage by as much as 22 dB. The plot of Figure 11 illustrates an example of this improvement. This plot shows the signal level on the trolley-wire/rail as a function of distance from the dispatcher's transceiver for the original installation that used the phone line as a low grade dedicated wire and for an installation of a true dedicated wire.

CONCLUSIONS

This investigation of the utility of a parallel low loss "dedicated" wire shows that it can be used to extend the vehicle to dispatcher voice communication range on a high loss trolley-wire/rail carrier phone communication system. The investigation shows that the ability of such a dedicated wire to extend range (except very close to the trolley wire) is not markedly influenced by its position relative to the trolley-wire/rail, the walls, nor the attenuation rate on the trolley-wire/rail line. Measurements made in an actual coal mine confirm the expectations provided by the theory. The recent work of Wait[6] which predicts a negligible loading influence on the dedicated wire of discrete loads on the trolley wire is confirmed by the experimental results.

FIGURE 11
SIGNAL LEVELS VS. DISTANCE



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Analysis of the Dedicated Communication Line in a Mine Tunnel for a Shunt-Loaded Trolley Wire

JAMES R. WAIT*

Cooperative Institute for Research in Environmental Sciences
University of Colorado/NOAA
Boulder, Colorado 80309

DAVID A. HILL

Institute for Telecommunication Sciences
Office of Telecommunications
U.S. Department of Commerce
Boulder, Colorado 80302

Abstract-The theory of transmission in a semi-circular tunnel containing two axial conductors is simplified for the case of radio frequencies. It is shown that the effect of ohmic losses in the wall can be represented as series impedances per unit length in the equivalent coupled transmission line circuits. This opens the way for a systematic calculation of the mode conversion phenomena that occur when one of the axial conductors has a discrete shunt load to the ground plane. Explicit formulas are obtained for the modal conversion coefficients. These are used to discuss the effect of a shunt load on the trolley wire on the performance of a radio frequency communication system. It is confirmed that the unloaded dedicated communication line provides for a low loss mode that is hardly affected by the shunt load on the adjacent trolley wire in the tunnel.

This research was supported by USBM Contract No. H0155008.

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*Full paper published in *IEEE Trans. on Communications*, Vol. COM-26, No. 3, pp. 355-361, March 1978.

Role of Controlled Mode Conversion in Leaky Feeder Mine-Communication Systems

DAVID B. SEIDEL and JAMES R. WAIT
Cooperative Institute for Research in Environmental Sciences
University of Colorado/NOAA
Boulder, CO 80309

Abstract-Using an idealized model, we consider VHF wave transmission in a straight mine tunnel that contains a braided coaxial cable. The controlled interchange of energy between the propagation modes can be used to optimize the communication range and allow for coupling to and from portable hand-held antennas. In particular, we show that the optimum lengths for maximum transmission through the leaky cable sections or mode converters predicted here are compatible to those predicted by approximate coupled mode theory.

INTRODUCTION

Recently, the leaky feeder concept has been utilized effectively in intra-mine communications [1,2]. It had been previously found that such a system using a braided coaxial cable will propagate two low frequency transmission line modes [3]. One of these eigen modes (defined here as the monofilar mode) is readily excited from an arbitrary point within the tunnel but suffers high attenuation because the return current flows mainly in the tunnel wall. Conversely, the other eigen mode (defined here as the bifilar mode) has low attenuation because the return current flows mostly in the sheath; but it is poorly excited for an external source in the tunnel. Thus, for an efficient limited access communication system, it is desirable to convert energy from one mode to the other. Such mode conversion can be obtained through insertion of axial nonuniformities in the coaxial line or within the tunnel environment. In this paper, transmission line techniques are used to develop a theory of controlled mode conversion for a braided coaxial cable in a semi-circular tunnel. The formulation will also account for the ohmic losses of the tunnel wall as well as the spatially dispersive form of the braid transfer impedance.

The specific type of non-uniformity considered here is when the cable has a section of higher leakage. The case of a number of such uniformly spaced sections is also analyzed. This configuration can be used in practical mine-communication systems wherein the roving miners may use portable walkie-talkie type transceivers.

FORMULATION

The model assumed is described in terms of a cylindrical coordinate system (ρ,ϕ,z) and is shown in Fig. 1. The tunnel wall is located at $\rho = a_o$ for $0 < \phi < \pi$, and the assumed perfectly conducting tunnel floor is located at $\phi = 0$ and $\phi = \pi$ for $0 < \rho < \infty$. The region defined by $\rho > a_o$ and $0 < \phi < \pi$ is a homogeneous medium with conductivity σ_e and permittivity ε_e. The coaxial cable with outside radius c is centered at $\rho = \rho_o$ and $\phi = \phi_o$. The region defined by $\rho < a_o$, $0 < \phi < \pi$ and $\rho' > c$, where ρ' is the radial component of a cylindrical coordinate system (ρ',ϕ',z) centered at (ρ_o,ϕ_o), is described by the free space permittivity and permeability ε_o and μ_o, respectively. The geometry of the coaxial cable is shown in Fig. 2. The inner conductor, of radius a, has a high but finite conductivity σ_w. The surrounding insulation of radius b is a lossless dielectric with permittivity ε. The braided sheath located at $\rho' = b$ is characterized by a surface transfer impedance Z_T. The outer dielectric coating has radius c and permittivity ε_c. We also allow for the possibility that a thin lossy film is located at $\rho' = c$ which is characterized by a transfer impedance Z_L. We assume that the fields of each mode of this structure vary as $\exp(-\Gamma z + i\omega t)$ where ω is the angular frequency and Γ is the complex propagation constant for the particular mode.

In a previous paper [4], we obtained a quasistatic mode equation for the low frequency quasi-TEM modes for this structure. There it was found that, in general, there are three distinct modes of this type. The corresponding propagation constants Γ_j were found to be the roots of a cubic equation with complex coefficients. In the special case where the lossy film layer is absent, there are only two modes, whose propagation constants are obtained from a quadratic equation. In each case, two modes can be identified as the well-known monofilar and bifilar modes. The third mode, when it exists, is referred to as the jacket mode [4].

Because the dominant low frequency modes of this system behave like TEM transmission-line modes, it is useful to treat the problem as a multi-conductor transmission line using transmission line concepts. For this purpose, characteristic impedances have been previously derived [4] for each line. Slightly modified expressions for those impedances are given as follows:

$$K_L = \left(\frac{1 + P + Q}{Q}\right) \frac{\Gamma \ln R}{N} ,$$

$$K_s = \Gamma\left[(1+P+Q)\frac{\ln R}{N} + (1+P)\alpha_c\right] , \tag{1}$$

and

$$K_i = \Gamma\left[\left(\frac{1 + P + Q}{P}\right)\frac{\ln R}{N} + \left(\frac{1 + P}{P}\right)\alpha_c + \alpha'\right]$$

are the impedances associated with the lossy film, the braided sheath, and the inner conductor, respectively. The current ratios P and Q are defined by

$$P \equiv \frac{I_i}{I_s} = \frac{Z_T(\Gamma)}{Z'(\Gamma) + Z_i} \tag{2}$$

and

$$Q \equiv \frac{I_L}{I_s} = (1+P) \left[\frac{Z_b(\Gamma) + Z_c(\Gamma)}{Z_L}\right]$$

where I_L, I_s, I_i are the currents on each conductor (see Fig. 3). Other factors in (1) and (2) are defined in [4].

The characteristic impedances defined by (1) are dependent upon Γ and consequently they assume different values for each mode. Thus, we have $K_L^{(j)}$, $K_s^{(j)}$, and $K_i^{(j)}$ corresponding to the propagation constant of each mode Γ_j for $j = 1,2,3$. In the special case when there is no lossy film (i.e. $\sigma d = 0$), I_L vanishes and we consider only the two characteristic impedances $K_s^{(j)}$ and $K_i^{(j)}$ for $j = 1,2$.

These characteristic line impedances can be used in conjunction with transmission line concepts to consider axial nonuniformities or discontinuities in the tunnel/cable system. For example, the cable could be loaded by discrete shunt or series loads. Alternatively, there could be a short length of the tunnel in which tunnel and/or cable parameters change value, either intentionally or non-intentionally. Here we will specifically consider the case in which the value of the braid transfer impedance is changed over a short length (ℓ) along the line. This corresponds in some sense to the mode converters suggested for use in tunnels by Deryck [5] and Delogne [1] who employed approximate coupled mode theory. Our multi-mode formulation is similar to that used by Wait and Hill [6] for analyzing the shunt-loading on a two-wire system in a mine tunnel.

The geometry for this case is shown schmatically in Fig. 4. In regions I and III, we have a sheath transfer inductance L_T and the transmission characteristics in these regions are given by Γ_j, P_j, Q_j, $K_L^{(j)}$, $K_s^{(j)}$, and

$K_i^{(j)}$. Similarly, region II is characterized by the sheath transfer inductance $\hat{L}_T$ as well as $\hat{\Gamma}_j$, $\hat{P}_j$, $\hat{Q}_j$, $\hat{K}_L^{(j)}$, $\hat{K}_s^{(j)}$, and $\hat{K}_i^{(j)}$. The line is excited by incident kth mode which impinges from the negative z direction. If we think of this as a junction problem, we will correspondingly expect a reflected wave in region I and a transmitted wave in region III, each containing all modes. Similarly, a forward and reverse traveling wave will be excited in region II. Thus we express the line currents and voltages in terms of reflection and tranmission coefficients, as follows:

$$I_\alpha^{(k)}(z) = \begin{cases} I_k[W_\alpha^{(k)}\exp(-\Gamma_k z) - \sum W_\alpha^{(j)} R_{jk}\exp(\Gamma_j z)] \ , \ z < 0 \\ I_k \sum \hat{W}_\alpha^{(j)}[A_{jk}\exp(-\hat{\Gamma}_j z) - B_{jk}\exp(\hat{\Gamma}_j (z-\ell))], \ 0 < z < \ell \\ I_k \sum W_\alpha^{(j)} T_{jk}\exp(-\Gamma_j(z-\ell)) \ , \ z > \ell \end{cases} \tag{3}$$

$$V_\alpha^{(k)}(z) = \begin{cases} I_k[W_\alpha^{(k)}K_\alpha^{(k)}\exp(-\Gamma_k z) + \sum W_\alpha^{(j)}K_\alpha^{(j)} R_{jk}\exp(\Gamma_j z)], \ z < 0 \\ I_k \sum \hat{W}_\alpha^{(j)}\hat{K}_\alpha^{(j)}[A_{jk}\exp(-\hat{\Gamma}_j z) + B_{jk}\exp(\hat{\Gamma}_j (z-\ell))], \ 0 < z < \ell \\ I_k \sum W_\alpha^{(j)}K_\alpha^{(j)} T_{jk}\exp(-\Gamma_j(z-\ell)), \ z > \ell \end{cases} \tag{4}$$

where α represents L, s, or i denoting the various conductors. The $\sum$ denotes the sum over j of all modes and I_k is the incident sheath current in the kth mode at $z = 0$. R_{jk} and T_{jk} are the coefficients for the reflection and transmission into the jth mode due to an incident kth mode. Similarly, A_{jk} and B_{jk} are the coefficients for the forward and reverse travelling waves in region II. $W_\alpha^{(j)}$ represents 1, P_j, or Q_j for α = s, i, or L, respectively.

In order to solve for these reflection and transmission coefficients, it is necessary to enforce the continuity of each current and voltage given by (3) and (4) at $z = 0$ and $z = \ell$. This results in a linear system of 12 equations and 12 unknowns (or 8 equations and 8 unknowns if $\sigma d = 0$) for each incident mode. The coefficients are then obtained by inverting the linear system using standard matrix techniques.

Actually, the structure we have described is a mode converter and its reflection and transmission coefficients are parameters which characterize its properties as a converter. For communications purposes, it is desirable for such converters to behave as forward directional couplers, and to couple energy from one forward-travelling mode to another at a rate determined by the particular application [7].

Before presenting some numerical examples, it is worthwhile to consider one important limiting case; that is, where $L_T = 0$. Now the cable braid is perfectly conducting and the interior and exterior problems are decoupled. For the interior region, we have one coaxial cable mode (the bifilar mode). For this mode, $P = -1$ and $Q = 0$. Thus, from (1), we find that for the bifilar mode $QK_L = K_s = 0$. Similarly, for the exterior region, we have two modes (the monofilar and jacket modes). For each of these modes, $P = 0$ and thus $PK_i = K_s$.

NUMERICAL RESULTS

In what follows, we adopt the convention that modes 1, 2, and 3 denote the monofilar, bifilar, and jacket modes, respectively. It was previously found that for typical values of σd, the attenuation rate of mode 3 (jacket mode), was orders of magnitude larger than that of modes 1 or 2[4]. Thus, for all practical purposes, the propagation effects of this mode are negligible. However, it is conceivable that significant amounts of energy could be converted to mode 3 from modes 1 and 2. Although the propagation effects of this energy can be ignored, the corresponding loss of energy in the other two modes may be significant. Consequently, in what follows, while the coefficients R_{ij} and T_{ij} for $i,j = 1,2$ are of primary interest, the possible influence of the jacket mode is allowed for.

In the examples given here, we have chosen the following typical tunnel parameters: $f = \omega/2\pi = 36$ MHz, $a_o = 2.0$ m, $\rho_o = 1.8$ m, $\phi_o = 45°$, $\varepsilon_e = 10\ \varepsilon_o$, $\sigma_e = 10^{-2}$ mhos/m, $a = 1.5$ mm, $b = 1.0$ cm, $c = 1.15$ cm, $\varepsilon = 2.5\ \varepsilon_o$, $\varepsilon_c = 3.0\ \varepsilon_o$, $\sigma_w = 5.7 \times 10^7$ mhos/m, and $\psi = 45°$.

We define the transmission-reflection ratio Δ_{ij} by

$$\Delta_{ij} = 20 \log_{10} |T_{ij}/R_{ij}|$$

Since we would like the mode converter to behave as a forward directional coupler, it is desirable to maximize Δ_{ij}. Fig. 5 shows the values of this parameter as a function of the segment length (ℓ) of the converter. In this case, $L_T = 1.0$ nH/m and $\hat{L}_T = 100$ nH/m. Values are given for both $\sigma d = 0.1$ mhos and $\sigma d = 0$.

We find that there are converter lengths which maximize Δ_{ij} (and thus minimizing R_{ij}) for various i and j. It is also noted in Fig. 5 that the presence of lossy film does not have much effect on the value of these optimum lengths; however, it does appreciably affect the values at these peaks. We see that the lossy film reduces cross reflections (R_{12} and R_{21}) but increases self reflections (R_{11} and R_{22}). Also note that Δ_{12} is equal to Δ_{21}. This can easily be explained using reciprocity arguments although, in general, for our mode normalizations, $T_{12} \neq T_{21}$ and $R_{12} \neq R_{21}$.

Actually, in practical applications, it is particularly important to maximize Δ_{12}. Thus, for example, suppose that we have mode converters equally spaced along the length of the tunnel. Then, since the attenuation of mode 1 is higher than that of mode 2, a signal converted to mode 1 from mode 2 (via T_{12}) will attenuate faster than that transmitted to mode 2. Thus, at some point down the tunnel this signal in mode 1 will become smaller than the signal reflected by the next converter from mode 2 into mode 1 (via R_{12}). Since the reflected signal is out of phase with the transmitted signal, it is essentially interference. Thus, Δ_{12} determines the distance between consecutive converters over which the reflected wave produces negligible interference.

Two other useful parameters are defined by

$$I_j = -20 \log_{10}|T_{jj}|$$

and

$$C = -20 \log_{10}|T_{21}T_{12}|$$

where I_j is the insertion loss in the jth mode in dB and C is the coupling loss in dB. C is important because, in a practical problem, the initial excitation will be predominately in mode 1 (monofilar). At a sufficiently large distance from this excitation, the predominant signal will be one which was converted to mode 2 (bifilar) by the first encountered converter, and then converted back to mode 1 at one of the following converters. Thus C represents the loss associated with this double conversion. Indeed this doubly converted signal will be dominant for distances (in meters) such that $C + \alpha_2 z < \alpha_1 z$ where $\alpha_j = 8.686 \text{ Real } (\Gamma_j)$ is the attenuation rate of the jth mode in dB/m.

Fig. 6 shows the insertion loss for the same parameters as in Fig. 5. Also, plotted is the curve $\alpha_j \ell$ which represents the attenuation of the same length of cable in the absence of the converter. Note that the insertion loss is to a large extent determined by this quantity, particularly for mode 1.

Figure 7 shows the coupling loss for the same parameters. An additional curve is shown for different values of sheath transfer inductance (L_T = 2nH/m, $\hat{L}_T$ = 40 nH/m). As one would expect, the coupling loss increases as the length of the converter is decreased, and also as the contrast between the two cable types decreased. Note that there is in each case a length ℓ which minimizes this loss. Delogne [6] has computed such a length using approximated coupled mode theory. His optimum length for maximum transmission coupling is given by $\ell_o = \pi/(\beta_2-\beta_1)$ where $\beta_j = \mathrm{im}(\Gamma_j)$ is the phase constant of the jth mode. This value is shown in Fig. 7 for L_T = 1 nH/m, $\hat{L}_T$ = 100 nH/m. It can be seen that this estimate by Delogne is relatively accurate for the model we have adopted. Note, however, that the simultaneous minimization of C and maximization of Δ_{12} is not, in general, possible (see Fig. 5).

Finally, we consider the case where identical converters are evenly spaced along the tunnel with period D. If the reflections from one of the converters are small, they can be neglected, and we can assume that the converters will not couple with one another. In a practical application, this system would be driven by a hand-held transmitter in the tunnel region. We choose a relatively analogous excitation to be a voltage source between the tunnel floor and the cable's braided sheath. This source is arbitrarily located between two converters at $z = -z_o$ (see Fig. 8). Similarly, signals will be received by a hand-held receiver in the tunnel region, and thus will be proportional to the field strength in this region. Since the fields outside the cable are proportional to the total axial current in the cable ($I_T = I_L + I_s + I_i$), we will compute $I_T(z)$ for each mode for this periodically loaded line.

Since the phase velocity of the two modes are different, signals travelling down the tunnel by different paths will, in general, have different phases.

The strongest "signal" can be properly thought of as the desired signal and the remaining weaker "signals" resulting from the different paths must be thought of as interference or noise. Similarly, reflected signals must also be thought of as interference.

The voltage source at $z = -z_o$ will excite both the monofilar and bifilar modes. If we ignore the reflections from mode converters and neglect the coupling to the jacket mode, the following expression can be obtained for the relative strengths of the two incident modes at $z = 0$:

$$\frac{I_T^{(2)}(0)}{I_T^{(1)}(0)} = -\left(\frac{K_s^{(1)} - P_1K_i^{(1)}}{K_s^{(2)} - P_2K_i^{(2)}}\right)\left(\frac{1 + P_2 + Q_2}{1 + P_1 + Q_1}\right) \exp[(\Gamma_1-\Gamma_2)z_o] \tag{5}$$

Note that, as expected, when $L_T = 0$, the bifilar mode is not excited and this ratio is zero.

Figure 9 shows the total current in the monofilar and bifilar modes as a function of z for periodically spaced converters, each of which is described by the parameters in Fig. 5 and with $\sigma d = 0.1$ mhos. The converter length ℓ was chosen to be 2.94 meters, corresponding to a maximum of Δ_{12} (i.e. see Fig. 5). The separation between converters (D) was chosen to be 39.73 meters. This choice was made in order that the ratio of signal to interference in the monofilar mode be the same immediately preceding and after the third converter (i.e. at $z = 2D$ and $z = 2D+\ell$).

Note that because of the difference in the attenuation rates of the two modes, (5) is dependent upon z_o and assumes its largest value for $z_o = d$. Even in this instance, for the data given in Fig. 9, its value is -92 dB, which is seen to be negligible compared to the current in the bifilar mode resulting from conversion from mode 1 to mode 2 at the first mode converter which is at -77 dB. Thus, although not shown in Figure 9, the interference in mode 2 is at least 15 dB below the dominant signal.

We see from Fig. 9 that the dominant signal in the monofilar mode is the original incident signal preceding the first 3 converters, but after the third converter, the dominant signal is that converted from the bifilar mode by the previous converter. Also, note that the monofilar interference immediately preceding each converter is a reflected wave from the bifilar mode, illustrating the importance of maximizing Δ_{12}. Finally, we see that, because fields in the tunnel are proportional to the total current, the received signal will essentially be the dominant monofilar mode signal and the bifilar mode currents must be thought of as additional interference.

We may generalize these results to similar systems in which the initial excitation is predominantly in the monofilar mode. We find that for large z, $I_T^{(1)}$ is bounded by (in dB)

$$\left|I_T^{(1)}\right|^{max} = -C - \alpha_2 z \tag{6}$$

and

$$\left|I_T^{(2)}\right|^{min} = -C - \alpha_1 \ell - \alpha_2 (z-\ell)$$

for $z > C/(\alpha_1-\alpha_2)$. For small z, $I_T^{(1)}$ is given approximately (in dB) by

$$\left|I_T^{(1)}\right| \simeq -\alpha_1 z \tag{7}$$

for $z < C/(\alpha_1-\alpha_2)$.

CONCLUDING REMARKS

Using an idealized model, we have confirmed the general principle of leaky feeder systems as promulgated by Delogne [1] and his colleagues. The concept that such systems can be designed on the basis of controlled rather than inadvertent mode conversion would seem to be very important [7]. But, of course, there will always be a certain amount of spurious or non-intentional mode conversion due to irregularities in the tunnel wall such as ribs and other changes in the cross-section. Actually, within the frame work of our theory, lateral changes in the effective surface impedance of the tunnel wall can be taken into account. This certainly remains as an important task for further work.

ACKNOWLEDGEMENT

We are grateful for the useful and cogent comments from Prof. Paul Delogne.

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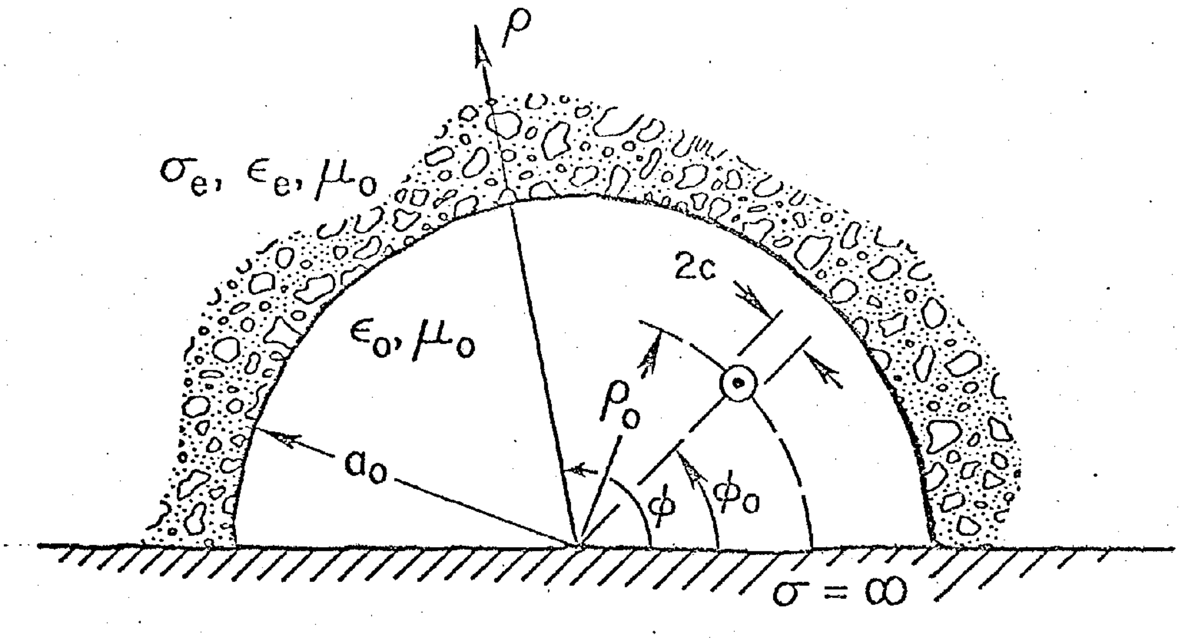

FIG. 1 Cross-section of semi-circular tunnel containing coaxial cable (not to scale).

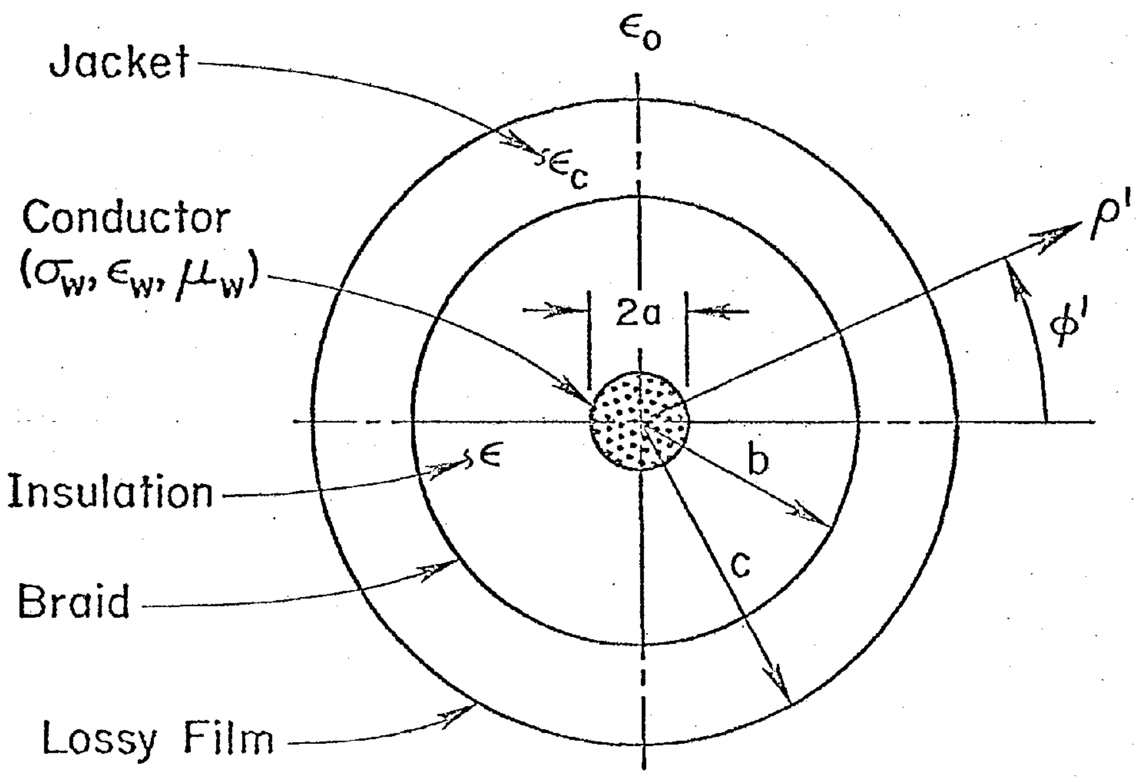

FIG. 2 The braided coaxial cable.

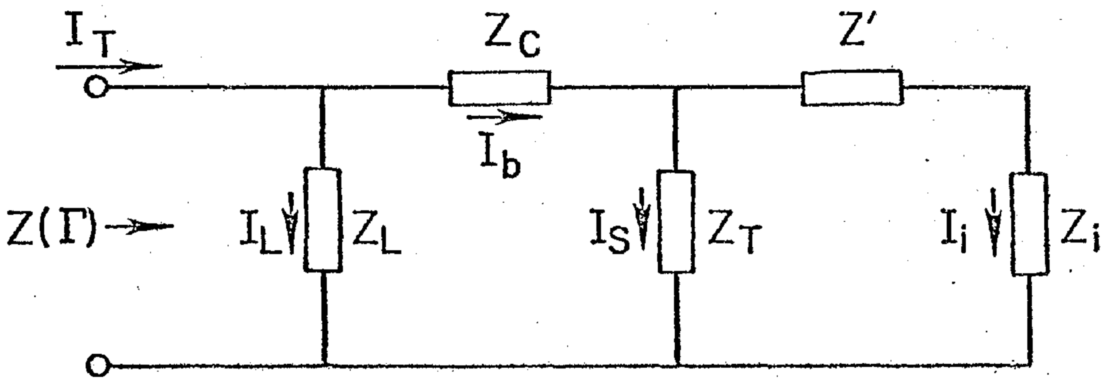

FIG. 3 The equivalent network which yields the effective series impedance of the cable in the quasi-static approximation.

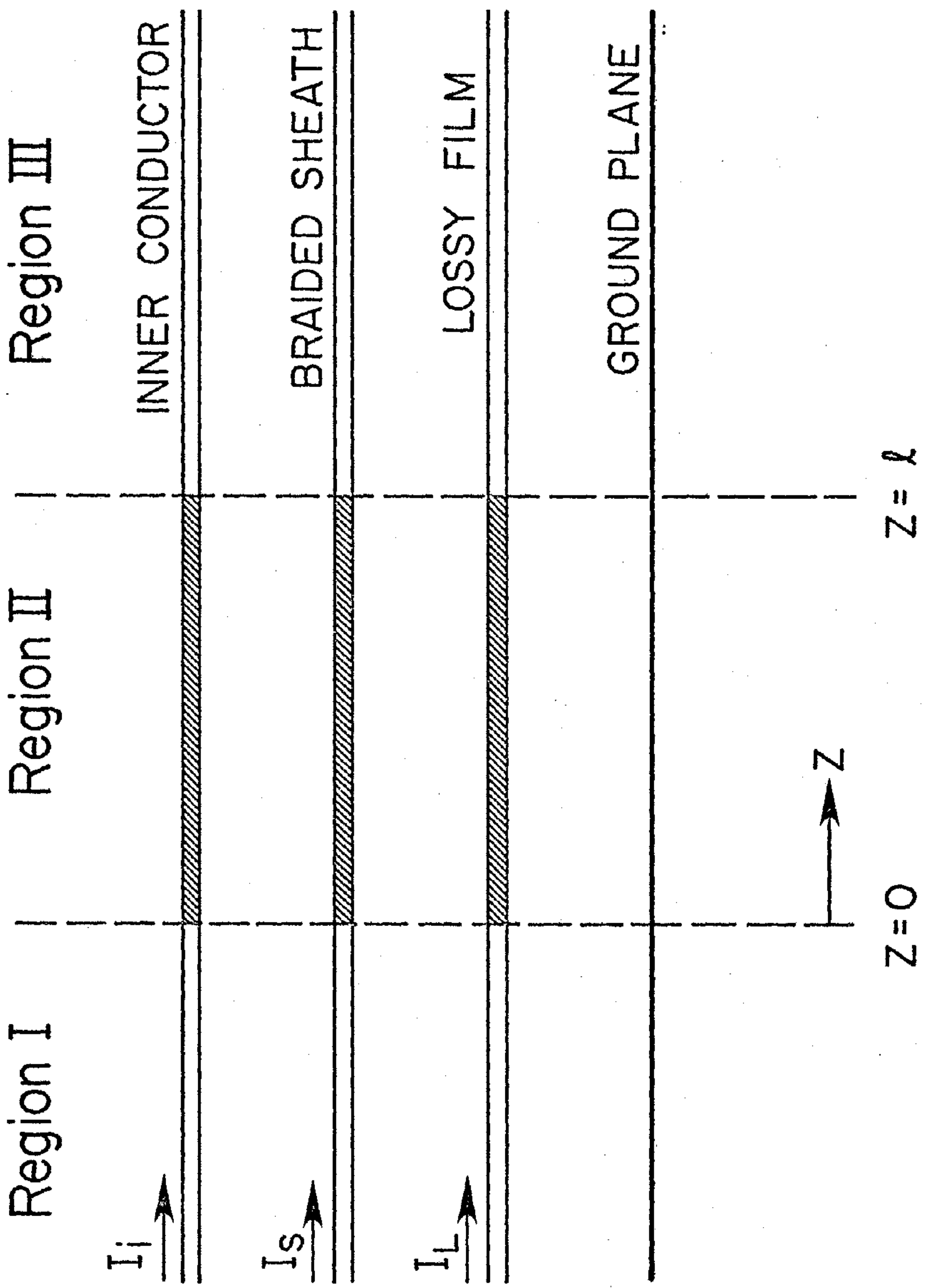

FIG. 4 Schematic diagram of multiconductor transmission line model of mode converter of length ℓ.

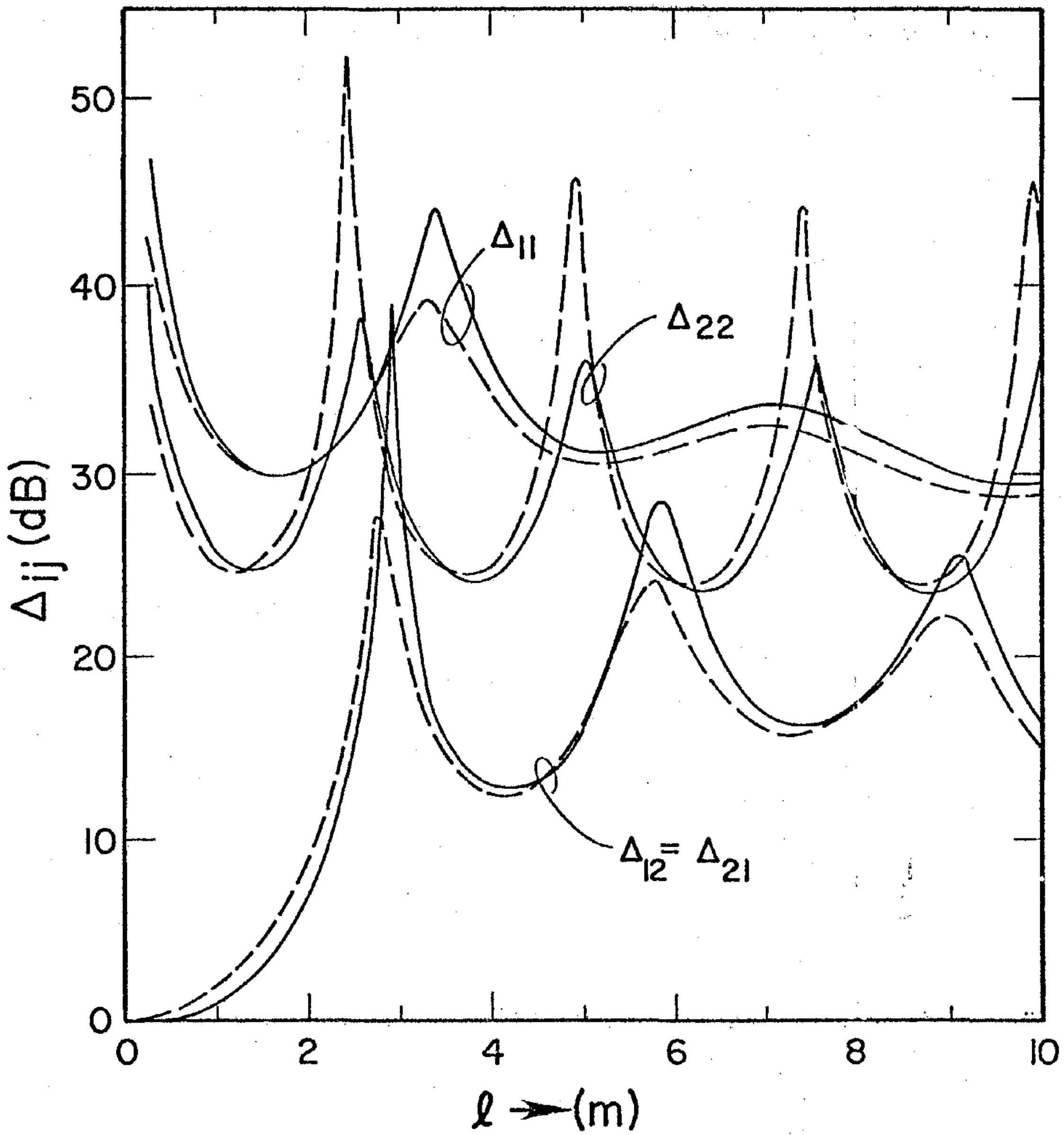

FIG. 5 Transmission-reflection ratio as a function of converter length.

σd = 0.1 mhos: ———————; σd = 0: - - - - - - - -.

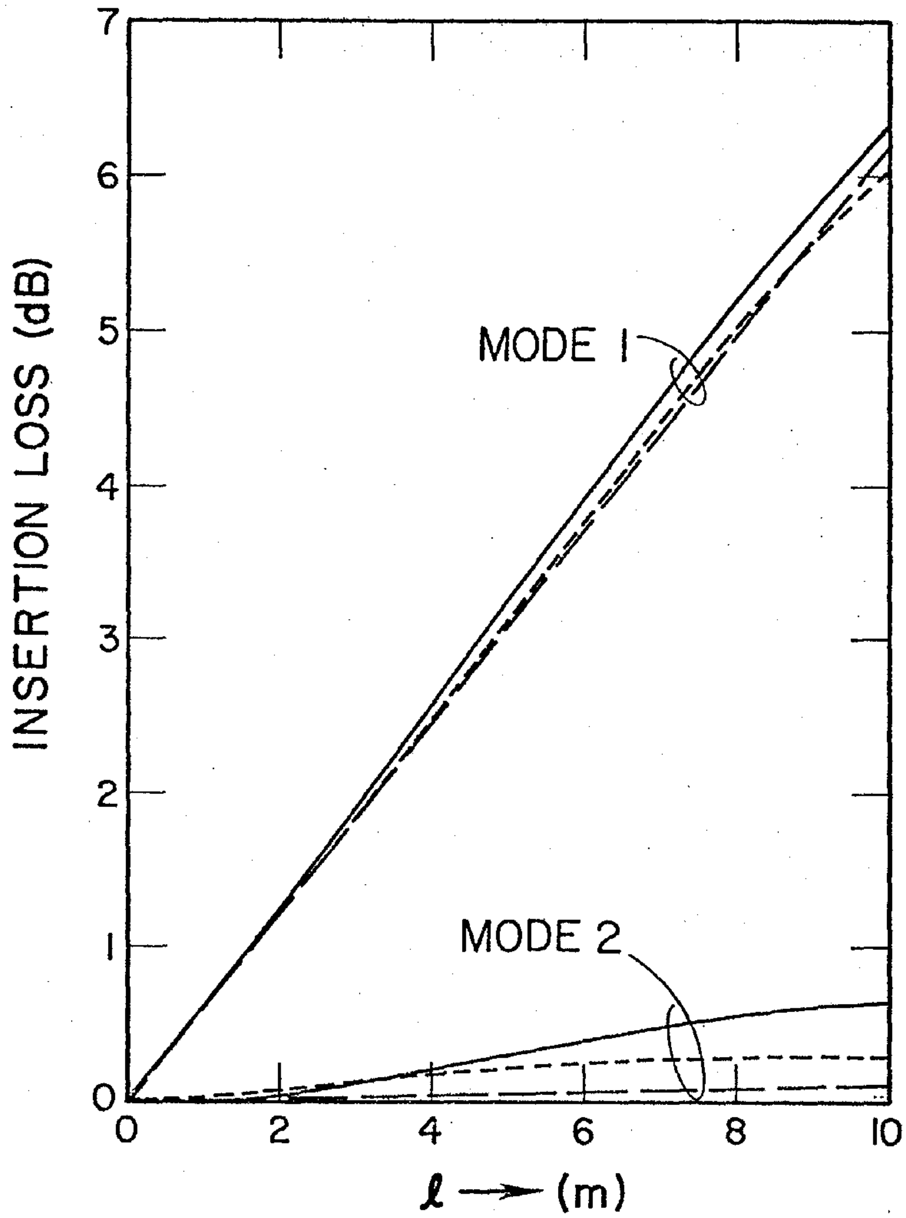

FIG. 6 Insertion loss as a function of converter length.

σd = 0.1 mhos: ————; σd = 0: – – – – –; $\alpha_j \ell$: ——— ———.

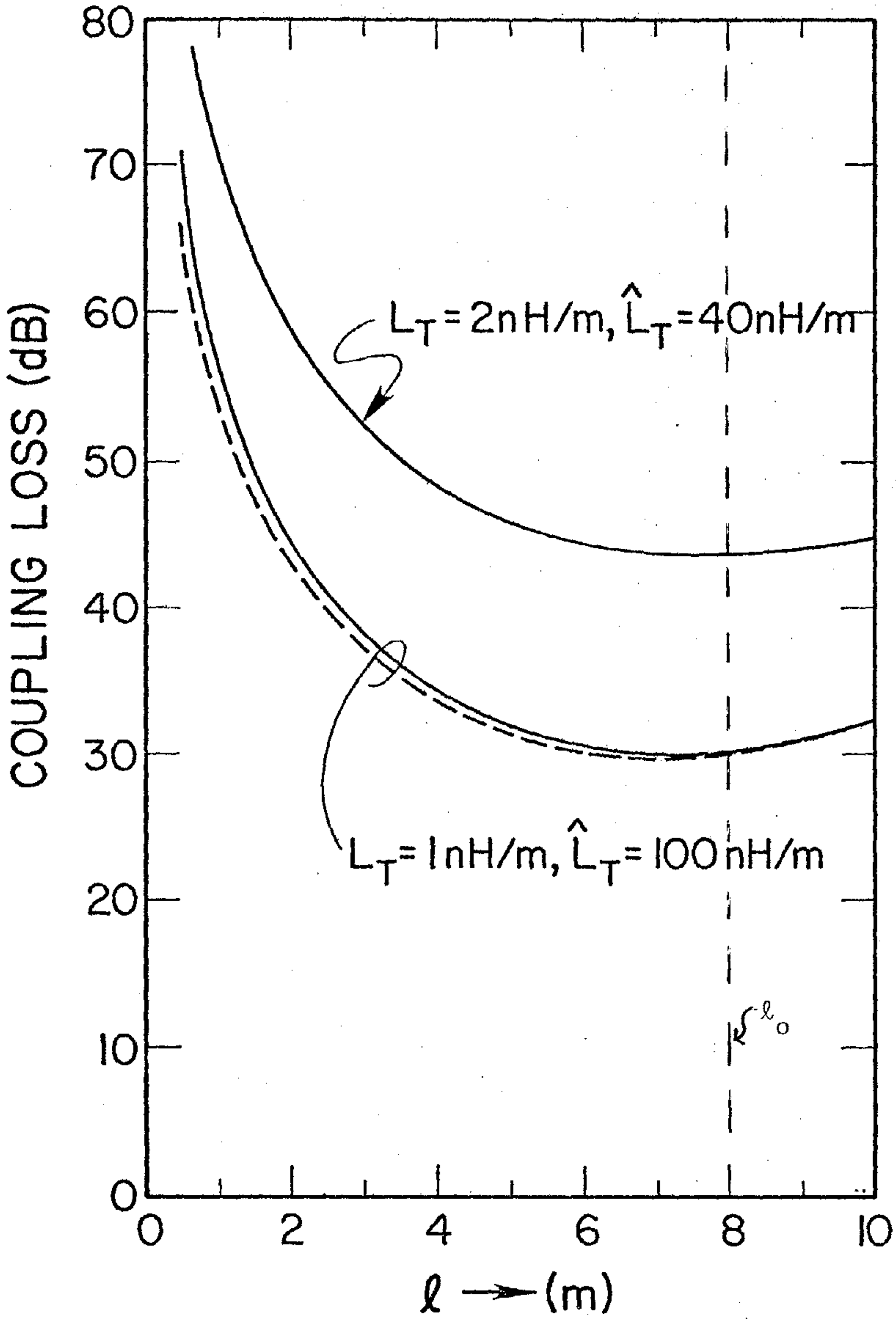

FIG. 7 Coupling loss as a function of converter length. σd = 0.1 mhos: ————————; σd = 0: – – – – – – –.

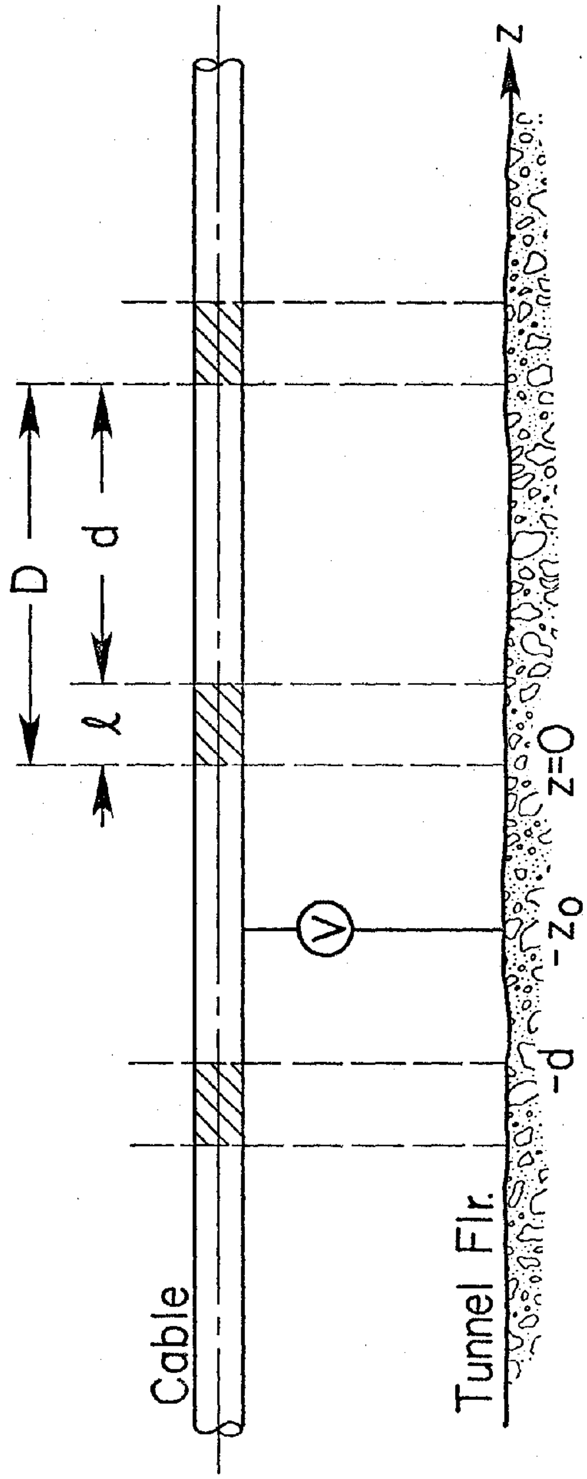

FIG. 8 Schematic diagram of uniformly spaced mode converters.

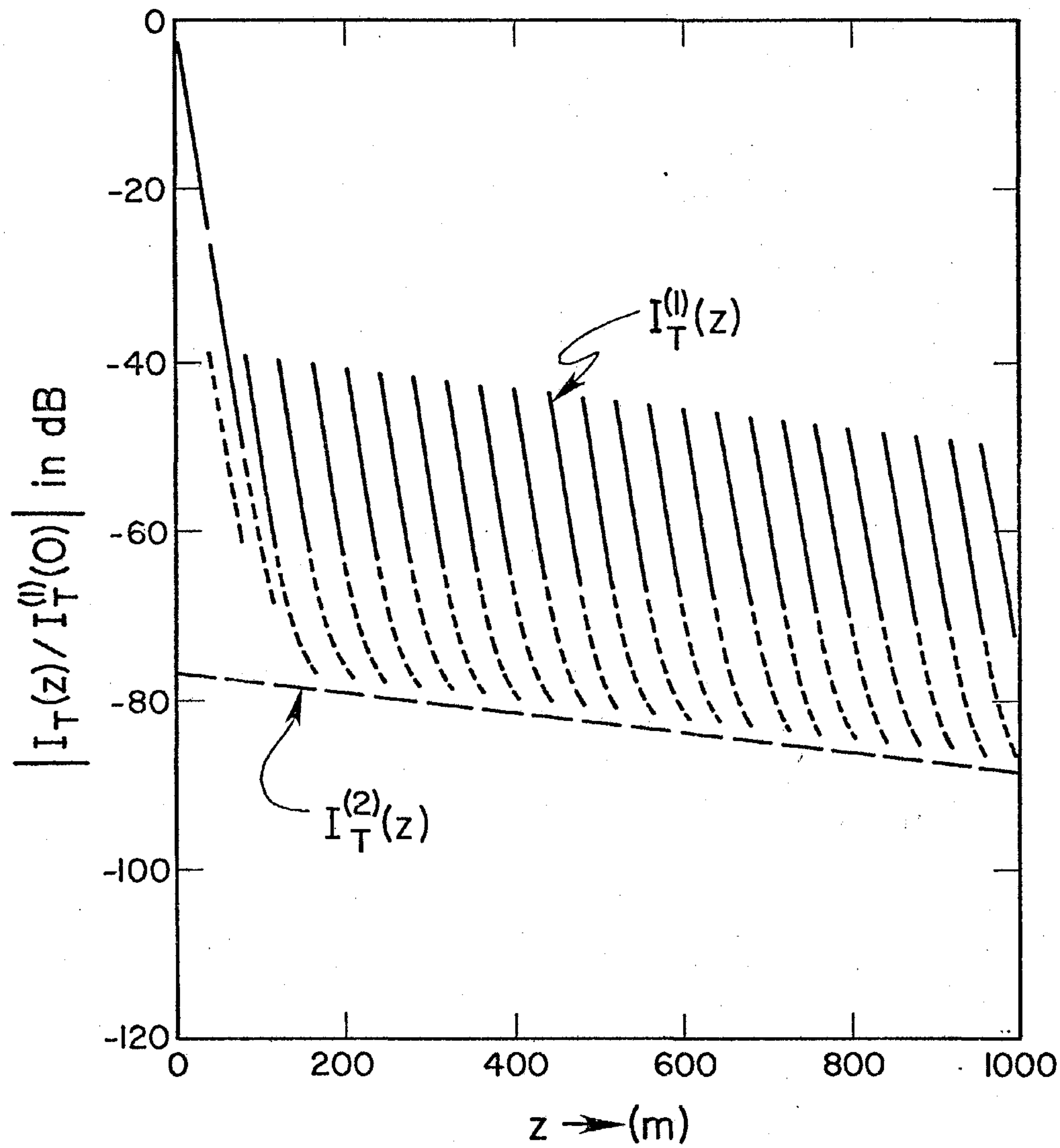

FIG. 9 Total axial current on uniformly loaded line. Dominant signal in modes 1 and 2: ————————; Interference in mode 1: - - - - - - - -.

PULSE TRANSMISSION ALONG CABLES IN CIRCULAR TUNNELS

D.A. HILL AND J.R. WAIT
ITS/OT, U.S. Dept. of Commerce
Boulder, CO 80303

The leaky-feeder technique for mine communications has recently been investigated in several theoretical and experimental studies [e.g. Special Issue of Radio Science, Vol. 11, No. 4, April 1976]. Most theoretical studies have concentrated on the attenuation rates of the dominant modes and the coupling loss from the cable to a nearby antenna. In addition to these signal amplitude effects, the bandwidth of leaky-feeder systems is of interest for cases where transmission of wideband signals is desired. In this talk, we analyze an idealized leaky-feeder channel which consists of a uniform leaky coaxial cable located at an arbitrary position within a uniform circular tunnel.

The frequency domain transfer function for the channel is first expanded about the carrier frequency in constant, linear, and quadratic terms in both amplitude and phase. Since a distortionless channel is one of constant amplitude and linear phase, the linear and quadratic amplitude terms and the quadratic phase term will cause pulse dispersion and limit the channel bandwidth. In earlier applications of this technique [1] to ionospheric pulses, it was not necessary to consider the amplitude distortion as a first order effect. The transmission of a sinusoidal signal with a Gaussian envelope is considered in detail. The transmitted pulse envelope is shown to be stretched in time and the sinusoidal carrier frequency is shifted. The pulse stretching is an inverse function of the channel bandwidth.

Specific numerical results are calculated for attenuation rates, phase and group velocities, and channel bandwidths for frequencies from 1 MHz to 100 MHz. The bifilar mode is of primary interest, but some results are also given for the monofilar mode. The phase and group velocities of the bifilar mode are remarkably frequency-independent and are primarily a function of the cable parameters. This is because the interaction of the leakage fields with the lossy tunnel wall is usually fairly weak. Consequently, the limitations of channel bandwidth due to both phase and amplitude dispersion are not severe.

A more complete transfer function is the mutual impedance between a pair of short dipole antennas located within the tunnel since this includes the antenna-cable coupling. The resultant channel bandwidth is decreased somewhat due to the introduction of the coupling terms, but it is still typically greater than 10% of the carrier frequency. A more complicated model for the surface transfer impedance of the cable braid which includes spatial dispersion has also been examined.

The conclusion of the analysis is that the bandwidth limitations for a uniform cable in a uniform tunnel are not severe. More complex channels which contain mode converters (or inadvertent mode conversion) are more likely to be frequency-sensitive and band-limited. It may also be necessary to consider third order phase terms for a more precise characterization of the transfer function [2]. Future work will consider these more complex channels.

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Some aspects of leaky feeder radio communication being studied by the UK National Coal Board

DAVID J.R. MARTIN

I. INTRODUCTION

The present UK programme of research and development in underground radio communication began in 1966 as a quest for a possible means of providing coal miners with personal two-way radio communication to keep them in contact throughout the mine. That aim is still a prime motivation of the continuing work and is still unfulfilled. However, through the extensive development of the leaky feeder principle an engineered system which has now been in production for several years provides a more modest cover - say, up to 10 km in end-to-end range with a limited facility for taking spurs or branches. These present restrictions, it may be noted at once, arise mainly from questions of intrinsic safety of the line-fed power supplies to repeaters rather than from any inherent limitation of the basic propagation method.

Although not meeting the requirements for true pit-wide communication, the present system (known as the 986 system) is proving extremely useful in providing more specialized two-way communication in association with underground transport systems - conveyor belts, locomotive systems and cable-hauled manriders. For these, the 986 system uses standard commercial two-way personal radio sets operating in the British VHF Low Band (68-88 MHz) and certified intrinsically safe.

More important than this, the 986 system is also being used extensively as a 'bearer' for continuous safety-check signalling from a cable-hauled manrider to its haulage engine, perhaps 3 km or so away, using frequencies in the same band but distinct from the two-way speech channel. Although the range required here is more modest than for pit-wide speech, there is a paramount need for absolute continuity of communication, free of dropouts caused by standing waves on the line or multipath effects in the tunnel. Furthermore, the number of such additional channels required in a single mine might be substantial with possibly overlapping range requirements, since such haulages as are being considered often operate in tandem and may each consist of two trains working in opposition on the same rope. Overlying this is still the basic need for at least one - possibly more - pit-wide channel for speech.

The emerging overall pattern is thus very complicated in terms of propagation requirements. With line-fed power restricted by intrinsic-safety considerations, and the risk of intermodulation interference between channels in repeaters operating as efficiently as possible (and thus possibly non-linearly), it becomes important to prevent propagation of signals in the leaky feeder highway beyond their required range. Broadly the proposed plan is to employ different frequency bands according to range requirements and by means of bandpass or bandstop filters to sectionalize the entire system for short-range channels while allowing the pit-wide channels unrestricted propagation. The use of in-line frequency changers to allow selected channels to pass a section barrier is also being considered.

II. SYSTEM CONFIGURATION

The present limitation on total system length is primarily a question of power-feed management and compliance with intrinsic-safety criteria, and it is not appropriate to go into the proposed solution here. More relevant is the choice of RF system configuration to provide the required flexibility of layout with any necessary degree of branching or networking. All such arrangements being considered are based on the present concept of using small in-line amplifiers - 'repeaters' in the line-telephony sense of the word - to make up for losses at frequent intervals, and all assume that it must be possible for the system to transmit and receive at the same time.

The most straightforward arrangement is to use repeaters which are themselves two-way devices; in their conventional form these would each comprise two back-to-back amplifiers with filters (and perhaps also hybrid transformers) to segregate the signal paths and ensure adequate stability. This approach leads to a fairly costly design of repeater, and the need for multiple banding would probably rule it out for the resulting complexity; nevertheless, a suitable design for single-band operation has recently been completed for evaluation.

The 'daisy chain' solution (Fig. 1) as used in the present production version of the 986 system has been described several times previously in the literature. It employs an extremely simple one-way repeater to provide two-way communication but requires a parallel telephone line to link the extremities of the system. It is also economically unattractive if many spurs are to be taken, since every extremity has to be terminated in a fairly expensive transmitter or receiver.

The experimental BDR (bi-directional routing) system (Fig. 2) has also been described before. It uses two leaky cables, which may be combined in a single construction. The repeaters are again simple one-way devices, but instead of being inserted in series with the lines they are bridged between them in alternate directions so that signals in each line are continually being amplified and passed into the other. Simple mid-span filters (which alternatively may be combined into the repeaters) ensure stability by breaking the feedback loops, and achieve progressive routing of the signals through the system in the appropriate directions.

A crucial advantage of this system lies in the 'tailback' configuration by which half the amplified signal energy from each repeater is fed back against the main signal flow as far as the preceding filter, with a corresponding effect for signals induced in the line from mobile transmissions. The minimum signal levels thus occur at mid-span rather than at the repeaters, with consequent halving of the total variation attributable to line loss between repeaters and a corresponding increase in minimum levels. This system is at present undergoing an operational trial in Maltby coal mine, England.

An alternative approach is now being studied reverts to the original daisy-chain approach of Fig. 1 but replaces the telephone line by a medium-frequency carrier link over the leaky feeder itself, the signals passing through the repeaters in the 'wrong' direction without being affected; at such frequencies proposed (below 2 MHz) in-line amplification would not normally be necessary. A version of this proposal known as 'IF return' is shown in Fig. 3, where it will be seen that the former terminal receiver is now replaced by a frequency changer, which itself may be a very simple line-powered device; signals originating at a mobile transmitter on reaching a terminal frequency-changer are converted to 455 kHz and returned to the unified base station for further processing and demodulation. As illustrated, the arrangement also solves the problem of providing spurs, since the frequency changers are inexpensive enough to be used at every extremity; the line filters indicated serve to prevent signals reaching more than one frequency changer and thus giving rise to heterodyne interference (as a further refinement, the frequency changers might be synchronized by passing a common 'local' oscillator signal through the repeater chain from base).

This system has been tried experimentally in a disused railway tunnel and has worked well in a simple unbranched run containing one repeater over the limited distance of 1 km available. It is believed that the arrangement could be extended to provide for more than one channel, using different IFs but the same frequency changer; such a trial is planned for a mine.

The converse of IF return is 'forward drive'. In this system, not illustrated here, the repeaters are turned round: the normal receiver is at base position, where there is also located the crystal-oscillator drive portion of the transmitter. Frequency modulated signals, at about 1.9 MHz, are passed through the repeater chain - again the 'wrong' way - to the terminal frequency changers which in this case multiply up to the final carrier frequency near 85 MHz.

The forward-drive system also has been tested in the tunnel. Again, it is expected that the system can be extended to multi-channel operation, with the number of channels limited - perhaps to two or three - by considerations of mutual intermodulation interference in the multipliers.

These medium-frequency link systems, like the present standard daisy-chain system in widespread use in the UK, use a new design of repeater intended for half-kilometre spacing. It is extremely simple, employing only one transistor and few other components. Current consumption is 2.5 mA from the line-fed 12-volt supply, gain is 16 dB and power output for base-station transmissions is 2.5 mW.

Most of the systems mentioned so far use a common wideband amplifier for 'go' and 'return' signals; the exception is the two-way repeater, which segregates the signals through separate amplifier paths. If more than two or three channels are to be accommodated it is very desirable that the paths should be so segregated; otherwise, the relatively high-level base-station signals will tend to cause intermodulation interference with the weaker (in the leaky feeder) mobile transmissions. If the complexity of the two-way repeater is to be avoided, and yet the signals in the two directions are to be amplified separately, a very straightforward solution is to employ a 'double daisy chain'. This calls for two parallel leaky feeders - which may be combined into a single cable - each with its chain of repeaters facing in opposite directions for the two directions of transmission. The repeaters would be co-sited in pairs, with each pair probably combined into a single enclosure. The arrangement is illustrated in Fig. 4.

A natural extension of the double-daisy-chain system is to introduce a third leaky-feeder, shared between the two directions of transmission to provide a tailback configuration similar to that inherent in the BDR system; this arrangement is illustrated in Fig. 5, where it is seen that the repeaters are no longer combined in pairs but staggered.

Double-daisy-chain systems are simple in concept and configuration, but their stability could be affected by cross-talk between the lines, especially if there are any reflections due, perhaps, to mismatches at the repeater inputs. Stability can always be ensured by including filters, but this is an undesirable complication and limits the flexibility of the arrangement.

All such basic system configurations are under appraisal as possible successors to the present system of Fig. 1. Each seems to have its particular merits and disadvantages according to the intended application, and it is likely that an overall plan will result in a hybrid arrangement, with the various grades of mine roadway served differently. In particular, some of the configurations are less amenable to the double-banding proposal.

III. DIVERSITY

A characteristic of VHF leaky-feeder communication in mine roadways is the pronounced local variation in signal level experienced, often amounting to 20 dB or so over a distance of less than one wavelength. When considered on a logarithmic scale the effects normally appear as needle-sharp dropouts below a mean level which otherwise varies only slowly with distance, following cable attenuation and repeater gains. Two-way speech communication is very tolerant of these disturbances but such interruptions in the transmission of data or coded signals from a moving vehicle will give rise to errors. This would be serious, for example, where the continued running of a haulage engine is dependent on the uninterrupted receipt of a coded safety-check signal from the train; in the extreme case where the train stopped with its antenna in the dropout it would be impossible to re-start the haulage.

The variations themselves arise from two causes: (1) discontinuities and interferences in the line structure and its mode propagation, and (2) multipath effects in the environment; it can be very difficult to distinguish between these categories in practice, since probing close to the line in an endeavour to exclude environmental effects will itself introduce a discontinuity to the line. Strictly, the two categories are probably inseparable anyway if the complete line and its environment are regarded as a unified propagating medium; but it is useful in practice to be able to make the distinction.

If the line itself is uniform, and there are no interactive discontinuities from the environment, the disturbances existing on the line will be confined to end effects (but this does make an assumption that there are not two or more modes having different velocities but identically the lowest attenuation constant - if there should be, the situation would be that of an indefinitely extended end effect). The end effects to be expected in an otherwise uniform line and environment have been variously predicted before and found to a greater or lesser extent in practice. But it is important also to realise now that the inclusion of a repeater in a leaky feeder, even though its size may be small enough not to disturb the single-wire uniformity of the line, is introducing a new source into the line and thus a new end effect. It can also in certain circumstances produce an amplified tail-end effect through 'repeater backwash', resulting in a more pronounced short standing wave before the repeater than the normal tail-end theory would predict; but this would only be expected with repeater gains far higher than are at present proposed.

Two of the system configurations described here (Figs. 2 and 5) incorporate a tailback feature, by which the line effectively doubles back on itself at each repeater as a parallel path to the main line. Assuming that a composite construction of cable is used, so that the single-wire modes of main line and tailback are indistinguishable, a very pronounced standing wave will result in each mid-span region between repeaters, where the main line and tailback signals are equal but propagating in opposite directions. This may or may not be observable away from the line amongst the general environmental multipath effects, but will be regarded for the present as a possible disadvantage.

The standard technique for combating multipath and standing-wave effects in radio communication is diversity, and some forms of it could be particularly suited to leaky feeder systems; for example, space diversity using mobile antennas spaced a half-wave or so apart is being examined as a way of achieving the required error-free transmission from cable-hauled trains. Frequency diversity in leaky feeder systems is being studied separately at the University of Surrey, with indications that quite close carrier spacing (500 kHz or so) might be adequate.

There is a particularly interesting form of frequency diversity to which tailback systems such as those of Figs. 2 and 5 would lend themselves. The technique would be to employ a carrier spacing corresponding exactly to the frequency at which the span length - the distance between repeaters - resonated at a half-wave length. For the case of the present experimental BDR system this frequency would be 260 kHz. The effect of such a particular spacing would be that in the mid-span region the two trains of standing waves would be exactly out of phase, with the nulls of one coinciding with the peaks of the other. Suitable combining of the signals at the two frequencies could thus provide a level substantially independent of the standing-wave effect resulting from the tailback connection.

This particular diversity technique is also expected to be effective against environmental multipath effects in the vicinity of the line in the same mid-span regions; any such benefit would of course be specially welcome here, where the mean level is at a minimum. The mechanism in this case is less obvious, but depends on an important basic assumption that the multipath effects resulting from two waves travelling in opposite directions in the line are substantially uncorrelated. This has yet to be demonstrated conclusively, but if true could turn to advantage what was previously regarded as a shortcoming of tailback systems.

IV. PERSONAL RADIO SETS

The 986 leaky-feeder system in current production uses standard VHF personal radio sets, but this feature is now preventing a wider acceptance of the system. Such sets are considered too large and heavy, too costly and insufficiently reliable or robust in mining conditions; the need for battery charging after every shift is also considered a serious disadvantage.

A purpose-designed radio set could avoid many of these criticisms by taking advantage of the characteristics of leaky feeder operation. First and foremost, the transmitter power - commonly at least 500 mW in a standard handportable set - can be reduced drastically, at least to 50 mW and perhaps lower; the immediate result of this would be an increase in battery endurance and a lighter battery. A reduction in receiver sensitivity and selectivity, and in particular the elimination of any need for second-channel rejection, would lead to a simpler and thus more reliable and cheaper receiver. The adoption of a lower frequency band than has hitherto been used - say, 40-50 MHz instead of 68-88 MHz - could simplify the transmitter and suit leaky feeder operation well.

Developments along these lines are proceeding in collaboration with a British firm, and prototype sets are undergoing evaluation trials. Although these first sets are hand-held, the objective is a radio equipment where the receiver, at least, is incorporated into the miner's helmet. The move to a lower frequency has aggravated the problem of combining efficiency with compactness in the transmitting aerial; a reasonable compromise appears to have been found, but the subject is seen to merit further specialized research.

V. CONCLUSION

Leaky feeder operation in the VHF band continues to offer the best promise for everyday radio communication requirements in British coal mines as at present foreseen. Most effective exploitation, however, demands a fuller understanding of the basic processes involved and in particular of the local signal variations observed in a non-uniform environment near a leaky feeder. Research into antennas for use with leaky feeders would also be very relevant.

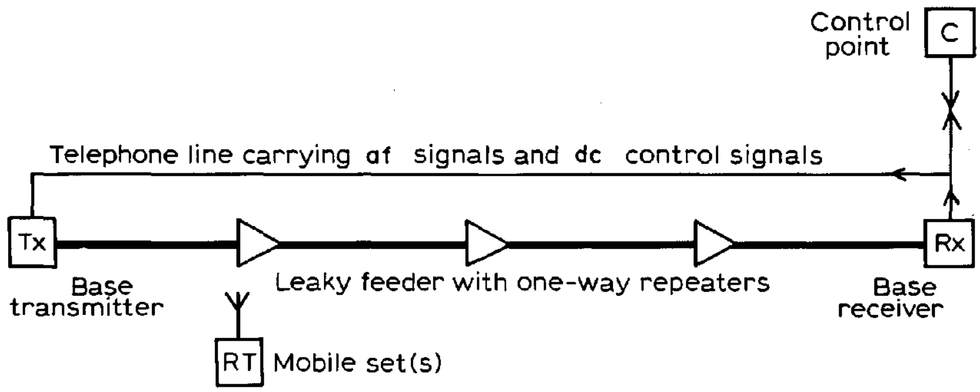

FIGURE 1. SIMPLE DAISY CHAIN SYSTEM

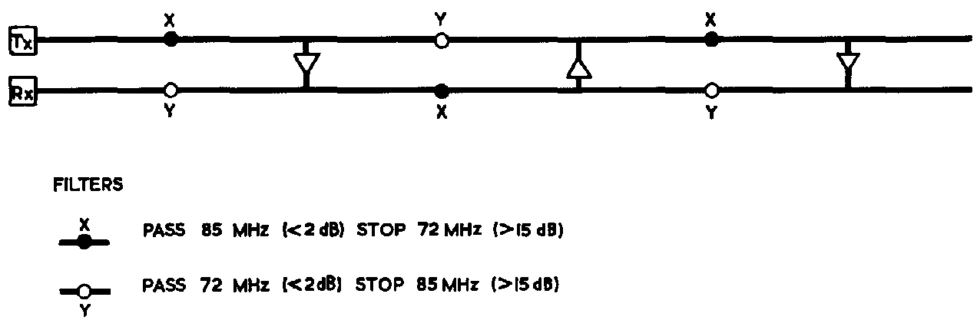

FIGURE 2. BDR SYSTEM

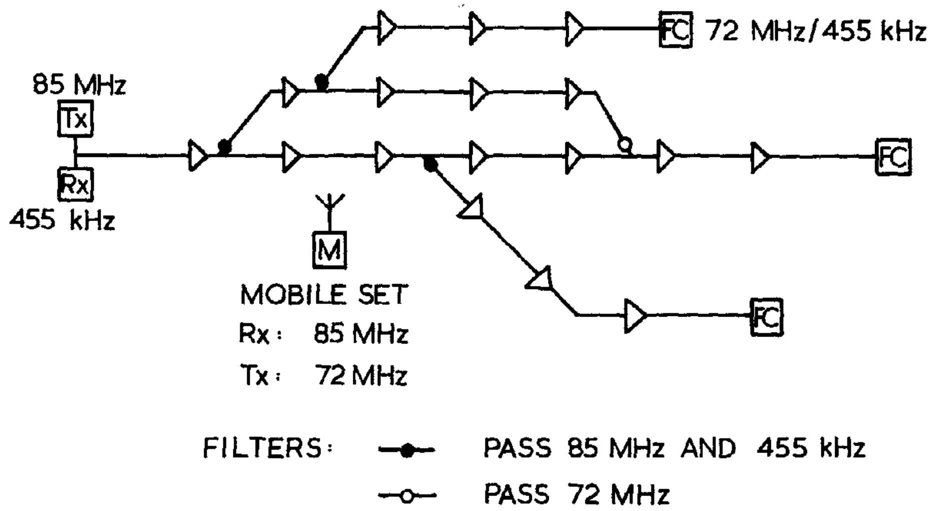

FIGURE 3. DAISY CHAIN SYSTEM WITH SPURS AND IF RETURN

FIGURE 4. DOUBLE DAISY CHAIN

FIGURE 5. DOUBLE DAISY CHAIN WITH TAILBACK

COMMENT BY N. MACKAY

An interesting method of extending the range of each length of cable used is to modify each section into a "graded" cable in which the cable coupling is increased with distance to compensate for attenuation. We have used such a system, in cooperation with Andrew antenna, and have increased the range of obstacle detection systems (using guided radar) by about 30%.

REPLY BY D.J.R. MARTIN

In our very first leaky feeder system, installed in Loncannet Mine in 1970, we did, in fact, use cable grading of necessity in one extra long section. But, for practical reasons, we have not continued the practice. It does, for instance, complicate the repair of cable damage or partial cable replacement. Grading cannot be used with the simple daisy-chain principle since the two directions of transmission would require opposite grading.

QUESTION BY P. DELOGNE

Your systems are duplex systems. Do they provide communications between mobiles?

REPLY BY D.J.R. MARTIN

Yes. The Mobiles themselves are simplex: the fixed system (base station and repeaters) has a duplex capability, and so can retransmit on the base-station frequency any signals received from a mobile. This allows direct conversation between mobiles in a simplex manner. In British terminology, this facility is known as "talk-through" and it can either be left "free-running" or kept under the control of a central operator.

COMMENT BY J.R. WAIT

One must bear in mind that leaky cables with low permittivity insulating material and high surface transfer can lead to near degeneracy between the coaxial (bifilar) and the monofilar modes. Then, inadvertent mode conversion is enhanced. On the other hand, systems based on controlled mode conversion (e.g. INEIX-Delogne type) allow for a systematic design with a more predictable performance.

COMMENT BY J.C. BEAL

There is a further subtlety in the loss in leaky coaxial cables. This is associated with the actual detailed structure. For the same coupling, i.e. field accessability from outside the cable, Kabelmetal cable has a quite wide slot and is quite susceptible to increased loss due to adjacent ducts, etc. Similarly, an RADIAX cable with a series of small holes, also has a high susceptibility due to very high local current densities around these small holes. Hence, the best leaky cable would be one with a semi-transparent outer conductor but with the lowest possible current density i.e. a uniform current density. In fact, the best practical approach appears, for this reason, to be the loose-braid coaxial cable. This may explain why some have observed the relative insensitivity of loose-braid leaky cables to adjacent tunnel wall effects.

FIELD MEASUREMENTS OF LEAKY-FEEDER SIGNALS:

Q. V. Davis, R. W. Haining, B. L. Critchley* and A. J. Motley

University of Surrey, Guildford, Surrey, England.

1. INTRODUCTION

The conventional leaky feeder communication system, as used in Britain at present utilises a loose-braided coaxial cable to propagate radio-frequency energy in mine tunnels with low loss. The propagation is dominantly via the quasi-T.E.M. coaxial mode. There also exists in the tunnel at least one other mode of higher attenuation. The consequence of this is the presence of "standing waves" in the tunnel environment, generally of length well in excess of a free space wavelength. Also, reflections from obstacles and so on create the usual fluctuations in signal level experienced in most mobile radio systems. It is these fluctuations which degrade the quality of normal leaky feeder communication systems.

Much experimental data is now available on the longitudinal amplitude variations in a tunnel but the fine details of the phenomenom have still not been fully explored. For this reason a carefully conducted series of measurements has been made to characterise, both in amplitude and phase, the three dimensional field distribution in a tunnel containing a leaky feeder.

In addition the possibility has been examined of overcoming the effects of dropouts in actual systems by use of frequency diversity. Most of the results to date are based on a system above ground, and aimed at achieving high digital data rates, using a carrier at around 40 MHz. Measurements of the variation of dropout position as a function of frequency have been made. Conclusions have been drawn as regards the optimum frequency separation in a diversity system as well as the reliability of a data link.

2. THREE DIMENSIONAL FIELD DISTRIBUTION

2.1 Experimental Arrangement

A detailed examination has been made of the variations, in both amplitude and phase, of the field distribution in a tunnel when a leaky feeder is used in the normal manner, in the v.h.f. low band at 72·5 MHz.

Fig. 1 shows the experimental arrangement and Fig. 2 the geometry of the tunnel, which is a straight, disused railway tunnel from which the tracks have been removed. It has brick walls.

* B. L. Critchley is now with Standard Telecommunications Laboratories, Harlow, Essex, U.K.

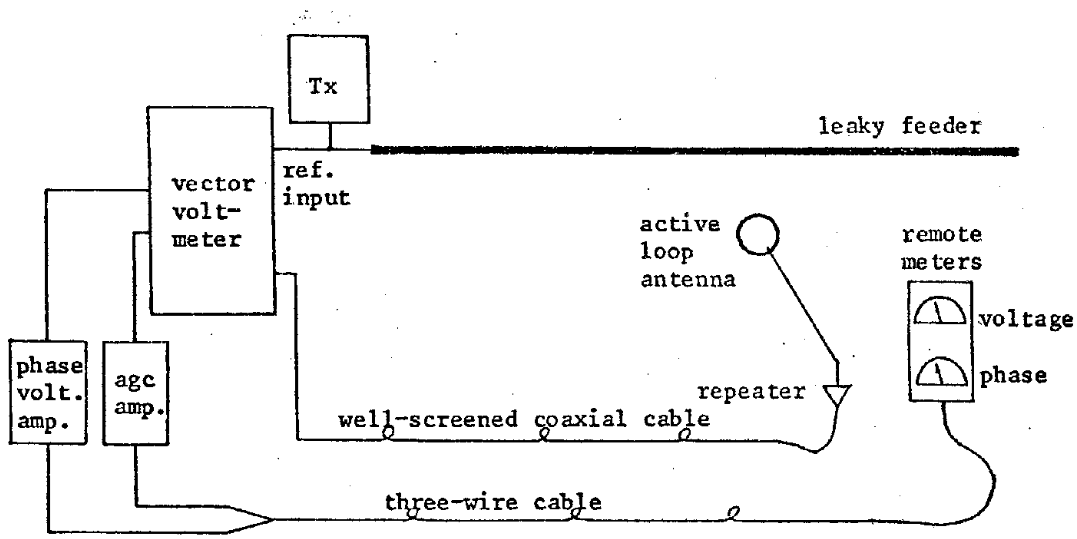

Figure 1 Arrangement for Measurement of Fields in Tunnels

The antenna was a small resonant loop feeding a transistor amplifier close to its terminals and connected to the instrumentation by well screened coaxial cable having ferrite rings along its outer surface. It was mounted on a long pole by means of which it could be placed in position without itself or its operator disturbing the field. The instrumentation cables in the tunnel were moved about and found not to disturb the measured fields.

It is hoped to carry out E-field measurements in due course, but in the first instance it was found that greater sensitivity could be gained from a small loop than from a short rod hence the decision to determine the H-field first.

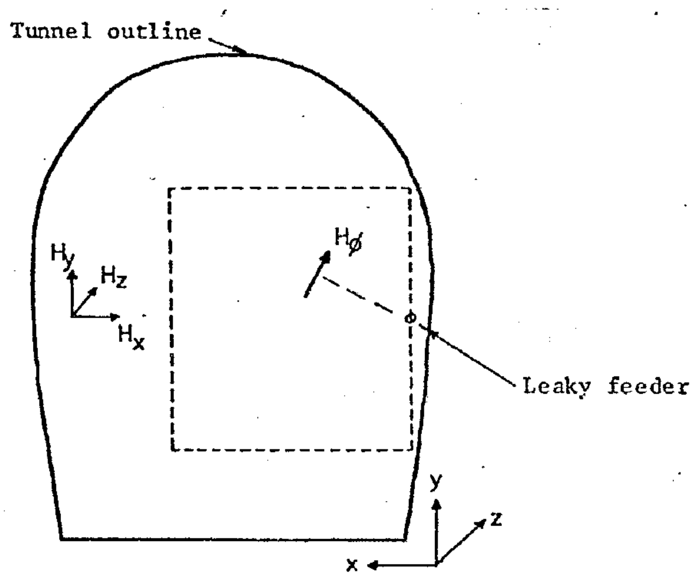

Figure 2 Tunnel geometry showing grid of measurement points

Figure 2 shows the situation in the tunnel. Measurements were made of the three orthogonal magnetic field components; horizontal, H_x, vertical, H_y and longitudinal H_z and from them the component H_ϕ was calculated (this being the component tangential to a radius from the feeder itself in the plane of the cross-section.) The procedure was to collect measurements on an accurately located grid in a vertical plane transverse to the tunnel and to repeat for successive planes over a distance of 120 cm along the tunnel. This is the greatest length which could be covered in a day, during which repeatability could be achieved.

Some of the more interesting features of the results have been selected for presentation in the following diagrams.

2.2 Field Strength Variation Across Tunnel

Figure 3 shows the variation in field strength across the tunnel at the height of the feeder for one longitudinal position.

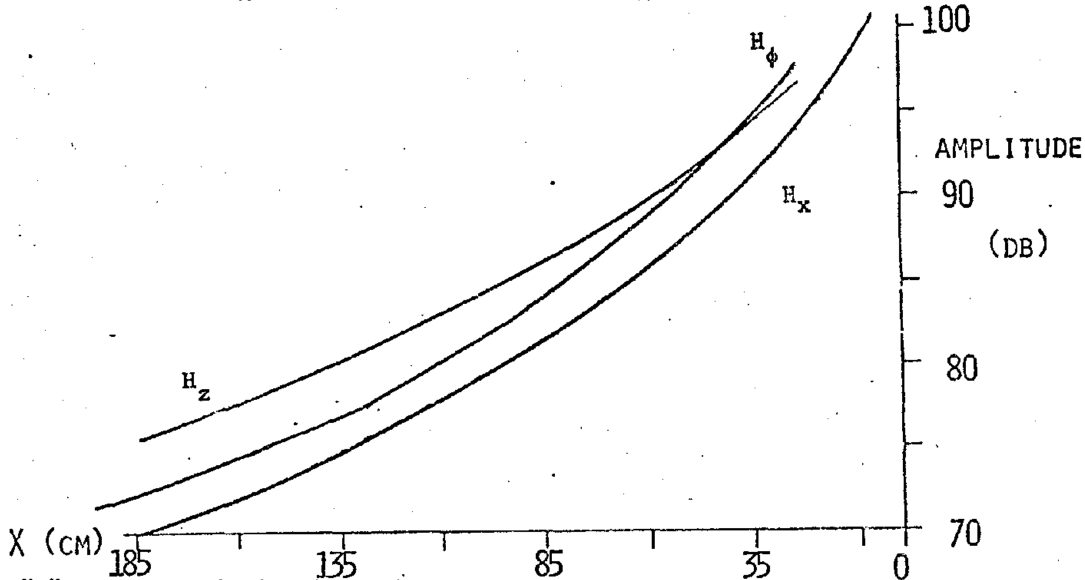

Figure 3 "H" Vector Variation in Horizontal Plane Through Leaky Feeder

Large variations exists in H_x along the tunnel so not too much emphasis need be placed on the absolute value of this component.

The variations fairly closely follow the law $H \alpha x^{-3/2}$.

2.3 Field Strength Variation Along Tunnel

The variation in field along the tunnel, figure 4, shows "standing wave" effects but these are considerably greater for H_x than the others; H_ϕ is particularly lacking in variation.

2.4 Phase Fronts

Figure 5 shows the phase fronts and contours of fixed amplitude in a horizontal plane at the same height as the feeder. This figure shows clearly the complicated nature of the pattern of the H_x component.

Note how the phase contours generally curve back slightly towards the feeder, or tunnel wall.

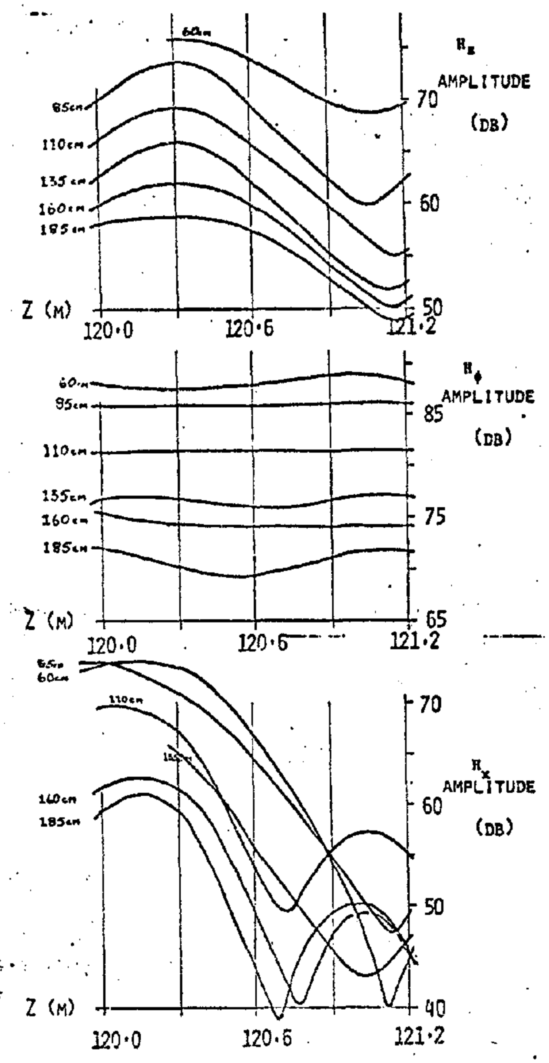

Figure 4 Field Strength Variation Along Tunnel at Specified Distances From Feeder, in Plane of Feeder

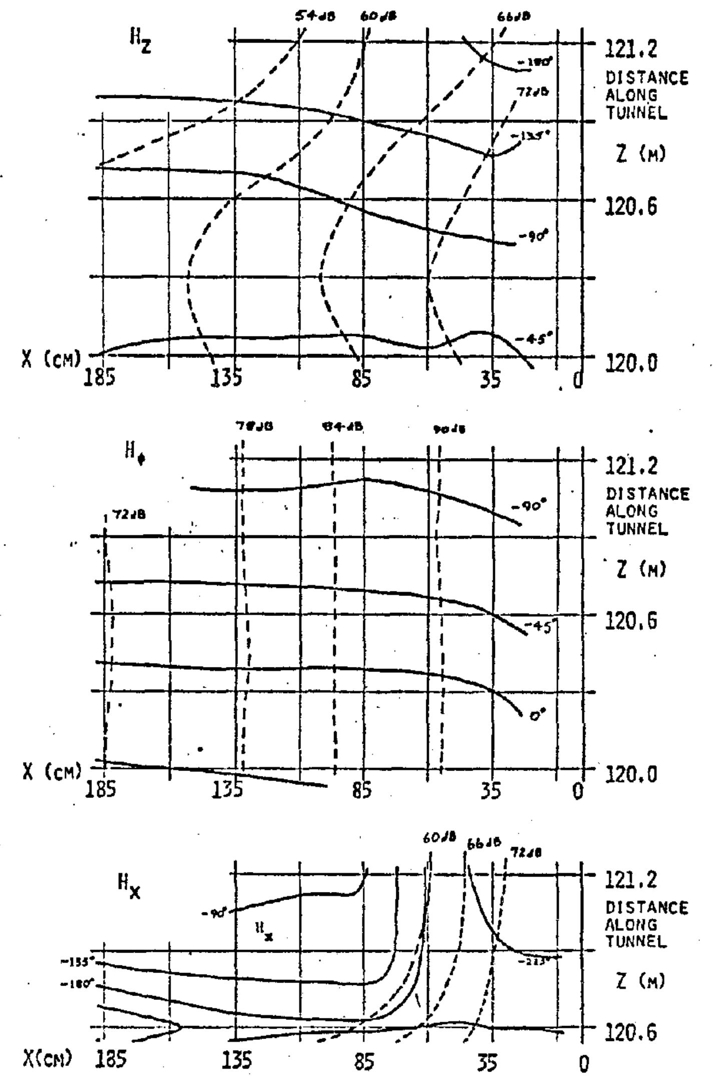

Figure 5 Phase Fronts and Amplitude Contours in the Tunnel, at the Height of the Feeder

As a final example, figure 6 shows the phase variation of H_ϕ over the height of the tunnel with its spread across the tunnel.

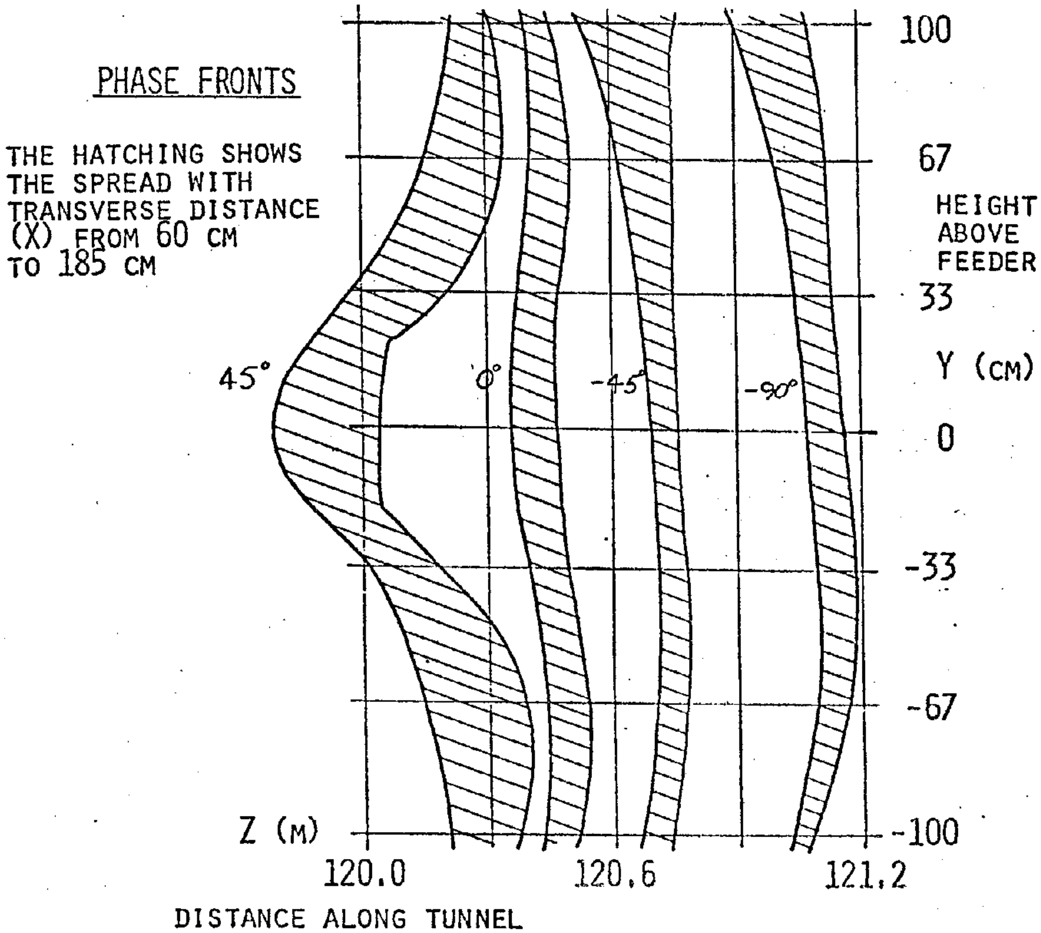

Figure 6 . Phase Fronts in a Vertical Plane

3. FIELD STRENGTH AND FREQUENCY

3.1 Signal Strength Variations at 40·5 MHz

A series of measurements were made at around 40 MHz, both in the tunnel and above ground, to determine the possibilities of high data rate transmission from a fast-moving vehicle to a fixed base.

In the tunnel the 40 MHz signal suffers little variation and the interesting results arise in the above ground case. Flunctuations can be so severe that frequency diversity techniques must be invoked for high data rate transmission.

Above ground the layout shown in figure 7 was designed to simulate the worst case conditions of a full-scale system. It was used with the experimental arrangement shown in figure 8. The feeder itself was a coaxial cable type B.I.C.C. T.3522 with a solid outer conductor having holes in it. Its coaxial mode attenuation is 25 dB/km. In a typical experiment a car with a centre mounted rod aerial was driven parallel to the line at a distance of 4 metres and radiated 250 mwatt. In some cases the feeder had a galvanised iron support wire suspended about 4 inches above it.

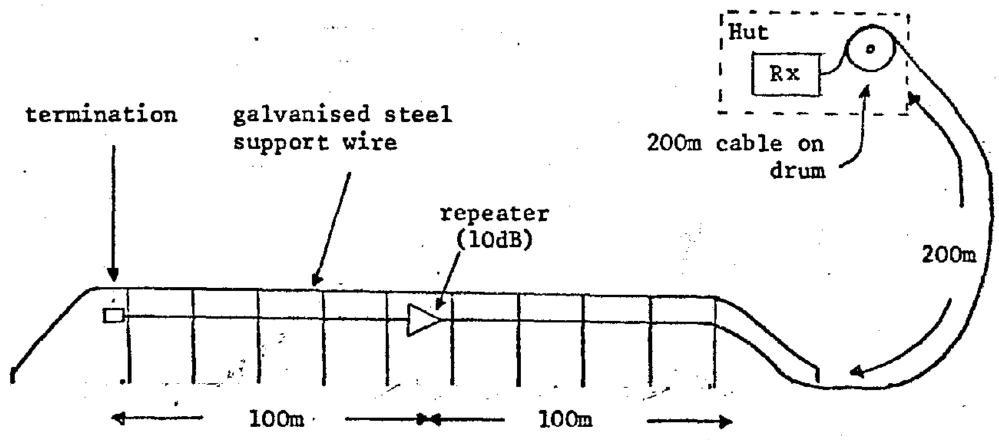

Figure 7 System Layout for Above Ground Measurements

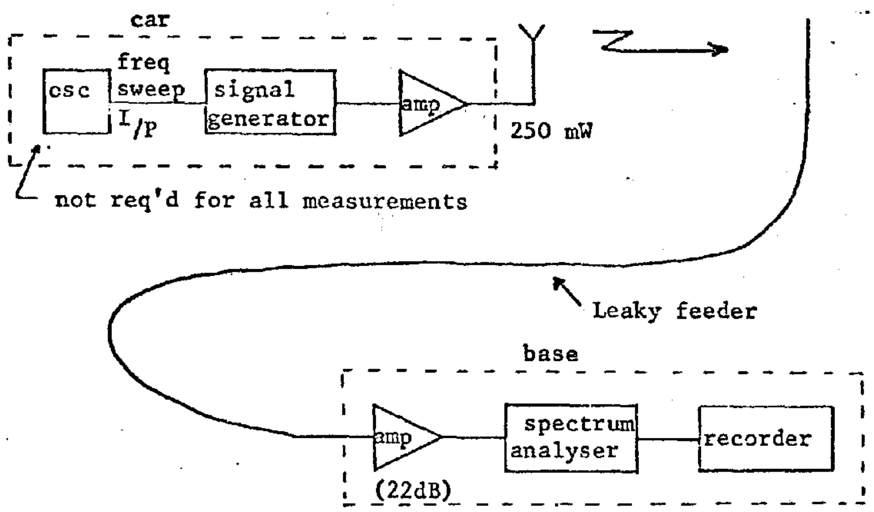

Figure 8 Equipment Arrangement

The traces shown in figure 9 are three typical results. The short standing waves are equal in length to one half of the free space wavelength. As can be seen the received signal level can fall by as much as 30 to 40 dB below its maximum. These fluctuations appear to be caused by reflections of the non-coaxial modes at the ends of the line.

3.2 Frequency Domain Measurements

Accurate measurements were made of received signal level against position of the car over a swept frequency range from 39 to 44 MHz. Some results are shown in figure 10.

The movement of the nulls is consistent with predictions based on far end reflections as the cause of the standing waves.

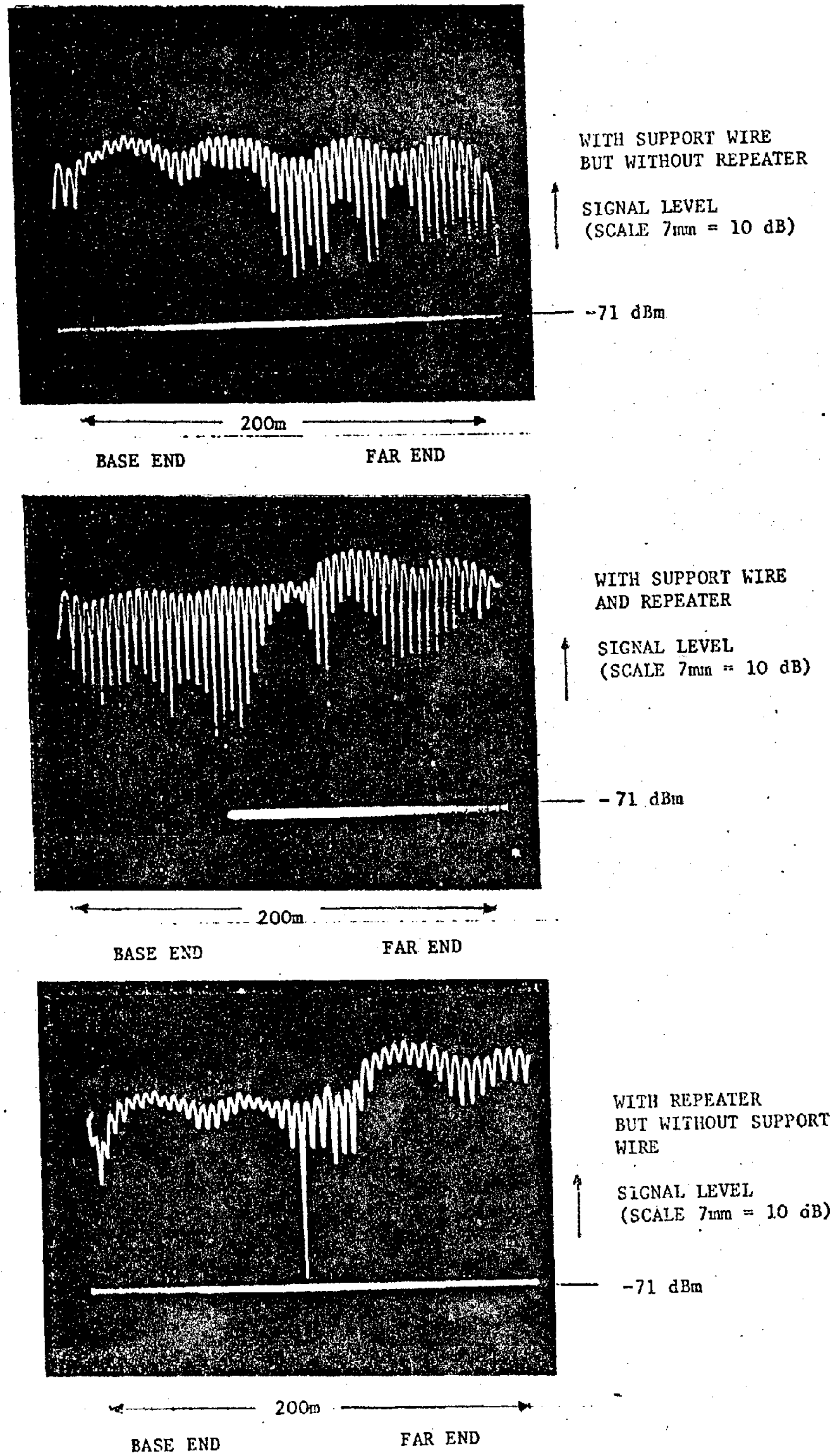

Figure 9 Typical Traces of Received Signal Level Variations Over Supported Section of Leaky Feeder Line

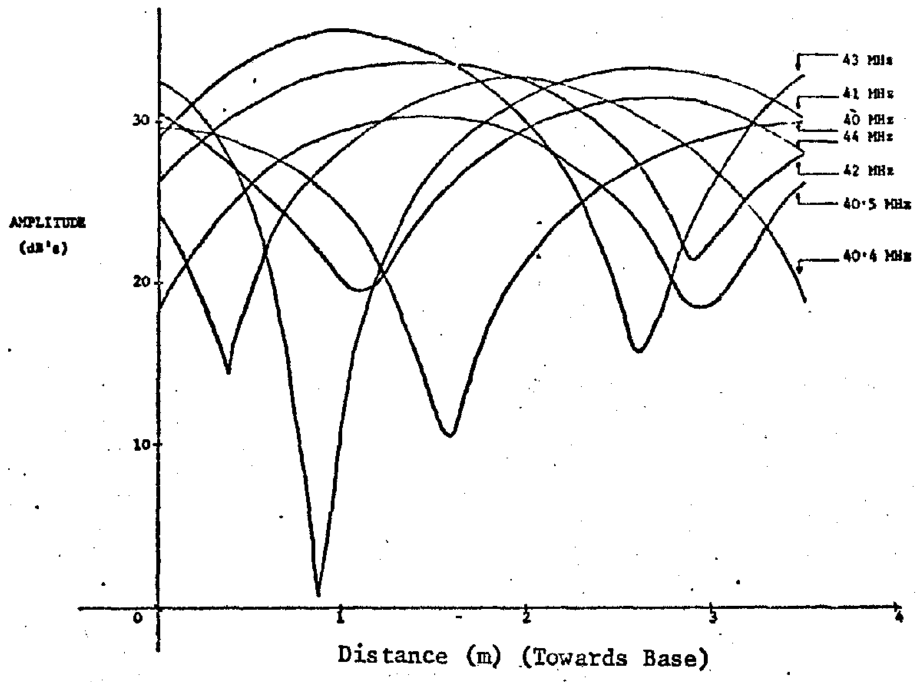

Figure 10 Received Signal Amplitude Versus Distance as a Function of Frequency

A similar test was performed in the tunnel using the arrangement of figure 11. The vertical rod aerial was placed on a metal tray, about 0·5 m square, on the ground in the centre of the tunnel. As seen from figure 12 the fluctuations are small, both with frequency and position. A similar result was obtained with the aerial raised 0·5 m. This result should be compared with figure 4 to see the effect on signal level variations of change in frequency from 40 to 72 MHz, in the tunnel.

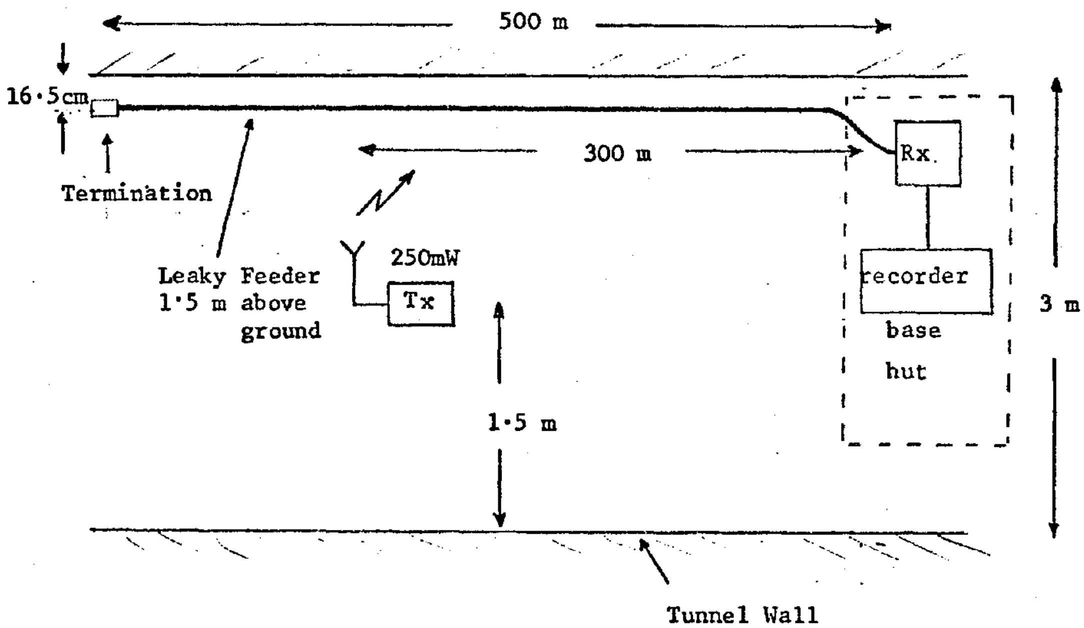

Figure 11 System Arrangement for Tunnel Measurements

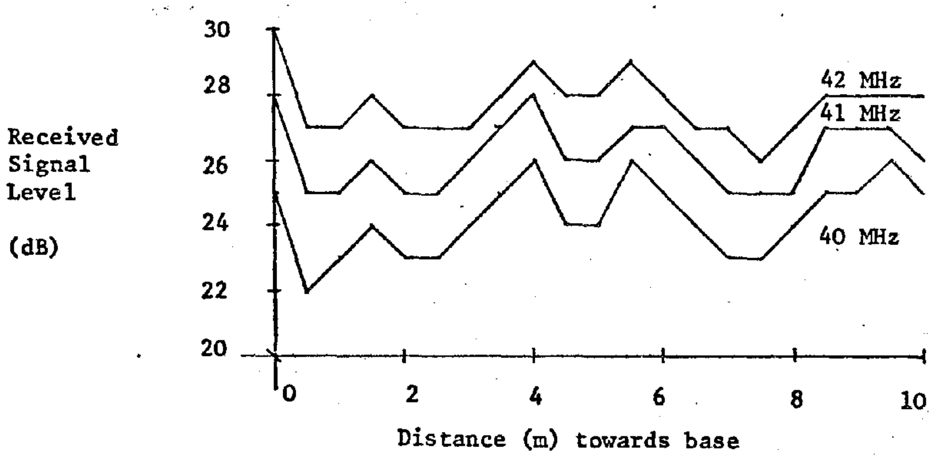

Figure 12 Received Signal Amplitude Versus Distance as a Function of Frequency

From figure 10 it is possible to determine the advantage to be gained, when in a null, by shifting frequency by a value Δf, that is of using the better of two signals separated by Δf in a frequency diversity system.

Figure 13 shows averaged results and these are compared with a theoretical performance curve derived on the assumption of a simple reflection at the far end of the line and nulls of depth 30 dB.

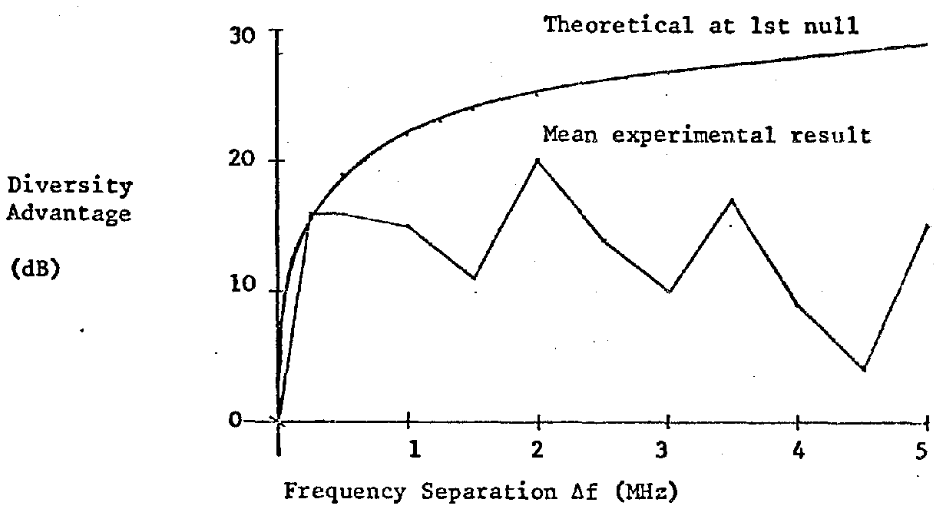

Figure 13 Diversity Advantage as a Function of Frequency Separation Δf

3.3 Mode Losses

Measurements were made of the propagation losses of the mode supported by the co-axial outer conductor and the earth, as a function of height and frequency, also of the combination of galvanised wire and earth, together with some low frequency ground constant measurements. Work on this aspect continues.

Figure 14 shows how these line losses vary.

SINGLE WIRE MODE ATTENUATION

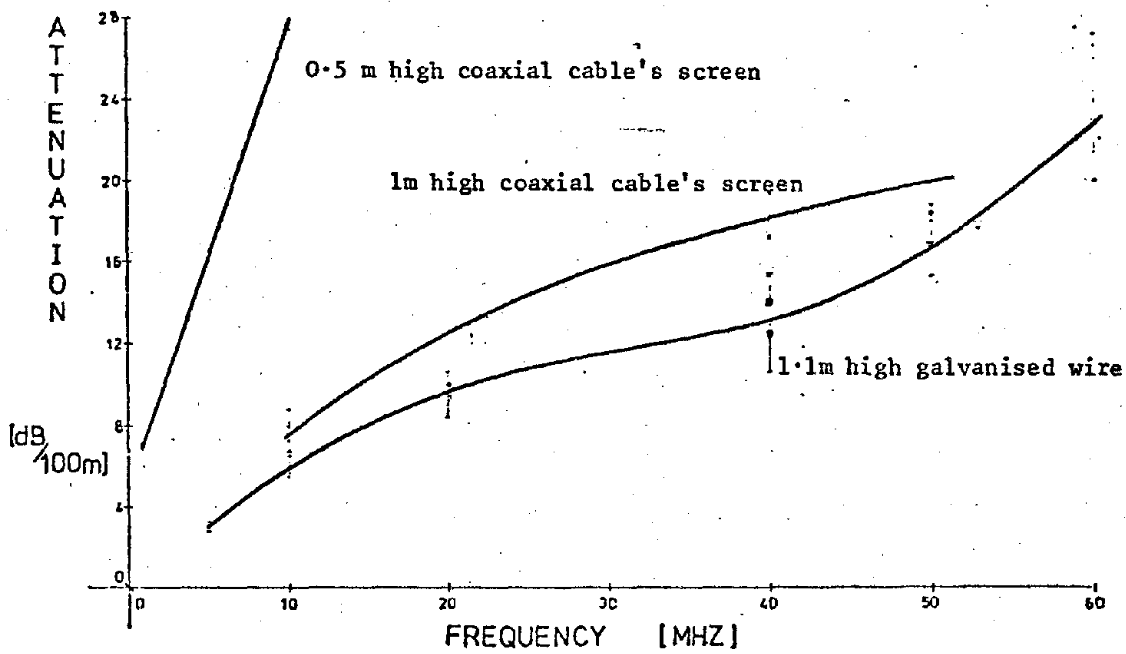

Figure 14

CONCLUSIONS

Much remains to be done but it seems that in the tunnel there could be advantages in lowering the frequency below that at which waveguide modes propagate. Correct choice of polarisation could also be helpful in eliminating standing waves. The H-vector measurements need to be supplemented by the electric equivalents before a clear picture can be established of the full detail in the tunnel.

Standing waves are much more severe above ground than in the tunnel. In particular when the galvanised iron support wire was present standing waves were particularly bad. The presence of this wire permits propagation of a low-loss bifilar mode and it is reflections of this mode at the far end of the feeder which contribute most to the standing waves.

Frequency diversity systems can give significant gain when signal fluctuations are present and at 40 MHz a frequency separation of anything over 500 kHz is probably sufficient.

The results indicate that high data rate (100kbits/s) communication can be achieved with error probabilities of about 1 in 10^4. With diversity and with correct matching to eliminate standing waves significantly better performance is possible.

COMMENT BY N. MACKAY

I would suggest that you make your measurements in such a way as to reduce all effect of the probes, transmitter, etc. We have found at Queen's that the only reliable means of measuring fields of this nature is to use telemetry links and very small probes. Even our probes are suspended by nylon cord and moved mechanically by remote means.

REPLY BY D.J.R. MARTIN (FOR Q.V. DAVIS ET AL)

The authors established that changing the disposition of the considerable slack in the auxiliary cables did not affect the readings, nor did the movements of personnel taking those readings. The probe itself was located on a wooden frame, and they deduced from this that the very presence of those cables and people was not significant to the results. However, they do appreciate that it would be better to remove all such extraneous cables, people, and objects if a suitable telemetry system (for amplitude and phase) could be devised.

COMMENTS BY J.R. WAIT
(Communicated to Q.V. Davis, 20 March 1978)

I was pleased to see you did those monofilar tests that we talked about just a year ago. These have just the expected form, i.e. the attenuation rate varies linearly with frequency in the region below 10 MHz for low wire heights while for higher frequencies and/or larger heights the attenuation varies roughly as $(f/\sigma)^{\frac{1}{2}}$. This behaviour is covered in my paper in Int'l. Jour. of Electronics, Vol. 42, No. 4, 377-391, 1977 where I discussed both uncoated and coated conductors over a half-space and at a fixed distance from the tunnel wall. Copies of this reprint are being distributed to workshop participants.

Your other results are also very informative. However, it appears that the 200 meter cable is not sufficiently long to really examine the nature of the modes on such uniform structures. However, it does seem that there is evidence of an interference or beating between the monofilar and bifilar modes in addition to the strong reflections at the end of the cable.

REPLY (COMMUNICATED) BY R.W. HAINING

Many thanks for your letter of 20 March, and in particular for your comments on our test results.

Further details of the BICC cable type T3522 are: Construction - PE sheathed apertured copper tape coaxial cable; Characteristic impedance - 75 ohms; DC resistance - 4.9 ohms/km outer, 4.1 ohms/km inner; Surface transfer impedance - 2.1 ohms/m at 30 MHz; Diameter of outer conductor - 10.1 mm; Diameter of sheath - 13.1 mm; Velocity ratio - 0.87; Coaxial mode attenuation - 25 dB/km at 40 MHz.

(for further comments see page 310)

UNDERGROUND RADIO COMMUNICATION TECHNIQUES AND SYSTEMS IN SOUTH AFRICAN MINES

B.A. AUSTIN
ELECTRONICS DIVISION
CHAMBER OF MINES OF SOUTH AFRICA
RESEARCH ORGANISATION
CORNER KEW ROAD AND LANDAU TERRACE
RICHMOND, JOHANNESBURG

Abstract

This paper reviews the research and development conducted within these laboratories over the past twenty-five years in the field of underground radio communication for use in mines. The results of measurements of the electrical characteristics of rock typical of the South African deep-level gold fields are presented, as are measurements of the electromagnetic propagation characteristics at various frequencies. The effects of electrical conductors, such as power cables, pipes and rails on signal propagation are noted and discussed, and the coupling between antennas and wave-guiding mechanism is examined. The paper also presents a review of the equipment developed within these laboratories for satisfying the various communication requirements and stresses the need for minimum size, yet exceptionally rugged, units which satisfy the requirements of intrinsic safety. The use of radio in mine rescue and fire-fighting applications is discussed and great emphasis is placed on a medium frequency system which propagates, over useful distances, directly through the rock strata. In many emergency situations no man-made wave-guiding system can be relied upon, particularly after falls of ground, and direct radio penetration of rock is vital to the success of the operation. Mention is also made of large scale installations using multiple base stations which have been operated very successfully in South Africa and have proved the feasibility of using such a system, remotely controlled via a two-wire line from surface, if necessary, to provide coverage throughout a mine. The problem of operator training in the use of radio and acceptance of it is also touched upon.

1. INTRODUCTION

The need for an effective and reliable means of communication between an underground fire-fighting team and their base was responsible for an investigation being conducted into the feasibility of using radio methods in mines. At the request of the Chamber of Mines an investigation into electromagnetic propagation through rock strata was conducted by the Telecommunications Research Laboratory of the South African Council for Scientific and Industrial Research in 1948.

Their report [1], presented in 1949, showed that electromagnetic signals would propagate through rock strata over distances of some hundreds of metres if the frequency of operation were in the medium frequency range of the radio spectrum. Of prime importance in an emergency situation is the need for equipment portability, and this restriction on size is one of the limitations on such a radio system, inasmuch as the antenna dimensions must allow for portable operation. Wadley [1], showed, after measuring in the laboratory, the conductivity σ, and relative permittivity ε_r of representative samples of quartzite and Kimberely shale that the rates of attenuation with distance were between 0,02 dB/m at 100 kHz and 0,06 dB/m at 1 MHz. Based on these figures and on the radiation performance of electrically small antennas, he produced a family of curves, Fig. 1, with distance between transmitter and receiver as the parameter, which showed that communication would be possible, directly through rock, over a distance of 600 m if the frequency of operation were about 300 kHz.

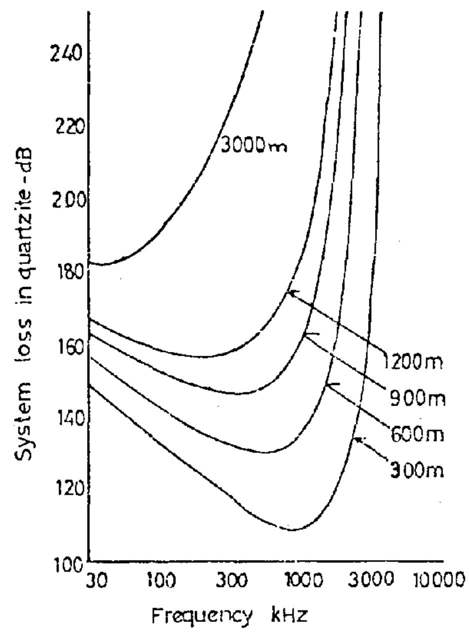

Fig.1.System loss in quartzite.

It was this information which prompted the Chamber of Mines to develop a radio system which would satisfy the needs of a fire-fighting, or "Prototeam", as it is called, by providing the necessary two-way communication. The peculiar set of circumstances which apply in underground fire-fighting operations impose their own restrictions on voice communication, and these will be discussed later in this paper.

11. THE UNDERGROUND ENVIRONMENT

South African gold mines are deep, the deepest vertical shaft being 3 596 m, the rock is amongst the world's hardest and the virgin rock temperature can be as high as 50°C. The services, such as electrical power require a mass of electrical cables for surface to underground sub-stations for reticulation throughout the haulages and travelling ways. Both compressed air and water is channelled around the mine in a vast network of steel pipes of various diameters, which are supported on the side walls of haulages and travelling ways. The movement of rock, men and materials is accomplished by both electric and diesel-powered locomotives running on steel tracks.

From an electromagnetic point of view the environment is complex. Over the great depths involved considerable stratification occurs with quartzite being the material in which gold-bearing conglomerates are found. Lying above the quartzite layers are andesitic lavas, and interspersed with the quartzite are bands of shale.

Wadley's [1], measurements in the laboratory provided the necessary information that radio propagation at medium frequencies would be feasible over useful distances, but the absolute range obtainable would depend very much on the in situ characteristics of the rock in any given area. The reasons for this are basically two-fold with the lack of homogeneity of the rock contributing a different "bulk" conductivity compared with the values measured in the laboratory [2], for homogeneous samples, and the fact that rocks in a laboratory environment and those at depth underground are vastly different as far as temperature, pressure and water content are concerned [3].

Fig. 2 shows values of electrical conductivity(σ) versus frequency for three rock types, samples of which were obtained from South African gold and platinum mines. The measurements were made at room temperature on freshly-cut samples from borehole cores. Fig. 3 shows values of relative permittivity (ε_r), versus frequency for the same samples.

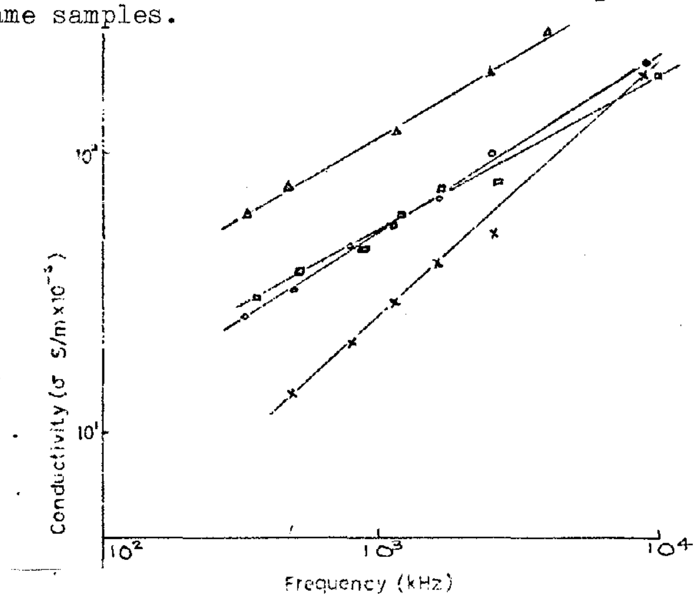

Fig. 2. Variation in conductivity with frequency for (Δ) pyroxemite, (○,□) quartzite conglomerate, (×) and anorthosite.

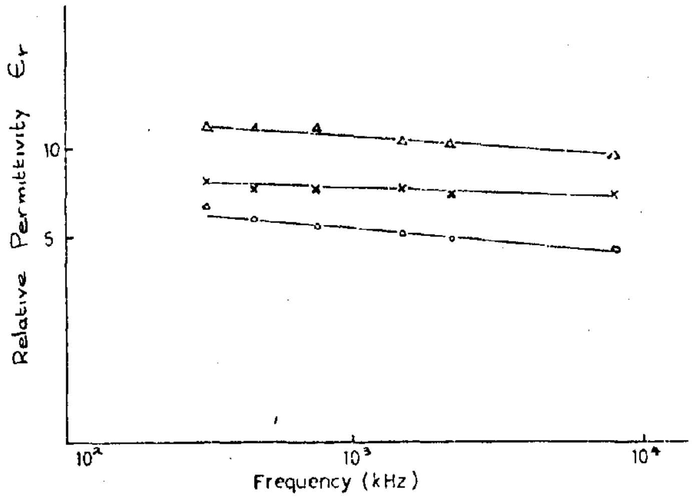

Fig. 3. Relative permittivity of the same samples.

111. UNDERGROUND MEASUREMENTS

Numerous tests have been conducted over the years in a wide variety of mines and mining situations in order to obtain a broad background on radio propagation phenomena underground. Communications requirements differ depending on the situation, and whereas the equipment initially developed was designed for use in rescue and fire-fighting applications the ranges achieved with it soon suggested that benefit could be derived in other areas of underground activity such as normal mining operations, and in engineering functions. This broadening of interests meant that portable-to portable contact was a requirement as well as portable-to-base. Transmitter output power and antenna size are thus both restricted when portable apparatus is required and the attainable range decreases. Qualitative tests over many years established that in areas devoid of any conductors, base station-to-portable communication was possible in gold mines over distances from 30 m to 800 m. This great variation in range is due to the local geology and hence no hard-and-fast rules can be laid down as to the useful operating range. It has been found however that by judicious siting of the equipment, communication can usually always be provided to suit some particular requirement.

With the assistance of conductors, particularly power cables, but also in areas where only pipes or rails exist, ranges well in excess of 1 000 m are attainable, with no effort being required on the part of the user. The mere presence of a power cable, for example, within

2 m or 3 m of the operator is sufficient to allow excellent quality voice communication over distances up to 3 000 m with another user similarly sited near the cable.

In order to try and understand the mechanism of propagation in areas not served by electrical conductors, a series of measurements were made at mf and hf. The area used for the experiment was in a haulage on 11 level of the Durban Deep Gold Mine near Johannesburg. The geology of the area consisted of quartzite interspersed with shaley bands. The depth below surface was approximately 500 m. Measurements were made of the variation in signal strength received from a transmitter at various points within the haulage as the distance between transmitter and receiver was changed. These tests were conducted at six separate frequencies, these being 335 and 903 kHz and 2, 4, 6 and 8 MHz. Small, typically 0,5 m diameter, single turn loops were used as both transmitting and receiving antennas, and particular note was taken of the orientation of these antennas when readings were taken.

The haulage was about 3 m square and tests were conducted with the transmitter and receiver within line of sight and at points when a 45° corner intervened in the path. No dramatic difference in readings was obtained, suggesting minimal affect of the air path on propagation.

Fig. 4 is typical of the results obtained [4], and shows the variation in received signal at 335 kHz with distance between transmitter and receiver with antenna orientation as the parameter.

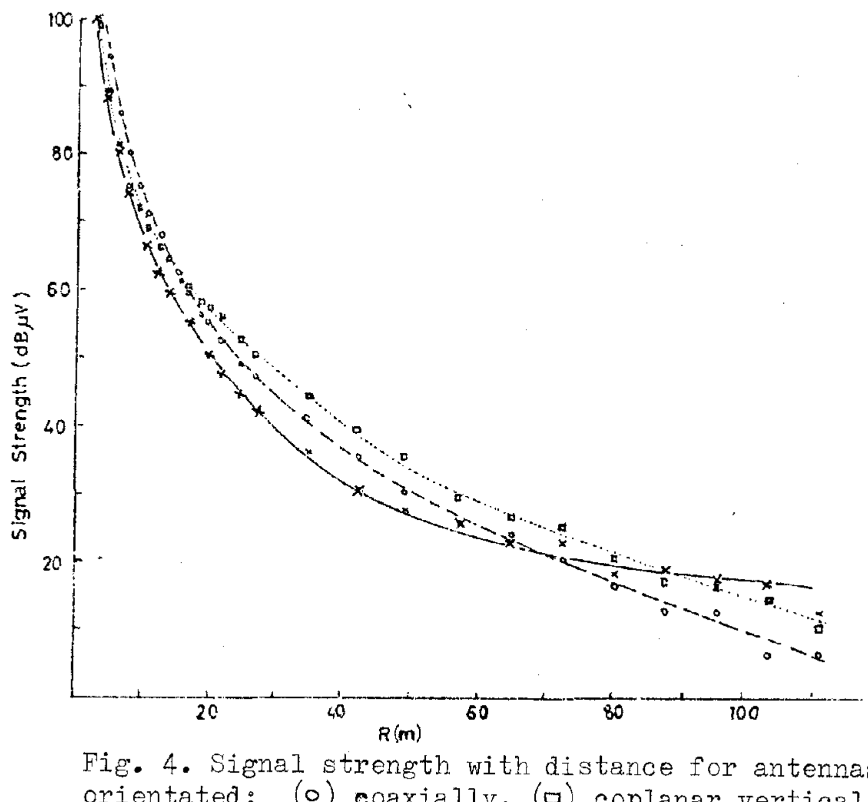

Fig. 4. Signal strength with distance for antennas orientated: (○) coaxially, (□) coplanar vertically and (×) coplanar horizontally.

The limit of 110 m was imposed by the measuring equipment and was not indicative of the maximum usable range obtainable. Perfect voice communication was possible 235 m apart, at which point further separation was impossible due to the configuration of the mine. Of particular interest is the effect of antenna orientation on signal strength. It will be noted from Fig. 4 that when source and receiver are close together, that is 15 m and less apart, the coaxial orientation of antennas produces maximum signal strength. Beyond about 15 m the coplanar vertical orientation takes over, and is itself surpassed by the coplanar horizontal orientation at about 90 m separation. The speech test at 235 m separation confirmed very definitely that the coplanar horizontal orientation was optimum. This result was somewhat surprising, and could only be explained as being due to a wave-guiding mechanism which was characteristic of the surrounding medium in the particular test area. It is interesting to compare this result with that of Emslie and Lagace [5], where best performance in their particular underground situation was found to occur with the antennas were orientated coplanar vertically. Their postulation of a particular wave-guiding mechanism produced good agreement between theory and practice. In the case reported here the propagation mechanism would seem to be somewhat different.

A least squares fit of the data shown in Fig. 4, based on the assumption that the coplanar horizontal antenna orientation was dominant in the "far field", produced a rate of attenuation α equal to 0,12 dB/m, see appendix A.

Similar measurements were made at five higher frequencies and results obtained are shown in Fig. 5. It is interesting to compare these measured values with those calculated by Wadley, from his laboratory rock sample measurements and this is done in Fig. 5.

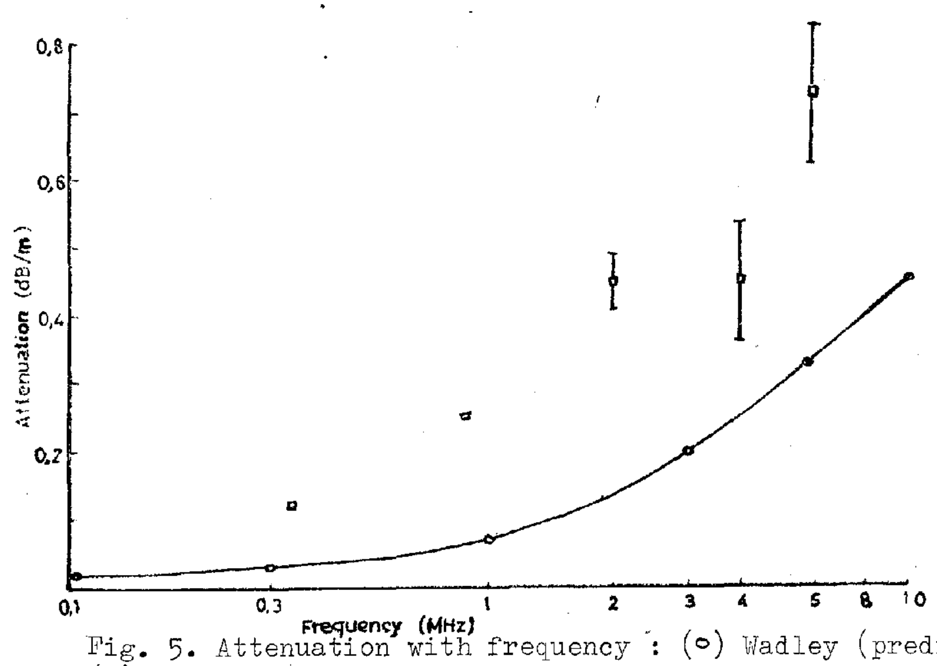

Fig. 5. Attenuation with frequency : (○) Wadley (predicted) (□) measured.

Based on the fact that the permittivity of the medium is virtually constant over the frequency range of interest (see Fig. 3), and on the general expression for the attenuation constant α, [6], an expression can be obtained for the conductivity, σ, of the medium,

$$\text{giving} \quad \sigma = \omega\varepsilon \left[\left\{ \frac{2}{\mu\varepsilon} \left(\frac{\alpha}{\omega}\right)^2 + 1 \right\}^2 - 1 \right]^{\frac{1}{2}}$$

where $\varepsilon = \varepsilon_0 \ \varepsilon_r$

$\mu = \mu_0$

$\omega = 2\pi f$

The results of this calculation are presented in Table 1.

TABLE 1

FREQUENCY DEPENDENCE OF ROCK CONDUCTIVITY

Frequency (kHz)	Conductivity σ (S/m) min	max
335	$3,35 \times 10^{-4}$	
903	$4,46 \times 10^{-4}$	
2182	$6,6 \times 10^{-4}$	$8,2 \times 10^{-4}$
4066	$5,4 \times 10^{-4}$	$8,4 \times 10^{-4}$
6052	$9,5 \times 10^{-4}$	13×10^{-4}

It will be noted that the conductivity measured in the laboratory was about two orders of magnitude lower than that calculated from propagation studies.

It should also be noted that the results presented above are dependent upon the transmitting and receiving antennas at any one frequency, having similar characteristics as they are moved from point to point. This assumes that the effect of the surrounding lossy medium on the antenna characteristics was constant throughout the test area. No verification of this was possible.

IV PROPAGATION VIA PIPES, CABLES AND RAILS

During the course of many underground tests and demonstrations of radio equipment the metallic objects in the vicinity had on propagation were noted and exploited. Whereas the maximum range attainable without the assistance of conductors might be 300 m, that obtained merely by having cables spanning the area in question, could be 3 000 m. Time and again this improvement in range by a factor of 10 or more has been noted. The coupling and propagation of energy via power cables, for example, is intriguing and the literature abounds with references as to the actual mechanisms which prevail. The work of Wait [7], [8], particularly, has shed much light on the problem and has gone a long way towards explaining many of the phenomena which have been encountered during underground testing. A working mine is not an ideal area in which to make quantitave assessment of this method of propagation because, more often than not, the array of cables, pipes, rails and conveyances make comparison between test areas very difficult.

It would be true to say that in the great majority of cases underground radio propagation as it is used on a day to day basis in South African mines is via this cable assisted mechanism. This is because so much of the activity in a mine takes place within a few metres at least of some conductor, be it a power line, rail, steel pipe or even winch rope. Experience has shown that two operators in radio contact need not even be coupling into the same conductor, as long as the two conductor systems are themselves in fairly close proximity to one another at some other point. The amount of "cross-talk" between cables, has been found to be very significant. The variation in signal strength encountered does not follow a standing wave pattern, but is rather a progressive decay.

The method used to couple energy into and out of the cables, pipes or rails is simple. The mere presence of any of these objects within some metres of the loop antenna of a radio transceiver results in signal strength improvement. The closer the loop is to the cable the better of course, but it is by no means a necessity for them ever to physically touch one another. The orientation of the loop relative to the cable is important. It has been found repeatedly that the plane of the loop must coincide with the direction of the pipe or cable into which it is coupling for maximum signal.

A significant finding during the course of experimental work associated with an underground-to-surface telemetry study, was, what has become known locally as the "sub-station effect". The requirement was to transmit telemetry information from three separate levels underground to a recording centre on surface. If indirect coupling of radio signals into the existing power system within the mine could be used to provide the necessary link between underground and surface, a considerable saving in time and money would be realized. Tests conducted between an electrical sub-station on surface and three sub-stations underground yielded excellent results as long as the antennas at each point were located fairly close to the common cable, within the sub-station. Moving any antenna a matter of metres outside of the sub-station, but still in close proximity to the same cable, caused a dramatic drop in signal strength. It has been postulated [9], that because the insulated conductors within an armoured cable, fan out to switchgear and transformers within a sub-station, a mode conversion [10], takes place in that the bililar mode [11], is excited on the inner conductors, whereas outside the substation only the monofilar mode is set up between the cable armouring and the surroundings and is propagated with a higher rate of attenuation [10]. Signals were transmitted very effectively by this means over a distance in excess of 3 000 m using 1 W transmitter power and small, 350 mm diameter, multi-turn loop antennas.

While on the subject of conductor assisted propagation it would be opportune to discuss the phenomenon as it applies in mine shafts. The structure of a typical vertical or incline shaft contains steel reinforcing members and an array of electric cables and steel cables or ropes, as they are called. Therefore it is entirely reasonable to expect electromagnetic energy to propagate within such a structure, and Wait and Hill, [12], have calculated the attenuation which might be expected at various frequencies.

Experience gained on many occassions in South Africa, with the same equipment as used underground, typically at 335 kHz and 903 kHz, has shown that communication is indeed feasible between an operator standing on surface, with the small loop antenna around his body and another operator, either within a steel cage travelling in the shaft, or standing on an underground station within 5 m or so, of the shaft. Numerous tests have shown the technique to be completely effective and reliable over shaft depths in excess of 1 700 m. Noteworthy here is the fact that no special toroidal, or other coupling devices are used in order to launch the signal into the shaft. Again the mere presence of conductors in the vicinity of the antennas is sufficient.

V THE DEVELOPMENT OF SUITABLE EQUIPMENT

Work was commenced during the early fifties to develop suitable hardware for use by Prototeams, based on the theoretical and practical results already obtained. A portable transceiver and base-station transceiver were designed, [13], and were used for underground trials and evaluation. The size of both items of equipment was the main disadvantage, but very worthwhile experience was gained by its use.

A second generation of transceiver was developed, [14], with marked reduction in size, and considerably improved performance through the use of single-sideband modulation A3J, as opposed to A3. Tests done at 1 W transmitter output yielded very promising results and suggested that for certain short-range applications a considerable saving in equipment size could be achieved by reducing the output power by 10 dB, and also use could be made of the predicted lower overall system loss if operation took place in the vicinity of 1 MHz (see Fig. 1).

A handheld SSB transceiver was designed which operated at 903 kHz and embodied a particularly simple and inexpensive method of SSB generation. The equipment was designed to merely provide evidence of useful range at the higher frequency and lower power. This it did so successfully that more than 100 units have been built during the past three years and they have been relied upon to provide all the communications in a mechanized stope at Doornfontein Gold Mining Company Limited near Johannesburg.

...

Fig. 6 shows the layout of this stope and the position of the rockcutting machines at the face. All machine operators are equipped with 903 kHz transceivers and communicate with either of the two stores from where the maintenance of the machinery is controlled. The travelling ways and gullys carry electric power lines and reinforced hydraulic lines and the signals propagate very effectively via these throughout this mechanized area. The base stations are merely higher powered versions of the handheld sets, but, because they are fixed installations, use can be made of larger antennas. These are 20-30 m circumference, single turn loops which are erected close to the electrical or reinforced hydraulic lines serving the face, and are equipped with suitable tuning and matching networks which are adjusted for optimum performance after installation.

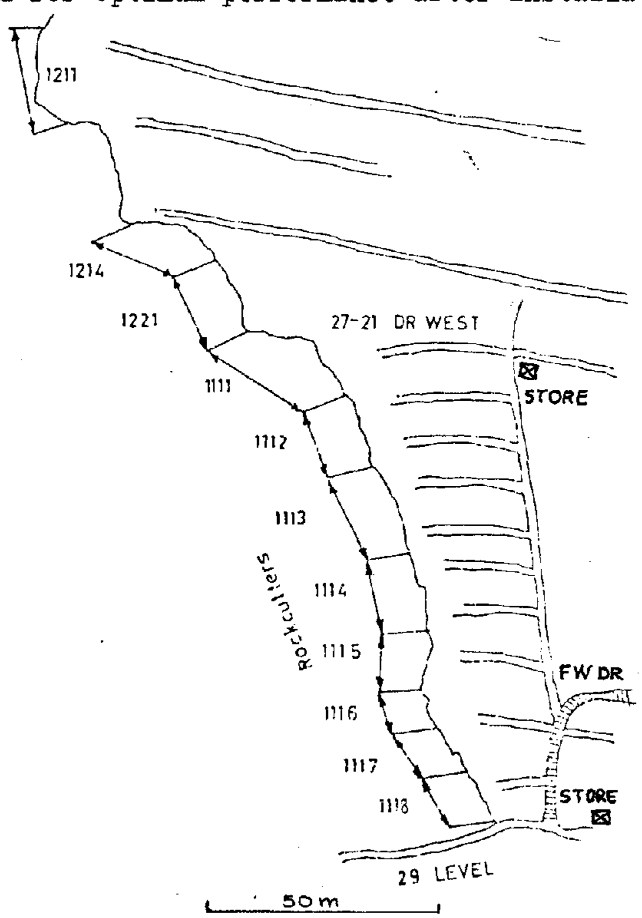

Fig. 6. Plan of Doornfontein gold mine mechanised stope.

An experiment was performed whereby a number of these base-stations, situated on different levels of a mine, were interconnected via a two-wire telephone line and remotely controlled. This remote control point was on surface and the base-stations were 2 500 m underground with a total length of line of about 3 000 m. All the base-stations were phase-locked to a 14 kHz reference signal transmitted down the lines from surface. It was found in practice however that phase-locking was unnecessary because any distortion caused at the signal overlap points from the various free running SSB transmitters was negligible. The minimum communication range from a base-station was about 150 m in some directions, but extended to more than 1 000 m along main haulage routes. The system was shown to be very effective and also has application in emergency operations where an unattended base-station can be positioned in a dangerous area, by a Prototeam for example, and be controlled from 3 000 m away.

The need for multi-channel equipment was now evident and a system has been developed, Fig. 7, which is frequency synthesized, providing 100 channels between 100 kHz and 1 MHz. The transceiver is attached to a web-belt assembly which contains the multi-turn loop antenna. When worn, minimum inconvenience is experieced by the operator, and both speech and tone signals may be transmitted, in order to accomodate the Proto brigadesman, who is unable to speak because of the snorkel type breathing apparatus which he wears. The receiver in the equipment is designed to exploit the fact that in many areas of a mine the electrical noise level is very low, or even non-existent, hence maximum receiver sensitivity, for the bandwidth available, can be used. The equipment is very rugged, waterproof and intrinsically safe, and, having only three controls, is very easy to operate. The requirement for possible paging facilities has been considered and will be incorporated should the need arise.

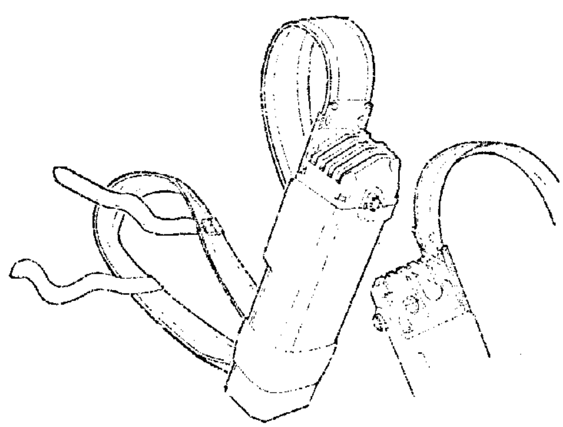

Fig.7. Underground Mine Transceiver.

VI OPERATOR ACCEPTANCE AND TRAINING

An area which is receiving considerable attention is that of the introduction of radio communications equipment to mines where previously the only form of contact between personnel has been via telephones. It became very evident during the period when the equipment was being tested that consideration must always be given to operator training. The possession of communications equipment is certainly no guarantee of communications, and a training programme which takes into account all levels of personnel involved must be part of the whole process of introducing radio equipment to the world of mining.

VIII ACKNOWLEDGEMENT

Acknowledgement is made to the Chamber of Mines of South Africa for permission to present this paper.

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APPENDIX A

Propagation of electromagnetic energy through a lossy medium may be described by the following expression, on the assumption that far field conditions apply.

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

where P_r = Power available at receiver terminals.
P_t = Power into transmitter antenna.
λ = Wavelength in the medium.
R = Separation distance between transmitter and receiver.
α = Attenuation constant of the medium.
G_t, G_r = Gains of transmitting and receiving antennas respectively.

If two measurements of received power P_{r_1} and P_{r_2} are made at distances R_1 and R_2 respectively from the transmitter, we may write

$$\frac{P_{r_1}}{P_{r_2}} = \left(\frac{R_2}{R_1}\right)^2 e^{2\alpha(R_2 - R_1)}$$

from which we obtain

$$\alpha = \frac{1}{R_2 - R_1} \ln \frac{V_{r_1} R_1}{V_{r_2} R_2}$$

Where V_{r_1} and V_{r_2} are the voltages across the receiver input at distances R_1 and R_2 respectively from the transmitter.

QUESTIONS BY R. GABILLARD

1. What kind of simplifing assumptions have you made for the calculation of the curves in Fig. 1?

2. Do you think it is possible to find a high frequency window in the kind of rocks where your mines are?

REPLIES BY B.A. AUSTIN

The figure is based on the addition of the losses due to the variation of rock attenuation with frequency given by

$$\omega\left[\frac{\mu\varepsilon}{2}\left\{\left(\frac{\sigma}{\omega\varepsilon}\right)^2 + 1\right\}^{\frac{1}{2}} - 1\right]^{\frac{1}{2}} ,$$

with typical values of σ and ε_r obtained from laboratory measurements of rock samples, and the "loss" with frequency obtaining for electrically small antennas. Any given small antenna follows a decreasing loss characteristic as the frequency is increased, with the absolute values being distance dependent. One of the factors contributing towards the H.F. window is σ being constant with frequency. Within the range of frequency measured with laboratory samples, this constant σ was not detected. This does not rule out the possibility, however, of suitable geological conditions existing in some areas underground.

COMMENTS BY J.C. BEAL

The main limitation of the South African system is very limited bandwidth, e.g. one voice channel; but leaky cable systems do have the major advantage of greater bandwidth allowing for multivoice channels and telemetry simultaneously.

REPLY BY B.A. AUSTIN

For voice communication using SSB, the bandwidth available with antenna Q of 100 is about 3 kHz, which is ideal. The highest frequency used in this equipment is about 1 MHz, allowing a 10 kHz bandwidth for the same Q. For other applications, therefore, a wider bandwidth is available at the HF end. It is felt that, for very many mining applications, the fact that MF allows the use of existing metallic conductors as guides rather than requiring special purpose cables, very much outweighs any bandwidth limitation.

QUESTION BY G.A. COLLINS

Regarding antenna size, you mentioned 30m. Have you looked at "super" loops with the same dimensions of the mines?

REPLY BY B.A. AUSTIN

We have not used any loop larger than about 30m circumference. Loops of circumference of a typical mine might be many kilometers in length. This might lead to installation and maintenance problems, and the loop would not be electrically small and this might introduce some undesirable directional characteristics.

COMMENT BY J.R. WAIT

The non-uniform current distribution on large loops would cause numerous problems, particularly in direction-finding applications.

QUESTION BY R.L. LAGACE

Why was SSB the obvious form of modulation to use from your application standpoint?

REPLY BY B.A. AUSTIN

The antenna design required maximum Q. typically $Q = 100$ at 300 kHz, giving a bandwidth of 3 kHz. Narrow band FM offers no S/N advantage over AM; in addition, FM suffers from a "threshold effect" at low S/N, which justifies using SSB. In addition, the low electrical noise level experienced in many mining area allows exploration of maximum receiver sensitivity. Typically, 0,1μV sensitivity for 10 dB S+N/N in the 3 kHz BW. The lower TX duty cycle of SSB conserves battery power.

TRAPPED MINER COMMUNICATIONS AND LOCATION IN METAL/NON METAL MINES

A. J. Farstad

Westinghouse Electric Corporation
Geophysical Instrumentation Systems
3655 Frontier Avenue
Boulder, Colorado

ABSTRACT

Electromagnetic propagation experiments have been conducted at a series of metal/non metal mines in the U.S. to evaluate the feasibility of using wireless uplink and sidelink communications in these mines in the event of mine emergencies. Some of the earlier work done for coal mine trapped miner communications does not directly apply to metal/non metal mines because of their significantly greater depths, wider variability in conductivity and different mine configurations and mining methods.

The results of this program clearly show that through the earth ULF systems are indeed feasible for these deeper mines provided that certain modifications are made in the frequency range and transmitting duty cycle of the trapped miner system as well as in the method of receiving the signals on the surface. To make the system compatible with the needs and desires of the mining industry, it is recommended that the additional sophistication required to detect and process the weak signals associated with the deeper mine depths be concentrated in the surface receiving antenna and signal processing equipment.

1.0 INTRODUCTION AND SUMMARY

The U.S. Bureau of Mines has developed an electromagnetic ULF system for detecting, locating and communicating with miners trapped in underground coal mines. This system operates on the principal of through-the-earth coupling between a transceiver carried by the miner underground and a transceiver on the surface operated by the rescue party. Signal strength field patterns observed on the surface are used to locate the miners position, while communications are established by a baseband voice downlink and a narrowband code uplink.

The trapped miner communications and location system has been fabricated in a ruggedized package and is presently undergoing extensive testing in numerous coal mines in the eastern coal fields of the United States. Results have shown that the system performance is more than adequate for the vast majority of coal mines in the U.S.

The Bureau of Mines has also sponsored a program to determine the requirements for trapped miner communications and location in metal/non metal mines, which as a rule are considerably deeper than the average coal mine. The objective of this program was to determine whether the existing trapped miner hardware could be effectively used in the metal/non metal mine environment and to identify whatever modifications would be needed to adapt its performance to the deeper class of mines. This paper discusses the metal/non metal mine environment along with several candidate schemes for communicating with and locating trapped miners in metal/non metal mines.

2.0 METAL/NON METAL MINE CHARACTERISTICS

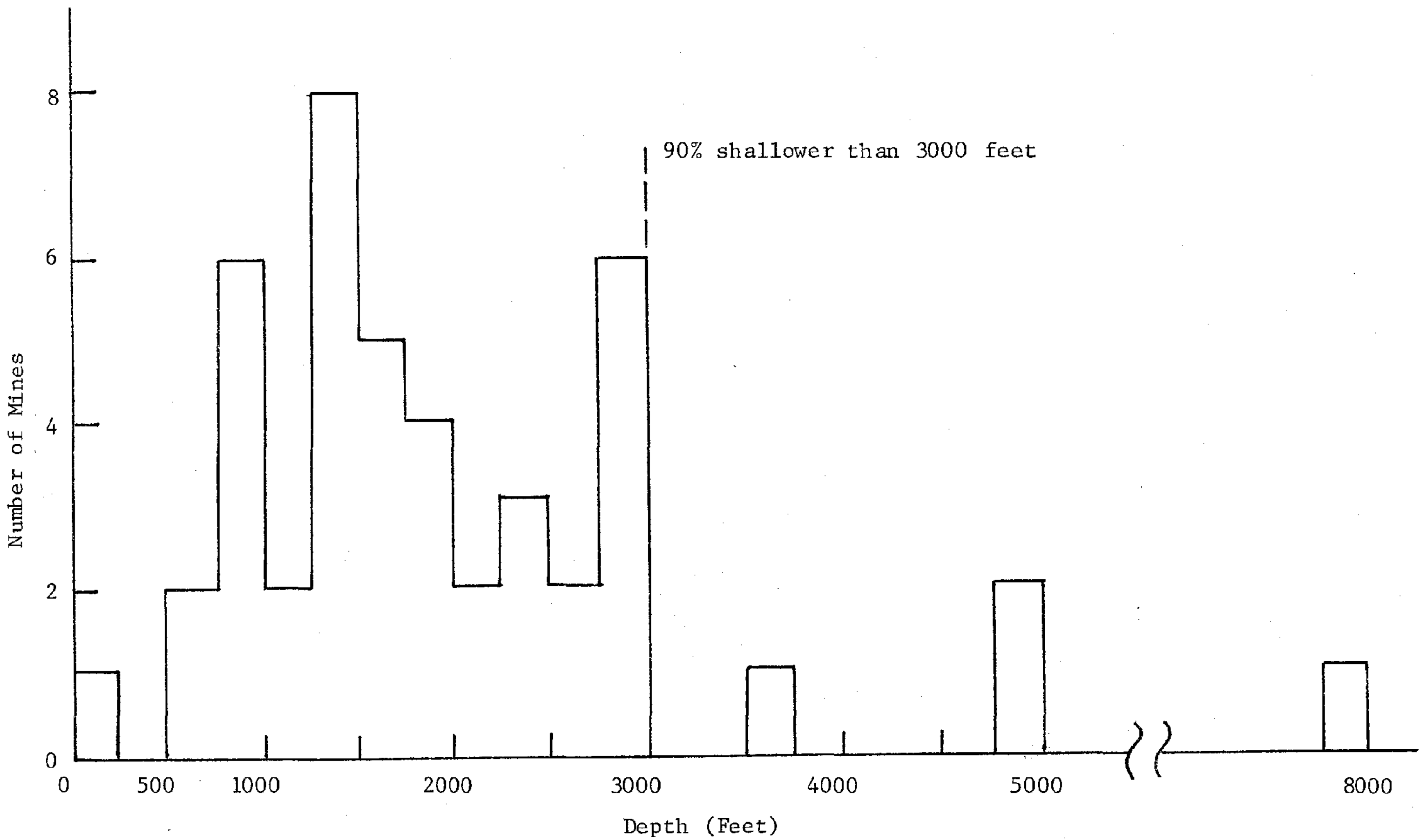
Mine Depth

The most critical mine parameter affecting the performance of a manpack trapped miner location system is overburden depth. Information on the distribution of maximum working mine depths in 45 of the major metal/non metal mines of the U.S. was obtained from the mine operators and is shown in the histogram of Figure 2-1. The 10% exceedance level (i.e. the depth exceeded by only 10% of all mines) for the 45 working mines is 3000 feet. If we use 3000 feet as a design goal, such that a manpack trapped miner location system must perform in all mines having depths up to 3000 feet, it will be operational in approximately 90% of metal/non metal mines in the U.S., assuming the sample to be truly representative of metal/non metal mines in general.

Overburden Conductivity

Overburden conductivity measurements were made at 7 mines on this contract. The results of these conductivity measurements were combined with measurements made by Westinghouse personnel on previous mine visits to arrive at the conductivity distribution curve for the

Figure 2-1 Distribution of Maximum Working Depths of Metal/Non Metal Mines in U.S. (Representative Sample - 45 mines)



17 mines shown in Table 2-1 and in Figure 2-2. Based on this particular representative 17 mine sample, the 10% exceedance level (i.e., the level of conductivity exceeded by only 10% of the mines) is about 1.15×10^{-1} mhos/m. It is important to note however that this unusually high value of conductivity for the 10% level was influenced greatly by two mines in particular; Berkely Pit Copper Mine in Butte, Montana and the International Salt Mine in Avery Island, Louisiana. Both of these mines had relatively shallow maximum overburden depths, 700 and 880 feet respectively. Thus it may be somewhat pessimistic to relate this particular 10% level to all metal/non metal mines since the deepest mines such as those found in the Coeur d'Alene district of Idaho and the Homestake gold mine in Lead, South Dakota have measured conductivities substantially lower than 10^{-1} mhos/m.

3.0 METHODS OF COMMUNICATING WITH TRAPPED MINERS

General

In determining practical methods of establishing rescue communications with trapped miners, one must have an understanding of: (1) The mine environment, (2) the mining industry, and (3) the mining personnel. Any electronic hardware designed for use underground must have the following characteristics:

(1) Must be reliable, even though infrequently activitated.

(2) Must be relatively inexpensive.

(3) Must be easily maintained.

(4) Must be simple to operate.

(5) Must be portable.

On the other hand, the equipment on the surface can be as complex and sophisticated as required since under normal rescue operations, those operating the surface equipment will be technicians thoroughly trained in its operation.

Basically there are two different means of coupling the surface and underground receivers to establish communications:

(1) Through the earth coupling.

(2) Coupling via wires or intermediate repeaters.

3.1 Through the Earth Coupling

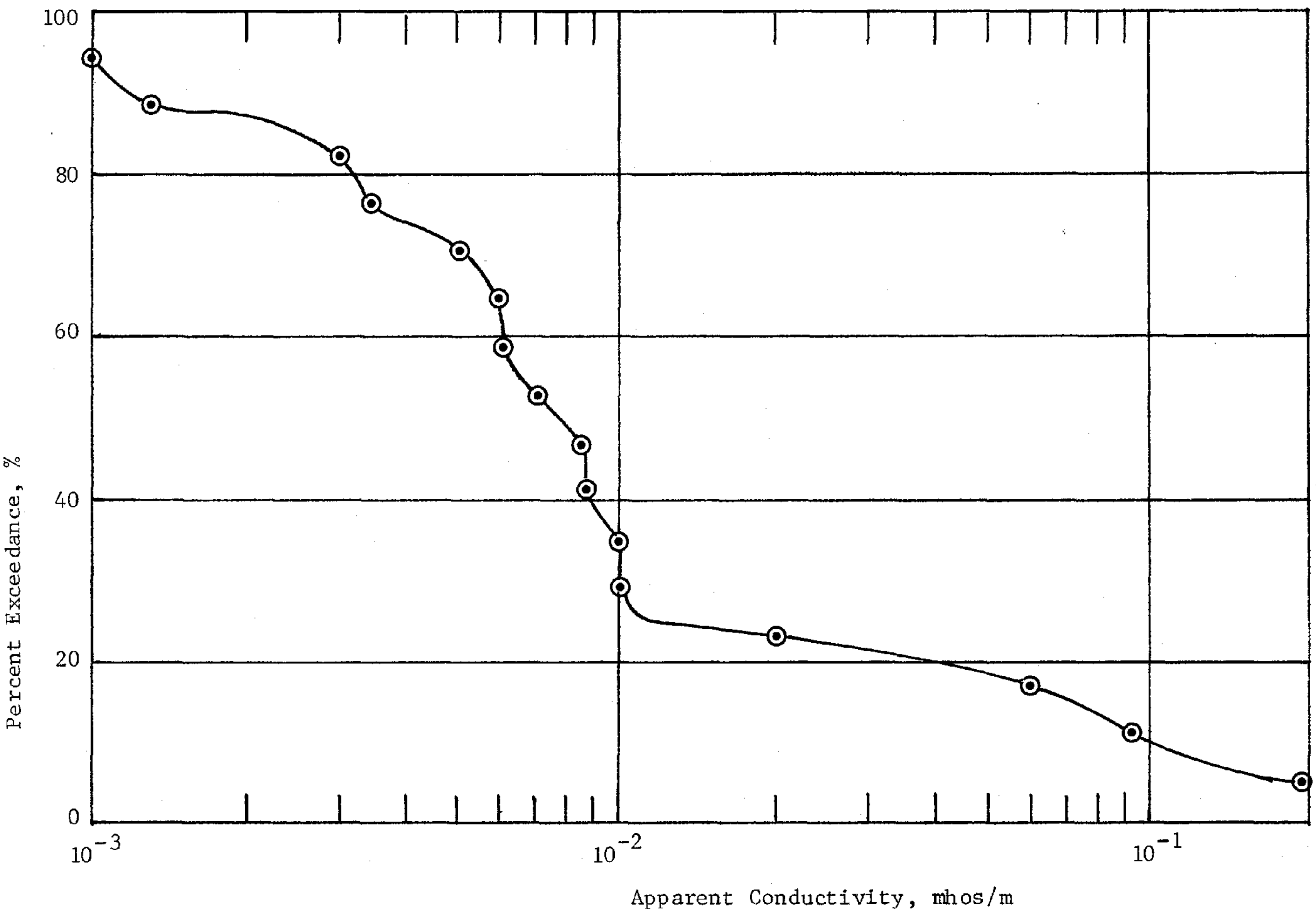
Much has been written in the literature about the behavior of low frequency radio waves in the earth (Ref. 1-4). Suffice it to say that low frequencies are needed to penetrate via through the earth

TABLE 2-1

Mine -Conductivity Distribution

Mine		Location	Apparent Conductivity	Percent of Exceedence
1.	Berkely Pit	Butte, Montana	1.95×10^{-1}	5.9
2.	International Salt	Avery Island, La.	9.3×10^{-2}	11.8
3.	Knob Hill	Republic, Wash.	6×10^{-2}	17.6
4.	Big Island	Green River, Wyo.	2×10^{-2}	23.5
5.	Fletcher	Viburnum, Mo.	1×10^{-2}	29.4
6.	Inexco	Jamestown, Colo.	9.9×10^{-3}	35.3
7.	LaTrobe	LaTrobe, Pa.	8.67×10^{-3}	41.2
8.	Copper Queen	Bisbee, Arizona	8.5×10^{-3}	47.1
9.	Ozark Mahoning	Rosiclare, Ill.	7×10^{-3}	52.9
10.	Galena	Wallace, Idaho	6×10^{-3}	58.8
11.	Caladay	Wallace, Idaho	5.9×10^{-3}	64.7
12.	Idarado	Telluride, Colo.	5×10^{-3}	70.6
13.	U.S. Tunnel	Idaho Springs, Colo.	3.4×10^{-3}	76.5
14.	Henderson	Empire, Colo.	3×10^{-3}	82.4
15.	Sunshine	Kellogg, Idaho	1.3×10^{-3}	88.2
16.	Homestake	Lead, South Dakota	1×10^{-3}	94.1
17.	TOSCO	Rifle, Colo.	1×10^{-3}	94.1

Figure 2-2 Conductivity Distribution of 17 Selected Metal/Non Metal Mines



paths especially in the deep metal/non metal mines described in Section 2.0. Simplified expressions for uplink and downlink through the earth propagation are as follows:

<u>Downlink</u> - Long Wire Transmitting Antenna

$$H_y = \frac{I\,|C|}{2\pi\, z} \quad \text{amperes/m}$$

<u>Uplink</u> - Dipole Transmitting Antenna

$$H_z = \frac{INA}{2\pi\, z^3}\,|G| \quad \text{amperes/m}$$

<u>Sidelink</u> - Dipole Transmitting Antenna (Coplanar Receiver)

$$H_z = \frac{INA}{2\pi\, d^3}\,|D| \quad \text{amperes/m}$$

where: I is the antenna current in amperes.
z is the overburden depth in meters.
$|C|$, $|G|$ & $|D|$ are attenuation constants for long wire and coplanar dipoles respectively, and are shown in Figure 3-1.

<u>Requirements for Through the Earth Uplink</u>

Based on the statistical study of mine parameters discussed in Section 2.0 for metal/non metal mines in the U.S., we can define a "worst case" mine as one having both a 10% exceedance depth and a 10% exceedance conductivity.

i.e., 10% exceedance depth = 3000 ft. (915 m)
10% exceedance conductivity = 1.15×10^{-1} mhos/m.

Assume f = 1000 Hz

Skin depth (δ) is computed as follows:

$$\delta = \frac{503.3}{\sqrt{\sigma\, f}} = \frac{1484}{\sqrt{f}}$$

$$= 47 \text{ meters at } f = 1000 \text{ Hz}.$$

Thus, 915 meters of overburden represents 19.5 skin depths of attenuation. This is consistent with an attenuation $|G|$ of 4.5×10^{-8}. The surface magnetic field strength (H_z) obtained in this case is computed as follows:

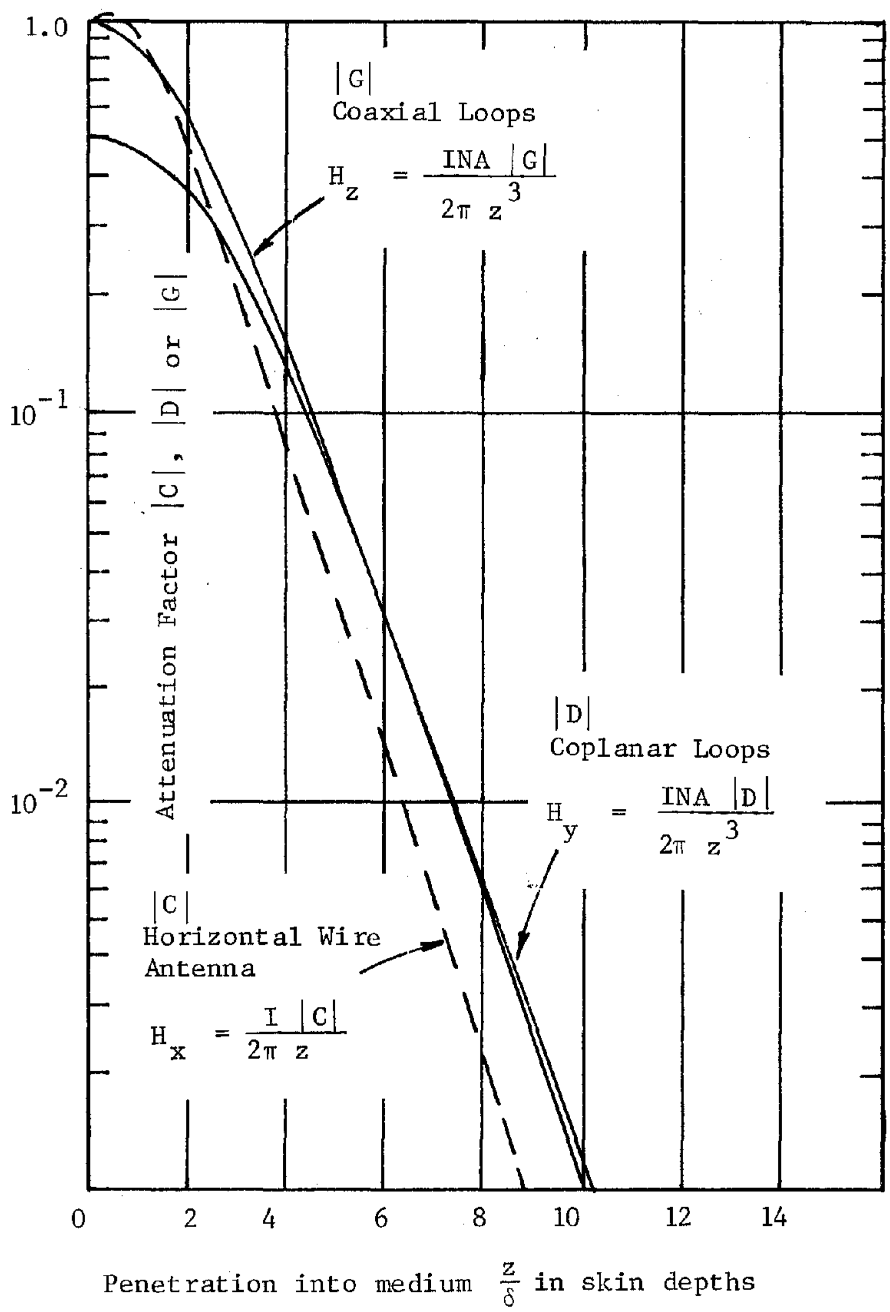


Figure 3-1 Electromagnetic Attenuation in Conducting Media

$$H_z = \frac{INA\,|G|}{2\pi\ z^3} = 9.43 \times 10^{-18} \times INA$$

where $|G|$ is the attenuation factor obtained from an extrapolation of the curve in Figure 3-1.

Assuming that a surface field strength of .1 μA/m is required for detection and location of a trapped miner, a magnetic moment of 1.06×10^{10} amp turns m^2 would be required for the 10% exceedance mine as described above. This value of transmitting moment is not practicable when one considers the portability and power restriction of the transmitter.

However if we reduce the frequency to 10 Hz, the skin depth becomes

$$\delta = \frac{503.3}{\sqrt{\sigma\ f}} = \frac{1484}{\sqrt{f}} = 470 \text{ meters at } f = 10 \text{ Hz}.$$

Thus, the 915 meter overburden depth is reduced to only 1.95 skin depths and the attenuation $|G|$ becomes .58. The transmitting moment required to penetrate this overburden is now computed as follows:

$$H_z = \frac{INA\,|G|}{2\pi\ z^3} = .1 \ \text{A/m}$$

$$INA = \frac{10^{-7}\ (2\pi\ z^3)}{|G|} = 830 \text{ amp turns } m^2.$$

which is a practical value for a trapped miner to achieve in an underground metal/non metal mine. Further enhancement of the detection capability on the surface can be achieved by using a long wire receiving antenna in place of a magnetic dipole (loop). By using a grounded wire electric antenna that is long with respect to the mine depth, the mutual coupling between it and an underground dipole antenna varies as inverse distance (1/z) rather than inverse distance cubed ($1/z^3$) as in the dipole-dipole case (3). The conventional use of the long horizontal wire antenna on the surface has been as a transmitting antenna to communicate a voice downlink signal to the trapped miners. However, the mutual impedance and the coupling relationship between the surface and underground antenna is the same regardless of which one is used as transmitter and which one is used as receiver. Thus the (1/z) coupling relationship is retained and signal detection capability is enhanced.

This concept has been proven out in practice on at least two occasions (5).

- At the Sunshine Mine in Kellogg, Idaho, a 1900 Hz uplink signal was received on a 4000 ft horizontal wire in a 50 Hz receiver bandwidth whereas the same signal could not be detected using a 500 turn 15" diameter receiving loop and the same receiver. (Overburden depth was 4800 ft.)

- At the Pocahontas Mine No. 1, Grundy, Virginia, a 1900 Hz uplink signal was received on a 2000 ft. horizontal antenna in a 3 Hz receiver bandwidth whereas the same signal could not be detected using a 500-turn 15" diameter receiving loop and the same receiver. (Overburden depth was 2300 ft.)

In view of the foregoing discussion, it can be summarized that detection of through the earth uplink signals can be accomplished in a so-called "10% exceedance" metal/non-metal using practical low power trapped miner transmitters, provided that frequency is reduced low enough to make through the earth attenuation relatively insignificant. For the 10% exceedance mine, frequency must be kept in the vicinity of 10 Hz for detection with magnetic dipole receive antennas but could go higher when using long horizontal wire receiving antennas.

There are logistic problems associated with using frequencies as low as 10 Hz when attempting to conduct hand-held direction finding operations. The earth's magnetic field induces high levels of noise in the loop for even the slightest movement of the loop. Consequently, when using a loop for direction finding of trapped miners at f = 10 Hz, the loop must be firmly planted in position using a tripod or equivalent, before readings can be taken at each individual position.

When using a horizontal wire antenna for direction finding, it is somewhat impractical to successively move the antenna around to locate the miners position as can be done with a loop. In the case of using a horizontal wire antenna for determining the trapped miners position, it would be more feasible to deploy an array of such antennas in the form of a grid network. Then, by analyzing the receiving pattern created by the signals received on all of the antenna elements, a judicial extrapolation of the trapped miners position can be made.

Requirements for Through the Earth Sidelink

In metal/non metal mines, unlike coal mines, there usually exists more than one level of mining activity going on simultaneously. Consequently, there is a requirement for knowing not only the miners horizontal coordinates but also on what mining level he is trapped. This introduces a requirement for horizontal propagation from the trapped miners transmitter to a section of the mine where the signal could be relayed up to the surface by some other means or the signal could be received by lowering a wideband receiver and loop. This would normally be done either at the shaft or a borehole, so the problem is one requiring a maximum sidelink propagation range from the miners area of entrapment to insure adequate signal detection at the vertical probe location. Figure 3-2 shows the magnetic field strength enhancement observed when the transmitting and receiving antennas are in close proximity (within 3 meters) to a major conducting pipe in the mine. This data was obtained at the Stauffer Chemical Trona Mine, Green River, Wyoming at a frequency of 1050 Hz. Figure 3-3, on the other hand, shows the measured sidelink data obtained in the Fletcher Mine, Viburnum, Missouri where proximity to local conductors was not a factor, except at distances greater than 500 meters from the transmitter.

The presence of pipes and other conductors occurring in most metal/non-metal mines can thus be utilized to advantage in relaying signals from a trapped miner to a shaft location for identification of vertical location of entrapment.

4.0 CONCLUSIONS

There are definite advantages to implementing a through the earth system for mine rescue communications and location. The most notable advantage is the simplicity and portability of the underground equipment required. For the deep metal/non metal mines of this study, excessive propagation losses can be minimized by (1) utilizing low frequencies in the 10-30 Hz range, and (2) utilizing long antennas for reception of signal on the surface. Vertical location determination can be obtained via sidelink coupling to the mine conductors and detection of signals at the shaft via a broadband receiver lowered down the shaft. Surface signal to noise ratios can be enhanced to the extent necessary by utilizing a digital processing receiving system and synchronous detection techniques.

Alternate methods of establishing trapped miner communications include:

UHF/VHF - Transceivers and Repeaters
MF - Transceivers and Repeaters

These techniques are being investigated by the Bureau of Mines and introduce a different set of advantages and disadvantages in solving the trapped miner communications problem. The main disadvantage in the above methods is the requirement for more extensive underground equipment. The main advantage is that voice communications are provided for both uplink and downlink and that the system can double as a routine communications link and therefore be ready for use in the event of emergencies.

Figure 3-2 Sidelink Propagation of ULF Horizontal Magnetic Fields, Stauffer Chemical Co.-Green River Wyo. Big Island Mine

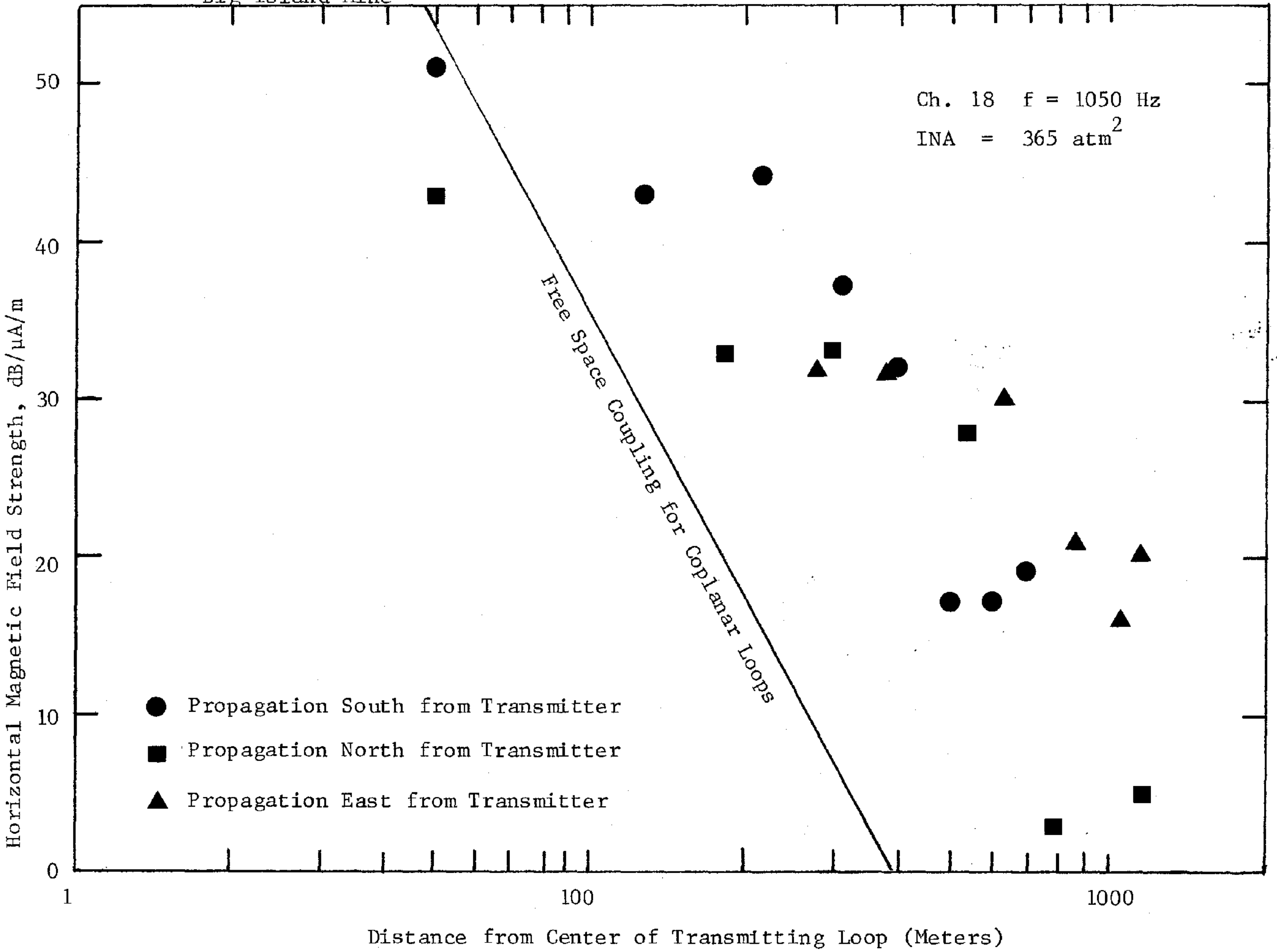
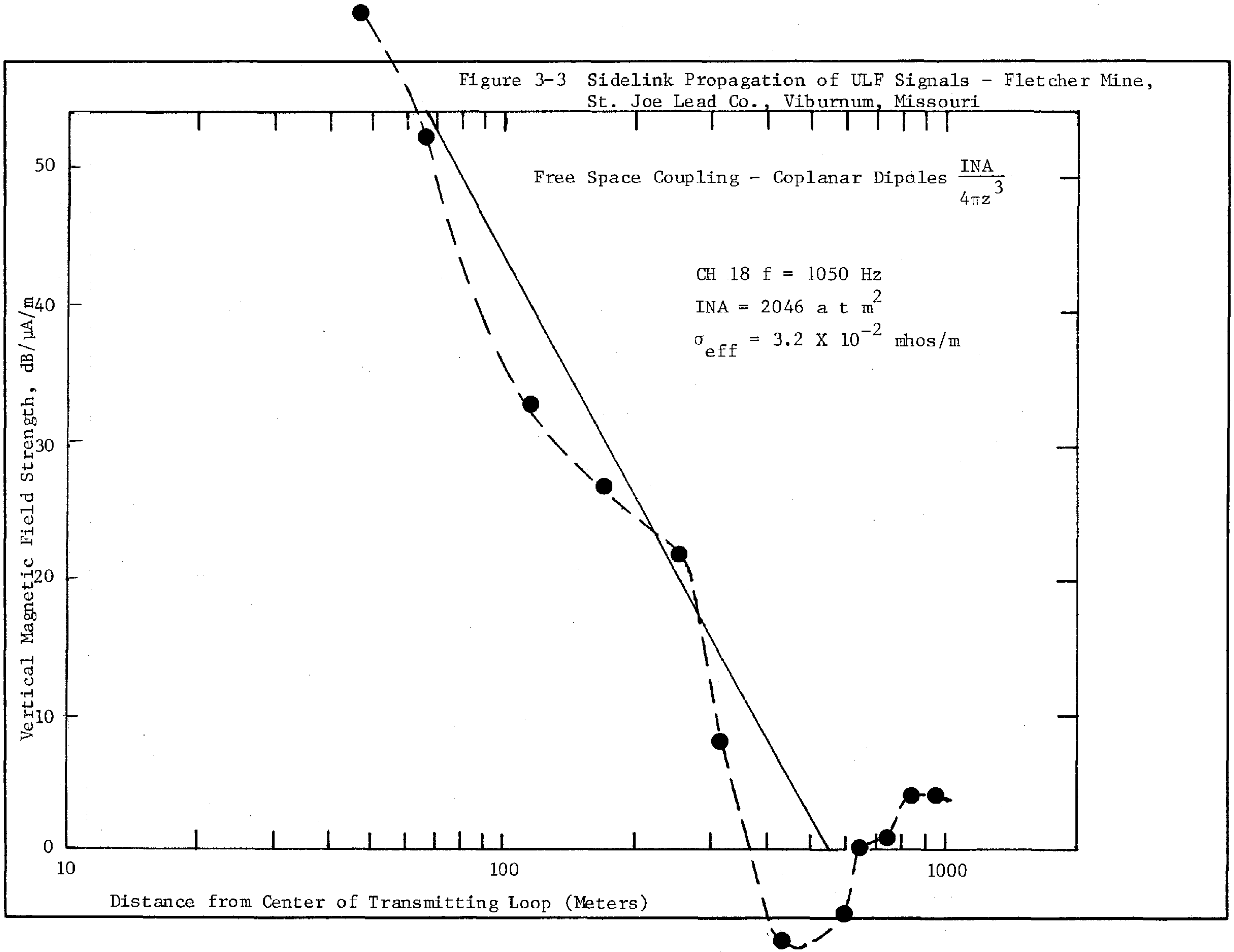


Figure 3-3 Sidelink Propagation of ULF Signals - Fletcher Mine, St. Joe Lead Co., Viburnum, Missouri



5.0 ACKNOWLEDGMENTS

The work described herein was conducted by Westinghouse Geophysical Instrumentation Systems under USBM Contract No. J0166100.

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QUESTIONS BY JOHN DURKIN

1. It is stated that substantial increase in signal sensitivity is obtained by using a horizontal mine antenna in place of a horizontal loop when receiving signals from submerged loop antennas. Does theory support this?

2. Is it advantageous to use horizontal wire antennas in transmitting voice to located mines?

3. Example worked out in paper showing a substantial decrease in needed magnetic moment when frequency is decreased is misleading. One must consider the increase in atmospheric noise, increase in receiver noise and necessary increase in receiver bandwidth.

REPLIES BY A.J. FARSTAD

1. The theory states that when the horizontal wire antenna is long compared to the distance away from it, the field strength in the medium,

$$H_y \approx \frac{I|c|}{2\pi Z}$$

where $|c|$ is the attenuation in the medium and Z is the penetration into the medium. By reciprocity, if current were induced into the receiving loop, then the mutual coupling would be such that the voltage received on the long wire antenna would also vary inversely with distance Z. In comparing the horizontal wire antenna with a horizontal loop as a receiving antenna, the respective signal sensitivities would depend to a great extent on the size of the horizontal loop. If the horizontal loop was large, it could be treated as four horizontal wire antennas connected together to form a loop giving a sensitivity comparable to that of the horizontal wire antenna. However, if the horizontal loop was small, it would be classified as a dipole and the $1/Z^3$ distance dependence would reduce its sensitivity substantially.

2. Whether it would be advantageous to use a horizontal wire antenna rather than a loop to transmit voice signals to an underground location would again depend upon the relative dimensions of the horizontal wire antenna and the horizontal loop. It may be advantageous from a logistic standpoint to deploy and retrieve a horizontal wire antenna as opposed to a loop. On the other hand, a loop has a lower ohmic resistance than a horizontal wire antenna and with proper tuning could be used to pass higher current than the horizontal wire.

3. See response to question below by H.K. Sacks.

COMMENT BY H.K. SACKS

I don't believe that the 10^7 improvement in performance is reasonable by shifting from 1000 to 10 Hz since receiver sensitivity and atmospheric noise are against you.

REPLY BY A.J. FARSTAD

The 10^7 improvement shown in the example case ($d = 915$ m, $\sigma = 1.15 \times 10^{-1}$mhos/m) only applies to the difference in propagation loss between $f = 1000$ Hz and $f = 10$ Hz. True, the receiver sensitivity and atmospheric noise become worse at 10 Hz. However, the atmospheric noise is the limiting factor* and it typically increases by about 20 dB** in going from 1000 Hz to 10 Hz, giving a net gain in signal to noise ratio of 10^6. If this is still not sufficient to extract the signal from the noise, digital processing receivers utilizing synchronous detection techniques can be used to further narrow the bandwidth and lift the signal out of the noise.

* Cryogenic magnetometers manufactured by SHE Corp., San Diego, specify sensitivities of $10^{-5}\gamma$ or .0796 μA/m at 10 Hz.

** Maxwell & Stone, "ELF and VLF Atmospheric Noise", Report No. 30P-6, Deco Electronics, 1964.

QUESTION BY G.A. COLLINS

What are your thoughts on digital processing receivers for VLF/ULF?

REPLY BY A.J. FARSTAD

In situations where the noise environment (either atmospheric or man-made) exceeds the received signal level in the bandwidth of the receiver, the signal can be extracted using cross correlation techniques in a digital processing receiver. Provided that the source frequency in the underground transmitter is stable, this same frequency can be synthesized in the digital receiver and by synchronous detection, be used to reduce the receiver bandwidth inversely proportional to integration time. The noise background can then be reduced approximately as the square root of the resulting bandwidth reduction without affecting the signal strength.

TRAPPED MINER THROUGH THE EARTH COMMUNICATIONS IN COAL MINES

Dennis Kalvels

Westinghouse Electric Corporation
Geophysical Instrumentation Systems
3655 Frontier Avenue
Boulder, Colorado

ABSTRACT

The U.S. Bureau of Mines has developed an electromagnetic location system for locating trapped miners in coal mines. This development program has led to the fabrication of a low power lightweight transceiver that can be attached to miner's belts, and a surface receiver and transmitter that can communicate with the underground transceivers. The test program now in progress has several aims.

1. Evaluate the performance of the trapped miner location devices in a variety of coal mines.

2. To establish a data base of transmission loss as a function of depth.

3. To determine whether uplink transmission loss characteristics could be determined from downlink measurements alone.

4. To determine effective conductivity, using magnetic field strength measurements.

Performance of the trapped miner location equipment has been demonstrated successfully in over three dozen mines ranging in overburden from 200 to 1400 feet. Discrete frequency tone signals (630 Hz, 1050 Hz, 1950 Hz, and 3030 Hz) pulsed at a very slow rate, are transmitted through the earth and measured in both uplink and downlink directions. Baseband voice is also transmitted and has been demonstrated through overburdens as thick as 700 feet.

A data base of transmission loss as a function of overburden depth is being compiled on this program. Eventually data from approximately 100 mines will be analyzed and extrapolated to a larger group of mines. Data obtained thus far indicate increased transmission loss with increasing frequency and with increased overburden; as can be expected.

However, exceptions were numerous, indicating that inhomogeneity and structural coupling are frequently present.

Measurements have shown that, in general, uplink transmission loss characteristics can not be extrapolated from downlink measurements because the small underground receiving antenna used in downlink measurements couples differently to mine conductors than the large area transmitting loop used by trapped miners for uplink transmission.

TRAPPED MINER THROUGH THE EARTH COMMUNICATIONS IN COAL MINES

An electromagnetic location system has been developed for locating trapped miners in coal mines. This system consists of a light-weight transmitter and receiver that can be attached to miners belts.

The transmitter is a compact unit, designed to attach to the top of a miners cap lamp battery for power. Each transmitter operates on one preselected discrete frequency tone pulsed at a slow rate. The frequencies used are from 510 Hz to 3030 Hz, spaced in 60 Hz increments between the 60 Hz power line harmonics. The transmitter is designed to be used with any available wire loop antenna. The nominal loop is standard #12 wire deployed around one or two coal pillars. This configuration results in a transmitting moment of up to 1000 amp t m^2.

The receiver is a larger multichannel unit approximately the size of a self rescuer. It also attaches to a miners belt and has its own battery pack inside. Each receiver has multichannel capability so that any transmitter can be received on one receiver. The receiver uses a 500 turn loop antenna fifteen (15) inches in diameter. With this mobile antenna, the receiver can be moved to locate the point of minimum horizontal field strength, and thus enable the user to fairly accurately locate the position of the transmitter.

Under the USBM, "Reliability and Effectiveness Analysis of Electromagnetic Location Systems," Contract #J0166060, the performance of four of these transmitters and several receivers is being evaluated in both uplink and downlink directions. In addition a high power transmitter, capable of producing a 20,000 Amp t m^2 transmitting moment, is being used to transmit downlink baseband voice. A voice receiver is used in the mine.

The U.S. Bureau of Mines contract titled, "Reliability and Effectiveness Analysis of Electromagnetic Location Systems," test program has several aims:

1. Evaluate the performance of the trapped miner location devices in a variety of coal mines.

2. To establish a data base of transmission loss as a function of depth.

3. To determine whether uplink transmission loss characteristics could be determined from downlink measurements alone.

4. To determine effective conductivity, using electromagnetic field strength measurements.

Through the earth communications has been achieved at over three dozen mines ranging in overburden from 200 to 1400 feet. Discrete frequency tone signals (630 Hz, 1050 Hz, 1950 Hz, and 3030 Hz) pulsed at a very slow rate, have been transmitted through the earth and measured in both uplink and downlink directions.

Downlink voice at baseband has been successfully used to a depth of about 700 feet. Using the voice downlink and pulsed tone uplink transmitters and receivers, a two way communications link has been set up between the mine and the surface.

Under emergency conditions, a miner would hopefully be equipped with a transceiver c onsisting of a voice receiver and a tone transmitter which he would display and initially send a location signal to the surface, through the earth. When detected at the surface and located, the surface rescue team could transmit down any voice instructions, questions, assurances, etc. to which the trapped miner could respond by simple pulse keying. This way a two way through the earth communications system could be established with the trapped miner.

Tone transmissions have been detected at depths of 1400 feet under low ambient noise conditions and through overburdens that exhibit low conductivities, i.e., .01 mhos/m. It is extremely complex and difficult to determine precisely the extent that structural anomalies such as track, cable, pipes, surface structures, etc. have on the strength of the received signal.

A data base of transmission loss as a function of overburden depth is being compiled in this test program. Figures 1 and 2 show the transmission losses encountered for the uplink and downlink tests conducted so far at over three dozen mines. The curves included indicate the expected loss that the electromagnetic field would experience in a homogeneous overburden having an effective conductivity of .01 mhos/m. The data indicates greater losses as though the effective conductivity was as high as .1 mhos/m.

Losses at the shallower depths appear especially high. At shallow depths the dipole approximation of the transmitting antenna used in making vertical magnetic field strength measurements become less valid. If the geometry is to be accounted for, the strength of the calculated field would decrease.

A further aim of this test program is to determine whether uplink transmission loss characteristics could be determined from downlink measurements alone. In using a large transmitting loop with an enclosed area of up to 1500 m^2 and a receiving loop of less than 1 m^2, some coupling differences may occur if the locations of the two antennas were reversed. It has been found that under conditions where the overburden has been homogeneous and free of local conductors, the uplink and downlink transmission losses compare quite closely. However, as the size and number of local conductors increase, the differences in uplink and downlink losses do also. It appears that local conductors provide different coupling to an antenna of smaller area than an antenna of larger area giving different uplink and downlink results. In comparing the data from a larger group of mines however, the differences tend to "cancel" out permitting in general a comparison between uplink and downlink transmission loss characteristics. This can be seen in Figures 3 and 4.

Effective conductivity determinations have been made from the vertical magnetic field strength measurements. Calculations result in effective conductivities between .01 and 1 mho/meter. In general, the conductivity of the overburden in coal regions is expected to average about .01 mhos/meter. Figures 4 and 5 show the downlink transmission loss data superimposed with expected loss curves for two conductivities (.01 mhos/m and .05 mhos/m). The .05 mhos/m conductivity curves appear more indicative of the losses found in actual field measurement, except for overburdens less than 300 feet.

Figure 6 graphically describes the difficulty in making effective conductivity determinations with the parameters of depth, frequency, and transmitting moment presently used. The difficulty lies in making an effective conductivity determination at shallow depths, (less than 100 m), and at lower frequencies (less than 1000 Hz) due to the lack of divergence of the conductivity curves. At 50m overburden, the conductivity curves do not diverge sufficiently to allow a conductivity determination unless the value of H_z is very accurate. Since accuracy obtained is 1 dB, accurate determinations can only be made at deeper overburdens. Determinations of effective conductivity from H_z data at depths less than 100 m are highly suspect especially at lower frequencies.

ACKNOWLEDGMENTS

The work described herein is being conducted by Westinghouse Geophysical Instrumentation Systems, Boulder, Colorado and is sponsored by the U.S. Bureau of Mines under Contract J0166060.

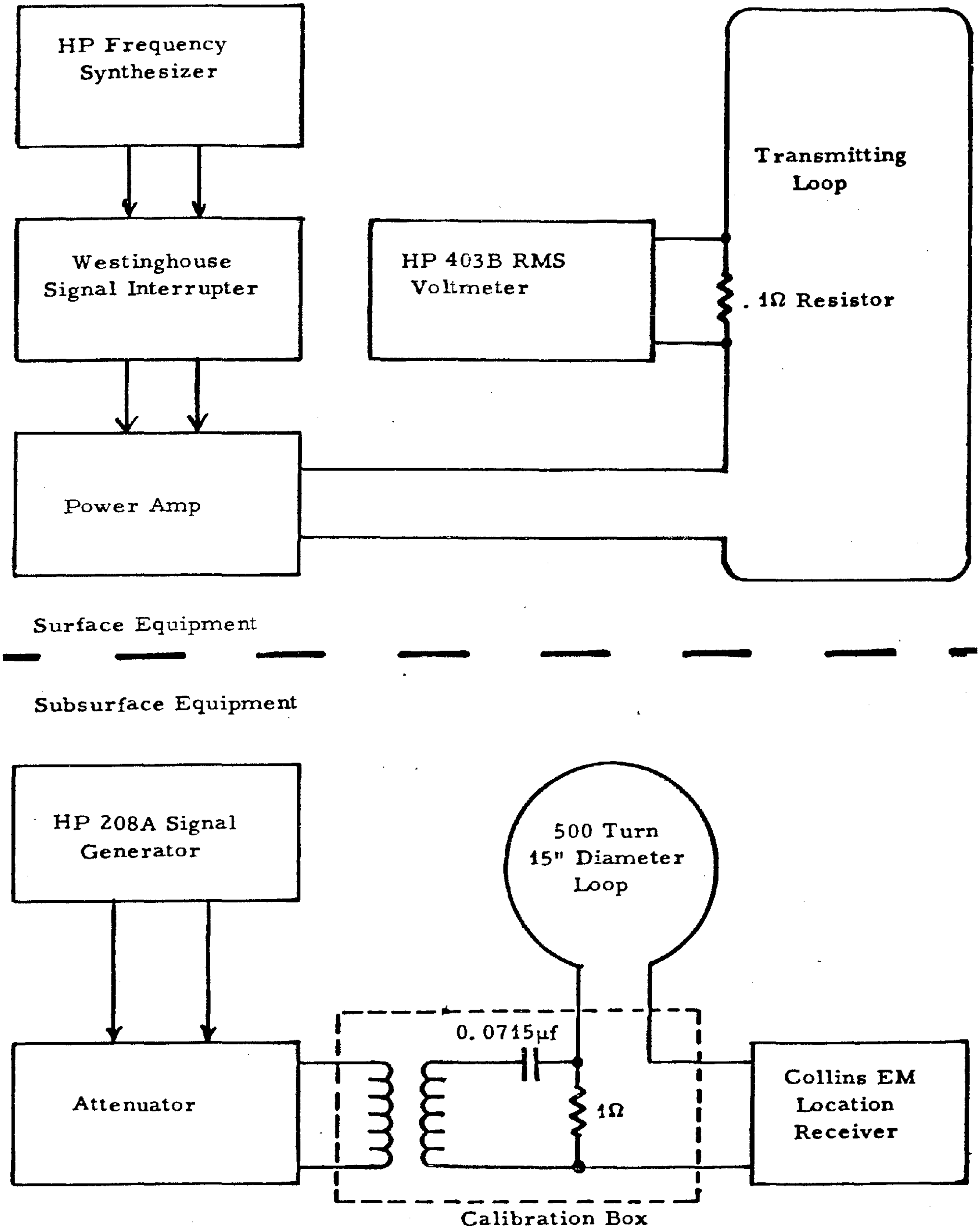


Figure 1 Downlink Test Set Up

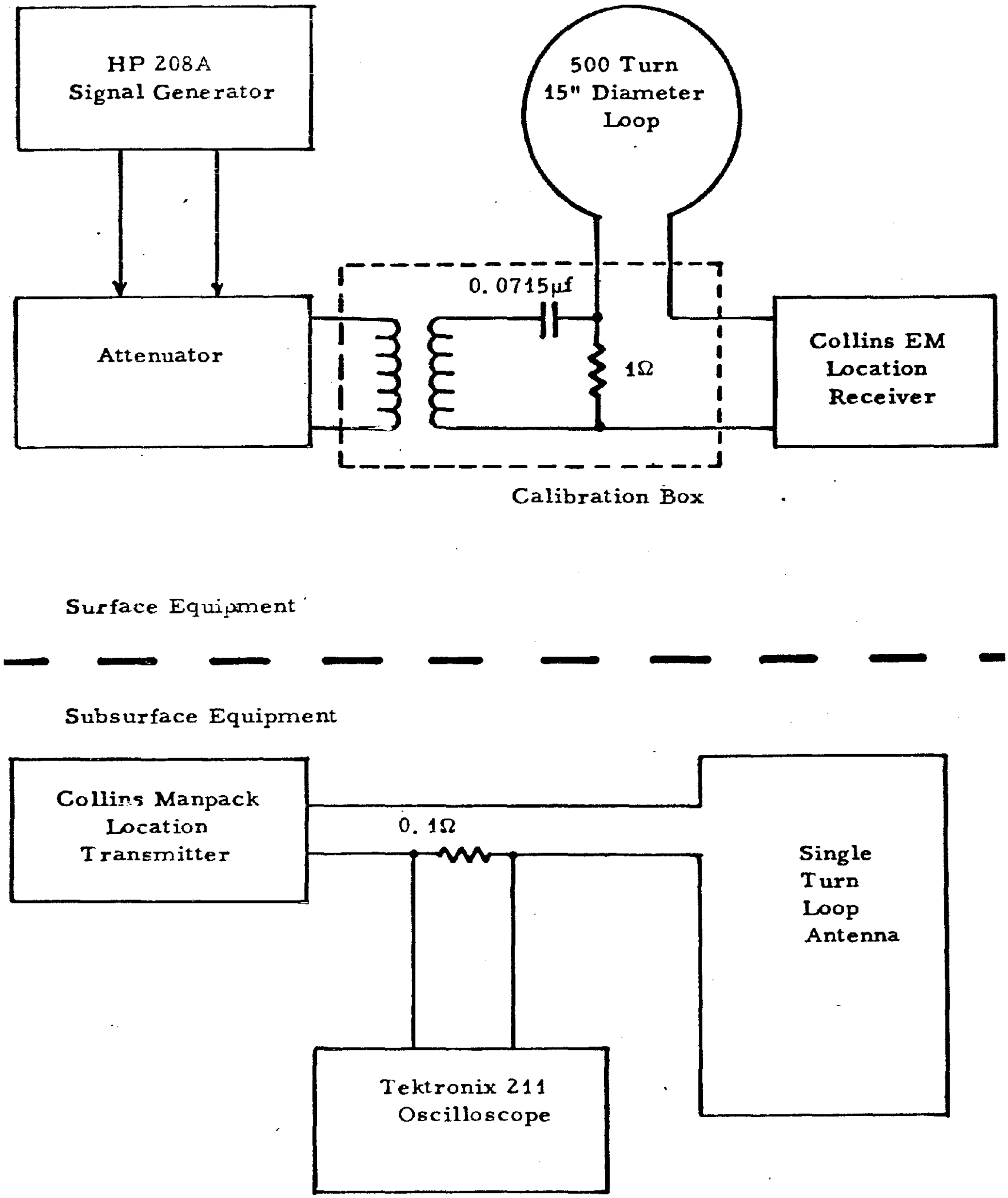


Figure 2 Uplink Test Set Up

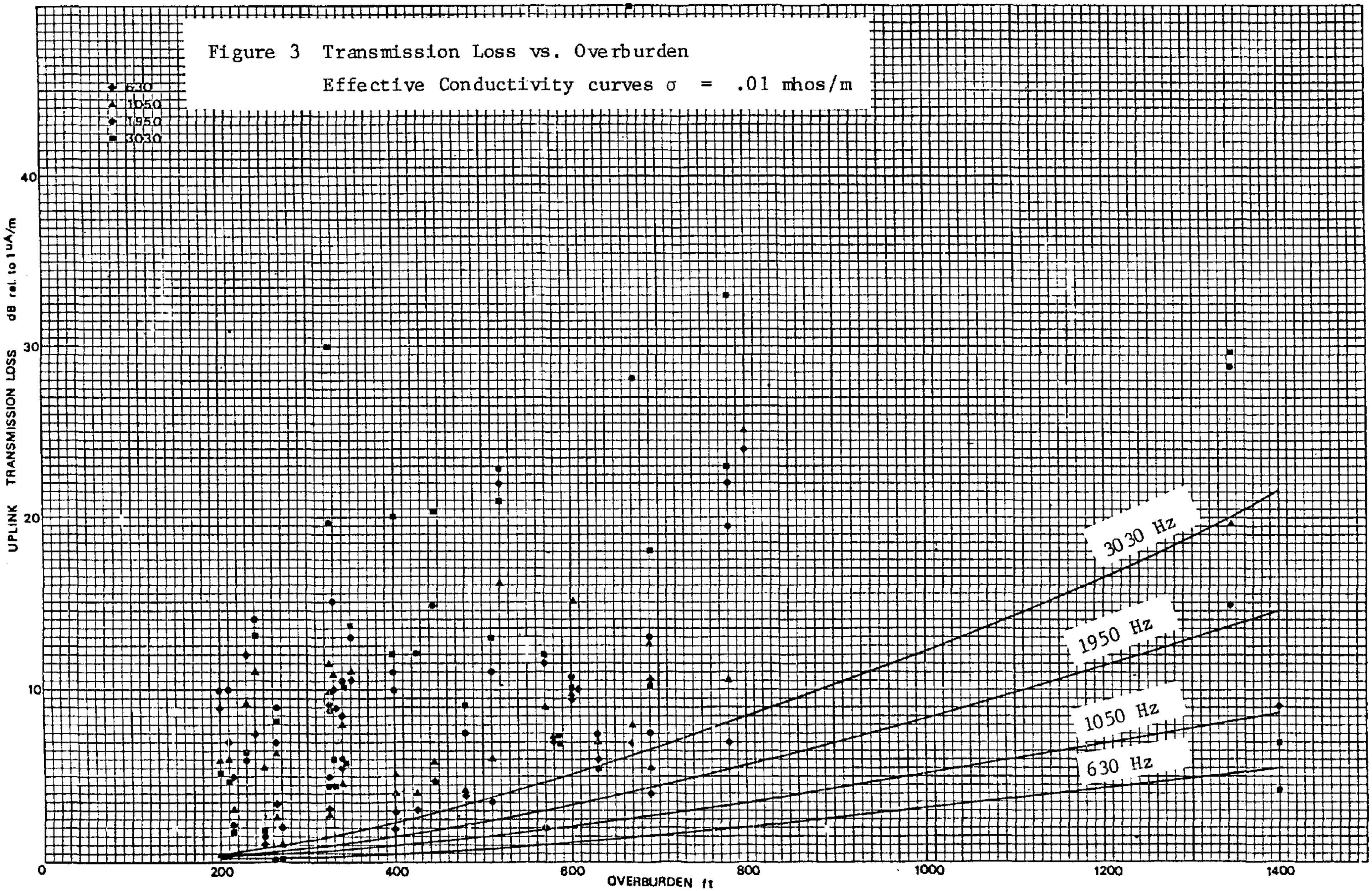


Figure 3 Transmission Loss vs. Overburden
Effective Conductivity curves σ = .01 mhos/m

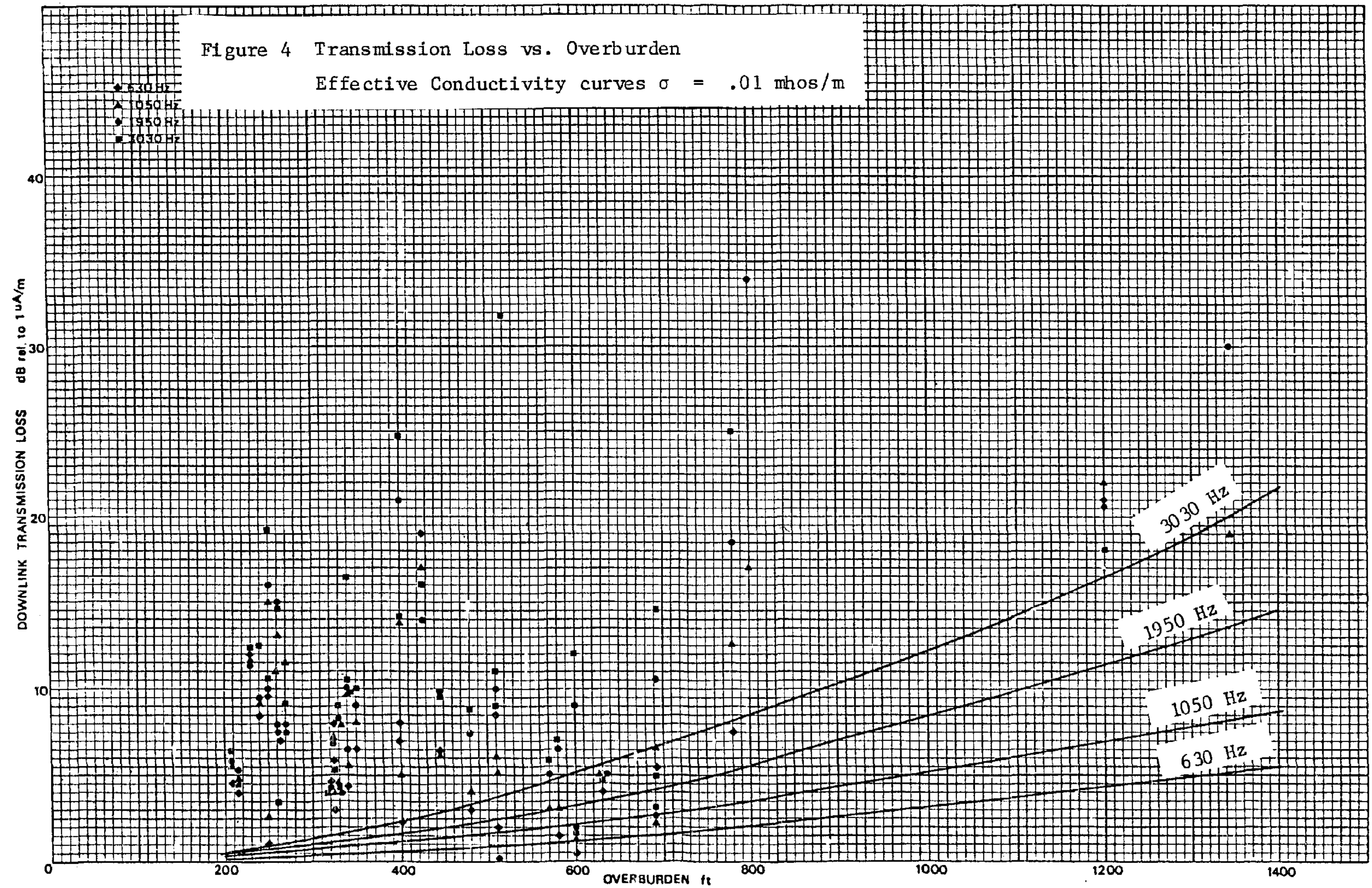


Figure 4 Transmission Loss vs. Overburden

Effective Conductivity curves σ = .01 mhos/m

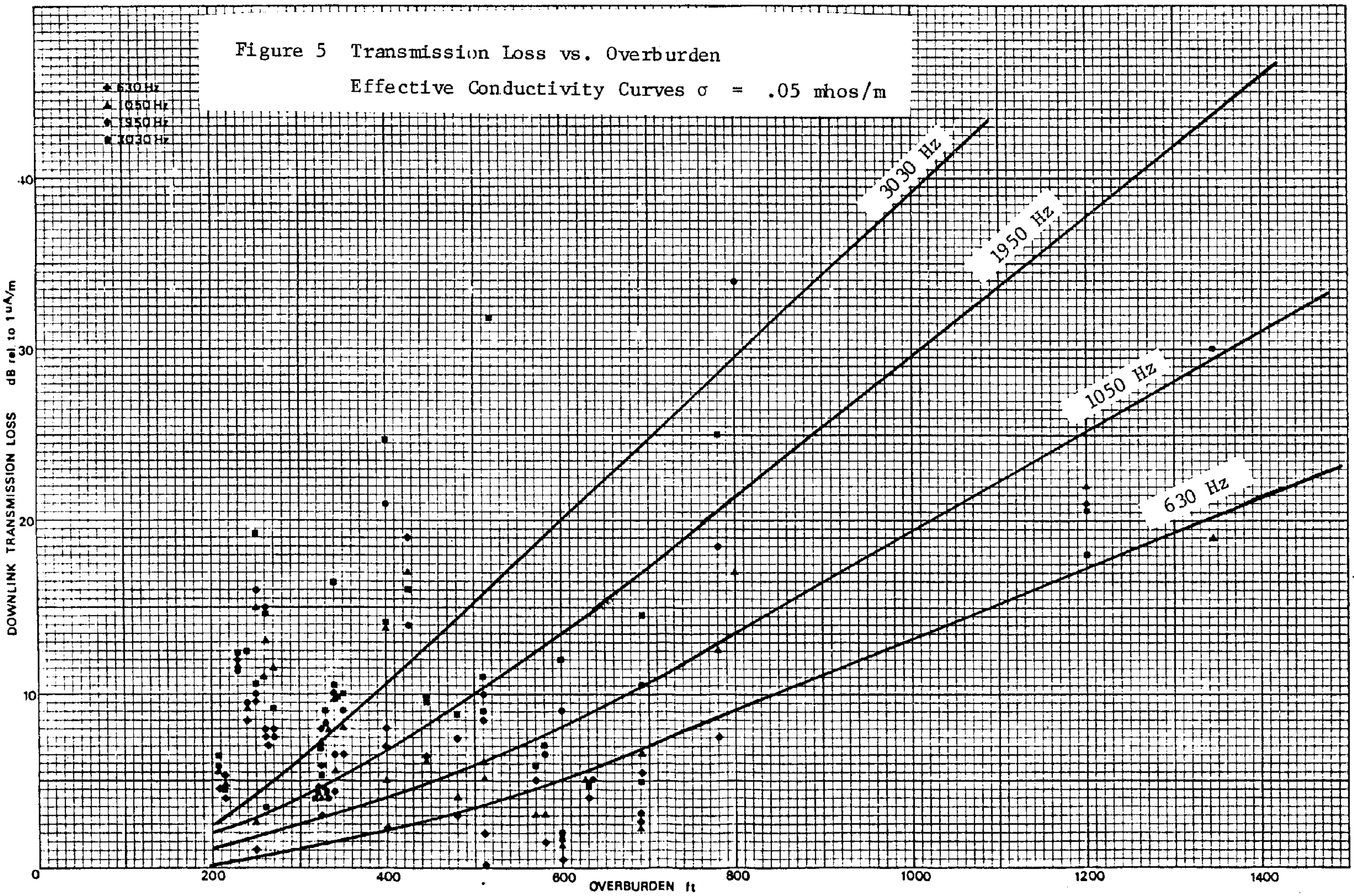
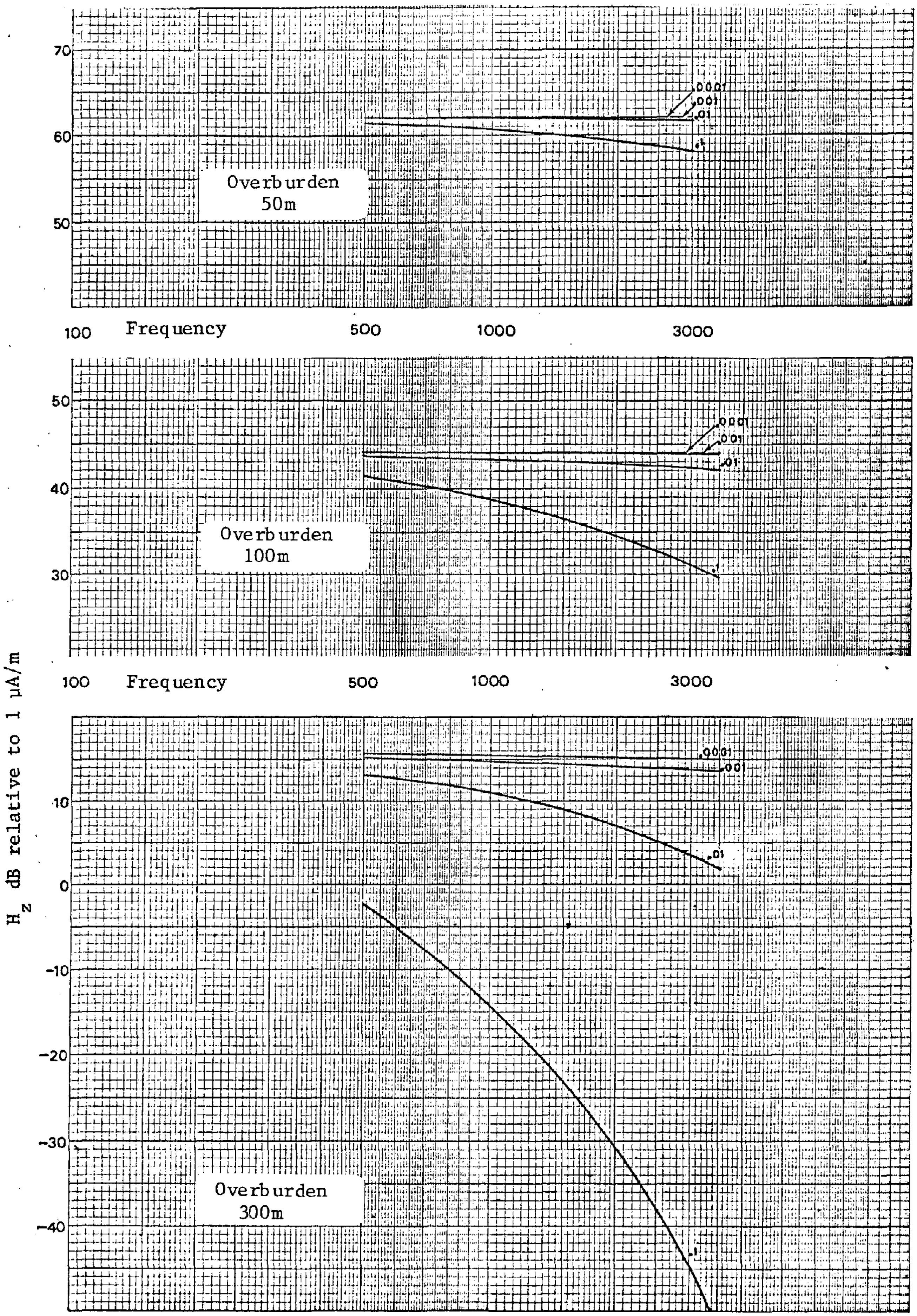


Figure 5 Transmission Loss vs. Overburden

Effective Conductivity Curves σ = .05 mhos/m

Figure 6 Effective Conductivity vs. Depth INA = 1000 Amp t m^2



QUESTION BY LAGACE AND WAIT

How were the conductivities measured and at what frequency, etc.?

REPLY BY A.J. FARSTAD
(metal non-metal mines)

Conductivity measurements were made at the metal non-metal mines using conventional 4-probe electrode arrays both from the surface and in the mine. The spacing of the arrays were expanded in a series of increasing separation to result in an effective conductivity sounding vs depth. Frequencies on the order of 1 Hz were used in the conductivity measurements to eliminate effects of inductive coupling.

REPLY BY D. KALVELS
(coal mines)

Effective conductivities were derived from the through the earth propagation data. From the transmission loss measurements, a value of conductivity that would produce that loss relative to the free space coupling was calculated. This technique becomes inaccurate when the loop dimensions are comparable to mine depth and can no longer be considered dipoles. Also, in mines exhibiting very low transmission loss at low frequencies, it is difficult to determine conductivity by this method.

QUESTION BY T.S. CORY

On the uplink, was the vertical field always the maximum field strength?

REPLY BY D. KALVELS

The maximum field was always measured, which most of the time was vertical. Occasionally distortion caused departure from vertical. In one case, the maximum appeared to be horizontal. The receiving antenna was always oriented for maximum coupling.

U.S. DEPARTMENT OF COMMERCE
Office of Telecommunications
INSTITUTE FOR TELECOMMUNICATION SCIENCES
Boulder, Colorado 80302
Room 242, RB 1

15 November 1977

Dr. H. Kenneth Sacks
Pittsburgh Mining & Safety Research Center
U.S. Bureau of Mines
4800 Forbes Avenue
Pittsburgh, PA 15213

(This letter is relevant to the previous two papers and the ensuing discussions)

Dear Ken:

Dave Hill relayed your request to furnish some quantitative information on the possible non-dipole aspect of Arnie Farstad's results for shallow depth transmission through the overburden. I had already discussed this problem with Arnie and he is aware of our earlier study of this problem.

The enclosed reprint is based on the surface to sub-surface transmission but the problem is reciprocal if you talk about the mutual impedance between a horizontal loop of finite size and a horizontal loop of infinitesimal size (i.e. both loops have vertical axes).

It is true that the neglect of the dipole approximation could lead to an apparent conductivity of anomalously high value. For the coaxial geometry the reduction of the normalized field is approximately $(1 + A^2)^{-3/2}$ where

$$A = \frac{a}{h} = \frac{\text{loop radius}}{\text{depth}}$$

This is valid for H < about 0.5. The situation is discussed in the attached reprint. There is a more detailed report (dated 15 March 1972) with numerical print-outs for the parameters indicated on the curves. To generate more numbers would probably require writing a new computer program. This could be done if you think there is a need.*

As Arnie Farstad points out, the conductivity determination from shallow depth/low frequency measurements are very rough. Such values are subject to all sorts of error due to geometrical mis-alignments and local conductors. The extrapolation of such conductivity data to higher frequencies and/or greater depths is really very questionable.

**Such results were communicated to John Durkin in early March.*

Best regards,

James R. Wait
Consultant to ITS/OT

P.S. I will be sending the Workshop abstracts within the next few days.

Enc: Rp 472: "Subsurface electromagnetic fields of a circular loop of current located above ground" by J.R. Wait and K.P. Spies, IEEE Trans., Vol. AP-20, No. 4, 520-522, July 1972.

cc: D.A. Hill
A.J. Farstad

A RADIO EQUIPMENT FOR MINE COMMUNICATION AND CONTROL.

Prof. P. Delogne (*)

Efficient radio communication in the mine requires the use of some type of leaky coaxial cable. The attenuation curves of coaxial cables suggest to use a rather low frequency. However the coupling of the mobile antenna to the monofilar mode improves with frequency. Electrical noise may also be important at low frequency. A compromise between these adverse effects lead to the conclusion that the band 5-10 MHz yields maximum range for mobile-to-mobile and fixed-to-mobile communications.

In this band the attenuation of good quality coaxial cables is as low as 0.5 dB/100 m and it is extremely useful to optimize the mobile transceiver performance, for every dB improvement yields 200 m range increase. The antenna is very short compared to the wavelength and it thus reduces to an electric or magnetic dipole. Careful evaluation including circuit problems lead to the choice of a ferrite antenna. The use of four ferrite rods brings 6 dB improvement w.r.t. a single rod.

It was equally important to optimize the modulation system. A comparative analysis showed that wideband phase modulation was the best solution provided it is combined with a threshold reduction technique. A 35 kHz deviation with a phase-locked demodulator yields good output quality for a carrier-to-noise ratio of 0 dB. This is a 7 dB gain compared to narrowband FM.

The mobile transceiver which has been developped according to these principles uses thick film integrable techniques. It will provide 10 frequency channels with ranges of 10 km for fixed-to-mobile and 5 km for mobile-to-mobile communications.

Finally it was interesting to consider operational aspects. As an option the base station may be equipped with a selective calling system requiring only two integrated circuits. The same system can be mounted in the mobile transceivers to provide simultaneously voice communication and remote control and signalling. The system includes a call memory and an automatic check of the channel status.

* Université Catholique de Louvain, consultant to INIEX, Belgium.

(This is the abstract for the following paper)

A RADIO EQUIPMENT FOR MINE COMMUNICATIONS AND CONTROL

Prof. P. DELOGNE, A.M. ANCKAERT & D. LEONARD
Catholic University of Louvain, Louvain-la-Neuve, Belgium

1. INTRODUCTION

The propagation of electromagnetic waves in underground works is now well understood and it appears that the largest linear range is obtained in the low-HF band [1]. However, this band is not used for mobile communications on the surface and no radio equipment is available. Special transceivers and base stations have to be developped for the mine. The only known equipment of this type is the CERCHAR X-Y-phone system operating at 7 MHz.

At these frequencies the specific attenuation of a good quality coaxial cable is as low as 0.5 dB/100 m and it is extremely useful to optimize the transceiver performance, for every dB improvement yields a 200 m range increase. Key points are the design of the antenna and the choice of the modulation system. This leads to RF bandwidths of the order of 100 kHz.

Operational aspects are also important. It is believed that the availability of high-quality and moderate-cost equipment will favour steadily increasing use of radio communications in the mine; it should not be surprising that in the future 10 or more radio networks will be operated simultaneously in a same mine. It is thus desirable that new radio equipment provide at least 10 frequency channels; for safety reasons, channel switching must not be accessible to the user but it should easily be realized by the mine electrical technicians. Selective call should be offered as an option. Some remote control facilities should also be included. All these aspects have been discussed with users before INIEX (Institut National des Industries Extractives, Liége, Belgium) committed the Communications Laboratory of Louvain University with the development of a radio equipment for mine communication and control. This work is now near completion.

2. CHOICE OF FREQUENCY AND ANTENNA

2.1. General

Propagation aspects [1], required channel bandwidths and number of frequency channels impose to spread the channels over about 5 MHz in low-HF band. It was also decided that the antenna had to be incorporated into the portable radio transceiver. It is thus very small compared with the wavelength (less than 0.005). It is well known that electrically small antennas in free space reduce to electric or magnetic dipoles; this is not fundamentally changed in a tunnel, though the radiation resistance is modified. The radiation of an antenna in a tunnel at these low frequencies can also be viewed as a coupling to the monofilar mode.

2.2. Radiation resistance of an antenna in a tunnel

In the first instance, it is useful to relate the antenna parameters in a tunnel containing a wire conductor to their values in free space. We start from the resonnable assumption that the antenna height h_e (ratio of the electromotive force to the incident electric field) and also its reactance X are not modified by the tunnel. The tunnel may be taken to be the reference tunnel used in [1]. Next, we consider two experiments shown on Fig. 1 in which the antenna is used for receiving and transmitting the monofilar mode; it is assumed that the monofilar mode is terminated on its characteristic impedance Z_0 and, as in [1], that $Z_0 = \eta = 377\ \Omega$. According to the reciprocity theorem, one has

$$V_1/I_1 = V_{m2}/I_2 \tag{1}$$

In experiment (a) the electric field incident on the antenna can be related to the monofilar mode power by eq.(2) of [1] and thus to I_1. Equation (1) allows to calculated the voltage V_{m2} and the power $P_{m2} = V_{m2}^2/\eta$ of the monofilar mode excited in both directions in the transmitting experiment (b). Finally, we obtain the radiation resistance R_{rt} of the antenna in the reference tunnel (taking into account that the radiated power is 2 P_{m2})

$$R_{rt} = 3770\ h_e^2 \tag{2}$$

The radiation resistance of an antenna in free space is related to the electric height by

$$R_r = \frac{\pi \eta h_e^2}{\lambda^2 D} \tag{3}$$

where D is the directivity gain (3/2 for a Hertz dipole). This yields the practical formula for the reference tunnel

$$\frac{R_{rt}}{R_r} \simeq \left(\frac{\lambda}{5}\right)^2 \tag{4}$$

Though this ratio is well above one in the low-HF band, R_{rt} will any way be extremely small for antennas which do not exceed 0.005 λ.

2.3. Antenna optimization

The interest of formula (4) is that it reduces the problem to that of an antenna in free space. Many theoretical studies have been published over electrically small antennas in free space but in this case the antenna dimensions are so minute that a realistic approach was needed. The equivalent circuit of an electrically small antenna comprises in series a reactance X, the radiation resistance R_r and a loss resistance R_ℓ. We define the "radiation Q-factor" by the ratio of the reactive power W stored in the antenna to the radiated power P_r. It is given by

$$Q_r = \frac{W}{P_r} = \frac{|X|}{R_r} \tag{5}$$

As a start point, we can consider that the reactive power is fixed by the battery power supply and circuit considerations. The objective is thus to minimize Q_r. It should be stressed that, given W, the actual Q-factor

$$Q = \frac{|X|}{R_r + R_\ell} \simeq \frac{|X|}{R_\ell} \tag{6}$$

bears no relation with the radiated power. In fact, a high Q-factor allows in practice to increase W, but Q will be limited by bandwidth considerations.

CHU |2| and COLLIN & ROTSCHILD have established a lower limit of the radiation Q-factor of an antenna which is bounded at the interior of a sphere with radius a.

$$Q_r > \left(\frac{\lambda}{2\pi a}\right)^3 \tag{7}$$

This is a lower limit because they consider only the reactive power stored outside the sphere. In the present case, this lower limit is extremely high : 750000 at 7 MHz for 2 a = 15 cm;

A careful examination showed that the electric and magnetic Hertz dipole could both have Q_r-factors one order of magnitude above the limit (7) on certain conditions. However the reactance of very short electric dipoles is so high that it should not be possible to develop a significant reactive power into them. The choice was thus limited to magnetic Hertz dipoles, i.e. multiturn loop and ferrite antennas. Given maximum dimensions of 4 x 6 x 20 cm, it appeared that a well-designed ferrite antenna was 3 dB better than the best loop antenna.

Formulas for the accurate design of ferrite antennas are available in the literature [4]. The self-inductance and the radiation resistance are both proportional to the square of the number of turns and Q_r is thus independent of this parameter; as a result, the reactance can be chosen at the most suitable value for circuit problems. A careful selection of the rod dimensions and material permeability brought the Q_r factor to the value 8.10^6 at 7 MHz, i.e. 10 dB above the theoretical limit. Finally, it was observed that the coupling coefficient of two parallel ferrite rods remained very small even when the spacing was as low as 2 cm; by using 4 parallel ferrite rods connected in series, the reactance is multiplied by a factor about 4, but the magnetic moment is multiplied by 4 and thus the radiation resistance by 16. As a result, the Q_r-factor is reduced by a factor 4. With this 6 dB improvement it is now 4 dB above the theoretical limit (7).

The antenna was designed this way at the frequency of 7 MHz, but it remains optimum at other frequencies. Indeed the radiation resistance R_r of a ferrite antenna in free space varies as f^4, so that Q_r varies as f^{-3} as well as the theoretical limit (7). However, formula (4) shows that the radiation resistance in the tunnel varies as f^2. This appears also from (2) where the effective height is given by

$$h_e = \mu_e \frac{2\pi S n}{\lambda} \qquad (8)$$

where μ_e is the effective permeability, S the cross-section of the rod and n the number of turns. The magnetic moment of the antenna used for transmission is $\mu_e nSI$; combining this with (2), we obtain formula (6) of [1]. Finally, when the antenna is used for receiving the monofilar mode in the reference tunnel, using (2) of [1], the antenna e.m.f. is given by $h_e P_m^{1/2}$.

3. CHOICE OF THE MODULATION SYSTEM

A definite advantage in developing a specific system for radiocommunications for the mine is that the channel bandwidth is not limited by existing surface regulations. Wideband frequency modulation is known to provide a signal-to-noise improvement over amplitude modulation. When using a classical discriminator, this advantage is gained only over the threshold, which corresponds to a C/N ratio of about 12 dB. However, as shown on Fig. 2, the threshold can be reduced using a phase-locked loop in place of a discriminator [5]. With a phase-locked loop an output S/N ratio of 30 dB can be obtained with a modulation index m = 10 , i.e. a frequency deviation of 35 kHz and an RF bandwidth of 80 kHz. This yields a 7 dB improvement compared with narrow-band FM (m = 3) with a discriminator and 18 dB compared with SSB. The output S/N ratio of

30 dB corresponds to a C/N ratio of 2 dB.

A digital technique is not interesting. Indeed, digital encoding of the voice will yields bitrates of 40 kbit/s or more. A subsequent phase or frequency modulation would require the same RF-bandwidth as the wideband FM described above but a C/N ratio of about 12 dB. Finally, the wideband FM with a PLL demodulator was retained.

4. BLOC-DIAGRAMS

Fig. 3 shows a simplified bloc-diagram of the portable transceiver. Frequencies are given for a channel at 7 MHz. The structure of the receiver is classical; it uses two intermediate frequencies at 10.7 MHz and 450 kHz where a CMOS integrated PLL is used. The image frequency is very high and lies in a frequency band 15-20 MHz which is not used for mine communications; the image selectivity is thus well ensured by the tuned ferrite antenna. The channel selectivity is obtained by two ceramic filters in the first IF. Much care was taken to reduce cross-modulation in the input stages. This was necessary because the mobile receiver may work very close to the cable in the immediate vicinity of a base station transmitting on the neighbouring channel.

In the transmitter, the antenna is used as the tuned circuit of a voltage control oscillator and a phase-locked loop with a frequency division by N = 256 locks the mean frequency on that of a crystal oscillator; this provides an excellent frequency stability with the advantage of an automatic fine tuning of the antenna circuit. This arrangement is also such that no noticeable spurious are radiated in the band 5-10 MHz occupied by the channels. Harmonics of the operating frequencies ly out of this band. A differential microphone is used for protection against the high acoustic noise levels existing in the mine.

Channel selection is obtained by changing all crystals and tuning the antenna in the receive mode.

Similar structures are used for the RF/IF part of the base station but with some differences, mainly in the transmitter. The channel selection is obtained by varying one of the frequency divisors and the VCO locks on the channel frequency anywhere in the band 5-10 MHz. It is followed by a power amplifier. Of course the antenna is replaced by connectors for monofilar wire conductor, bifilar line and coaxial cable. Several base stations operated on different frequency channels may be connected to the same leaky cable with isolations between them as low as 15 dB; particular attention was paid to intermodulation problems and it was necessary to include a channel filter at the RF output of the base station.

5. PERFORMANCE

It has been verified on the prototype that the expected ouput S/N ratio of 30 dB was obtained for a C/N ratio of 0 dB. It was thus possible to obtain the following sensitivities :

- portable receiver : the minimum required magnetic field is 0.02 µA/m
- base station : 0.2 µV/75 Ω , i.e. - 153 dBW.

These sensitivities will be maintained if it appears that industrial electrical noise is not significantly higher than thermal noise, which may be true between 5 and 10 MHz. Regarding transmission, the base station power is 2 W and the portable equipment is equivalent to a magnetic moment of 27 mA.m^2.

This corresponds to an improvement of roughly 15 dB w.r.t. the French X-phone (7 dB on the sensitivity and 8 dB on the magnetic moment). As a result, expected ranges using a good quality coaxial cable with mode converters are as follows :

- base station to mobile : 8500 m
- mobile to base station : 7000 m
- mobile to mobile : 3000 m

assuming a realistic position of the cable in the tunnel. Finally, it is important to note that more than 10 frequency channels will be available and can be used on the same cable.

6. OPERATIONAL ASPECTS

6.1. Structure of networks

The leaky coaxial cable can of course be established in a star network covering all important tunnels as well as long wall faces. Remote control (for instance from the surface) of the base stations is frequently required in the mine; for this reason the radio part and the low frequency part of the base station have been separated into two cases that must be interconnected by a telephone line. Several low-frequency parts may be connected to the same line to control one base station; they all hear the traffic on this radio network but only one of them can transmit at a time. Each low-frequency and radio case contains a balancing battery which may be powered locally from the mains supply or through the telephone line by the other cases.

6.2. Calling system

In the basic option, a general call in the network is obtained by modulating the carrier at the maximum deviation with a tone frequency. This tone is taken in the frequency dividers of the transmitter phase-locked loop. The same general call is used by the base station and by the mobile transceiver. However, at the end of a call emanating from a mobile transceiver, the base station automatically sends a tone back during 100 ms. This provides to the mobile a check of the channel status.

As an option, selective calling of the mobile transceivers by the base station is possible. The code consists in the succession of three different tone frequencies among six and provides up to 150 different calls. In this system, the same MOS-integrated circuit can be used as encoder or decoder, but not simultaneously. Two such circuits are implemented in the mobile transceiver and when it is called, it will automatically repeat its own code and give this way a check of reception to the base station. The selective call is also memorized in the mobile by turning on a LED and maintaining a permanent tone.

A further option provides the mobile transceivers with radio control possibilities. Indeed, if it is equipped with two selective call MOS-integrated circuits, it now becomes able to transmit orders to the base station and receive from the latter a check of correct reception. Due to limited place for the push-buttons, the number of orders emanating from a mobile transceiver is limited to 6 . This option will provide the miner with voice communications and remote control with a single transceiver.

7. CONCLUSIONS

It is believed that this equipment will solve many radio communication and remote control problems that are encountered in the mine.

The equipment which has been developped offers wide possibilities. A small prototype series will now be realized and operated for a few months in a mine in order to check the realiability in the mine environment and also the usefulness of some operational choices. It will then be made commercially available.

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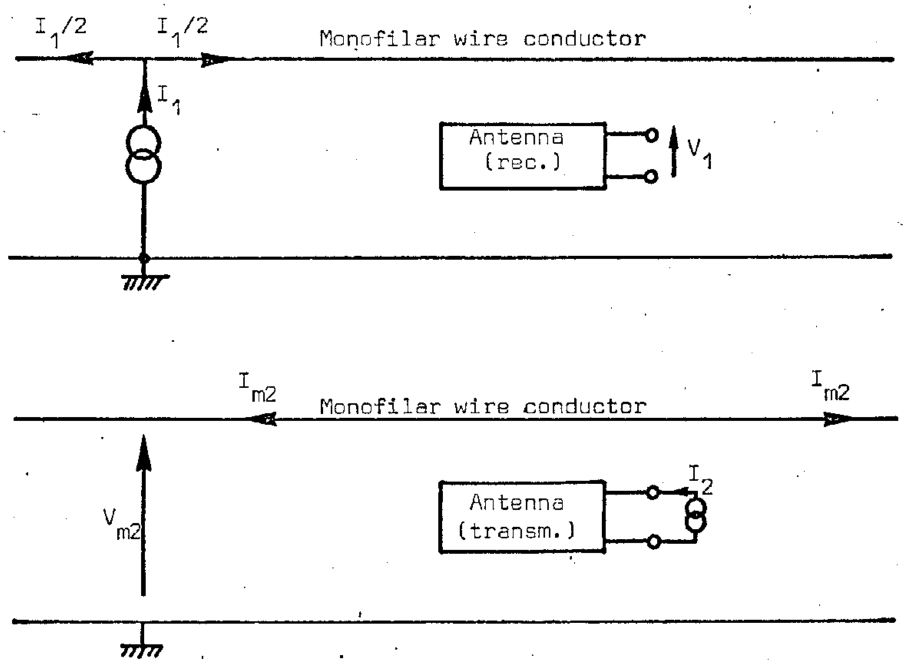


Fig. 1. Two experiments for the application of the reciprocity principle.

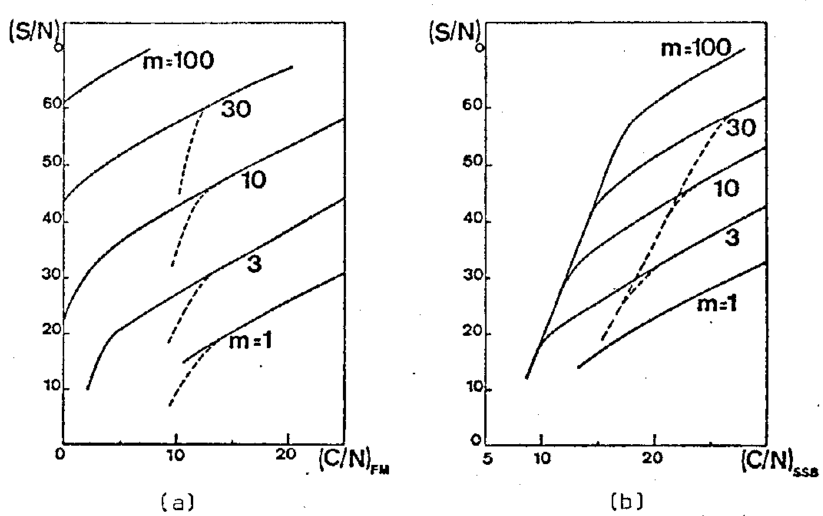


Fig. 2. Performance characteristics of FM.
(a) Abscissa is the C/N ratio in FM bandwidth
(b) Abscissa is the C/N ratio in SSB bandwidth
Continuous lines for discriminator. Dotted lines for PLL.

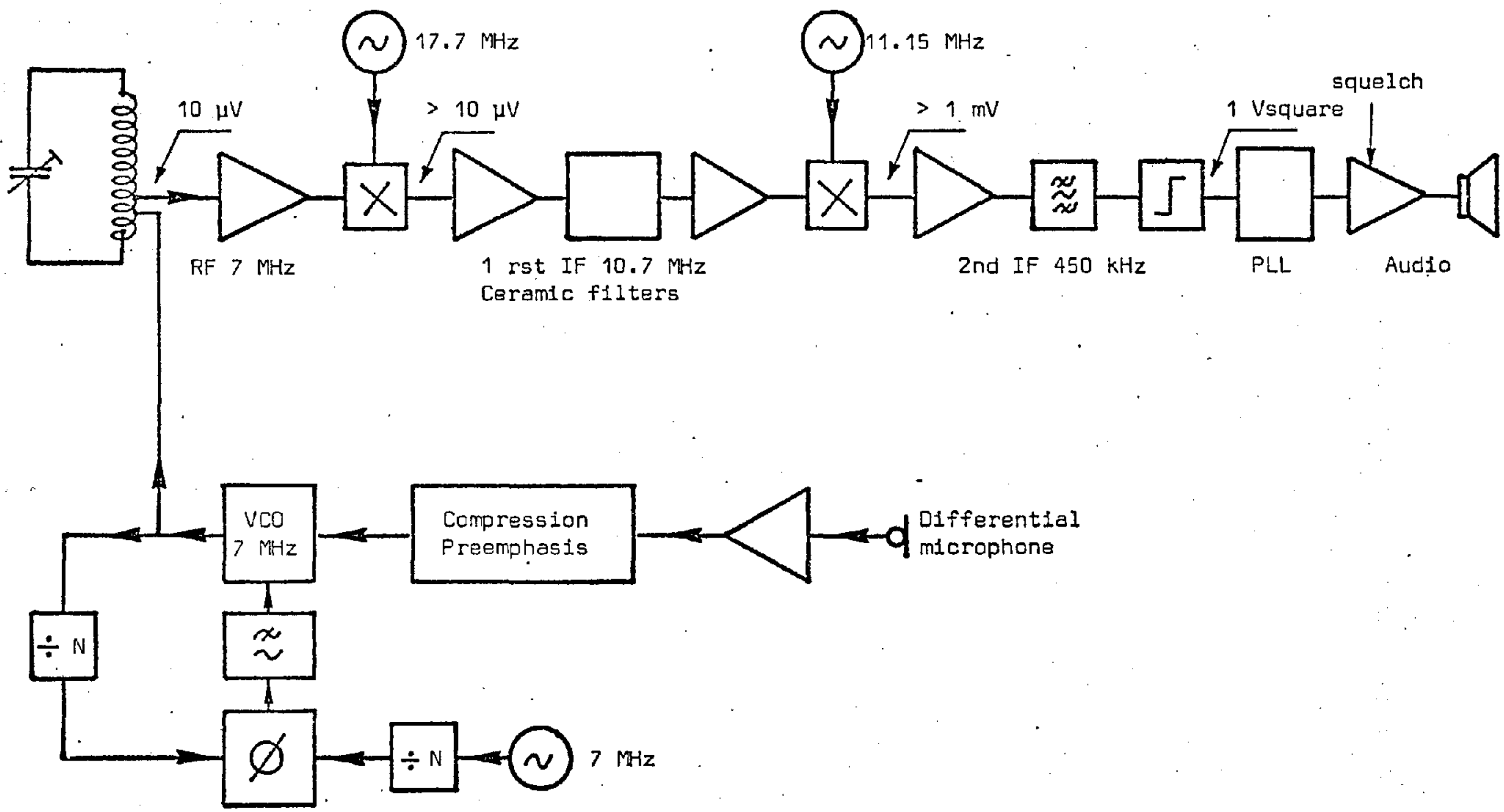


Fig. 3. Bloc diagram of the mobile transceiver.

COMMENT BY D.A. HILL

Our calculations for the ratio of the radiation resistance of a dipole in a tunnel to that in free space agree with your λ^2 dependence in the quasi-static regime. However, we also have a complicated dependence on antenna position and orientation which does not appear in your expression. Does your expression represent a special case?

REPLY BY P. DELOGNE

Yes, the antenna is oriented for maximum coupling to the monofilar mode.

EXCITATION OF THE COAL-SEAM MODE IN A CONDUCTOR-FREE REGION

Medium frequency waves can propagate in a conducting coal seam, bounded above and below by more conductive rock, in an approximate TEM transmission line mode with the electric field vertical and the magnetic field horizontal.[1] The magnetic field components produced by a transmitting loop antenna located at the center of the coal seam in a vertical plane are, in terms of cylindrical coordinates with the origin at the center of the loop, the z-axis vertical, and the loop in the r-z plane:

$$H_\phi = \frac{imk}{4} \cos\phi \frac{d}{dr}\left\{H_1^{(2)}(kr)\right\} \tag{1}$$

$$H_r = -\frac{imk}{4} \sin\phi \cdot \frac{1}{r} \cdot H_1^{(2)}(kr)\,. \tag{2}$$

$H_1{}^{(2)}$ is the first order Hankel function for outgoing waves, k is the propagation constant of the coal-seam mode, and m is the effective magnetic moment per unit length, distributed along the z-axis, that is equivalent to the magnetic moment M of the loop antenna located in the coal seam waveguide.

The effects of the currents induced in the rock by the transmitting loop can be represented by an infinite set of uniformly spaced loop images, each of magnetic moment M distributed along the z-axis with complex spacing[2]

$$D = h + (1-i)\,\delta_r, \tag{3}$$

where h is the thickness of the coal seam and δ_r is the skin depth in the rock. The magnetic moment density m, which couples to the TEM coal seam mode, is the zero-order component of the Fourier series that represents the set of magnetic moment images and is given by:

$$m = \frac{M}{D} = \frac{M}{h + (1-i)\,\delta_r} \tag{4}$$

The higher order Fourier series components couple to higher order coal-seam modes which are highly damped, and can therefore be ignored.

This work was supported by the U.S. Department of the Interior, Bureau of Mines, Pittsburgh Mining and Safety Research Center, under USBM Contract HO346045.

The propagation constant k is given by

$$k = \beta - i\alpha \tag{5}$$

where α and β can be calculated by means of the usual parallel plate transmission line formula:

$$\alpha + i\,\beta = \sqrt{(R + i\omega L)\,(G + i\omega C)} \tag{6}$$

which in this case takes the form:

$$\alpha + i\,\beta = \sqrt{(2Z_s + i\omega\mu_o h)\left(\frac{\sigma_c}{h} + \frac{i\omega\epsilon_c}{h}\right)}. \tag{7}$$

Z_S is the surface impedance of the rock, given by

$$Z_S = \frac{1 + i}{\sigma_r\,\delta_r} \tag{8}$$

where

$$\delta_r = \sqrt{\frac{2}{\omega\mu_o\,\sigma_r}}\quad , \tag{9}$$

and σ_r, σ_c are the rock and coal conductivities, ϵ_c is the coal permittivity, and ω is the angular frequency. The permittivity of the rock ϵ_r is ignored since $\omega\epsilon_r \ll \sigma_r$ for the frequencies of interest here.

On taking the asymptotic form of the Hankel function, we find for the direction $\phi = 0$, in the plane of the loop, that the magnitude of the azimuthal component of magnetic field is:

$$|H| \simeq \frac{M\,(\alpha^2 + \beta^2)^{3/4}}{(8\pi)^{1/2}\,[(h + \delta_r)^2 + \delta_r^2\,]^{1/2}} \cdot \frac{e^{-\alpha r}}{\sqrt{r}} \tag{10}$$

This expression is approximately valid at ranges for which $\alpha r > 1$.

ATTENUATION RATE VERSUS FREQUENCY

Medium frequency transmission measurements have been made by T.S. Cory[3] in six coal mines. From a communications point of view, one can make a very meaningful comparison of the various mines directly from the experimental data, by plotting the attenuation constant α versus the frequency f. For this purpose, Equation (10) shows that if, for a given frequency, the experimental values of $H\sqrt{r}$, expressed in decibles, are plotted against r, a straight line of slope proportional to α should be obtained.

Figure 1 shows such a plot of $H\sqrt{r}$ versus r for Consolidation Coal Company's Robinson Run Mine (No. 95) at a frequency of 477 kHz. It is seen that the experimental points plotted in this way do indeed conform to Equation (10). The slope of the best straight line drawn through the plotted points gives an attenuation constant $\alpha = .01253\ m^{-1}$. It is to be noted that the range r has, for convenience, been divided by an arbitrary standard range of 100 m in the derivation of $H\sqrt{r}$ from the data. The value H = 24.0 dB re 1 μA/m on the straight line at this standard range, along with the slope $\alpha = 0.01253\ m^{-1}$, completely specify the experimental data at this frequency.

On repeating this procedure for the data at each frequency used in the experiments, we obtain the α versus f plot shown in Figure 2. It is seen that α increases monotonically with f. This type of curve is found for all the mines.

Figure 3 shows α versus f plots for a large number of mines. It is seen that the mines fall into three categories, representative of the seams in which they lie, namely: the Pittsburgh seam, the Pocahontas No. 3 seam, and the Herrin No. 6 seam, which are in order of increasing attenuation rate α.

DETERMINATION OF THE CONDUCTIVITIES

By separating the real and imaginary parts of Equation (7) we obtain the following expressions for α and β:

$$\alpha = \left\{ (1/2) \left[(p^2 + q^2)^{1/2} + p \right] \right\}^{1/2} \qquad (11)$$

$$\beta = \left\{ (1/2) \left[(p^2 + q^2)^{1/2} - p \right] \right\}^{1/2} \qquad (12)$$

where:

$$p = \frac{2(\sigma_c - 2\pi f \epsilon_c)}{h} \left(\frac{\pi \mu_o f}{\sigma_r} \right)^{1/2} - 4\pi^2 f^2 \mu_o \epsilon_c \qquad (13)$$

$$q = \frac{2(\sigma_c + 2\pi f \epsilon_c)}{h} \left(\frac{\pi \mu_o f}{\sigma_r} \right)^{1/2} + 2\pi f \mu_o \sigma_c \ . \qquad (14)$$

For an assumed value $K_c = 6$ for the dielectric constant of coal, we can determine the values of σ_c and σ_r that give the least square fit between the theoretical and experimental α versus f curves. The solid line in Figure 2 corresponds to $\sigma_c = 3.0 \times 10^{-5}$ Mho/m and $\sigma_r = 8.5 \times 10^{-2}$ Mho/m. The RMS error in α is 0.0018, which is about 5% of the mean α over the frequency range covered. Table I shows values of the σ's determined in this way for the various mines.

FIGURE 1
PLOT OF $H\sqrt{r}$ VERSUS r AT f = 477 kHz
ROBINSON RUN COAL MINE

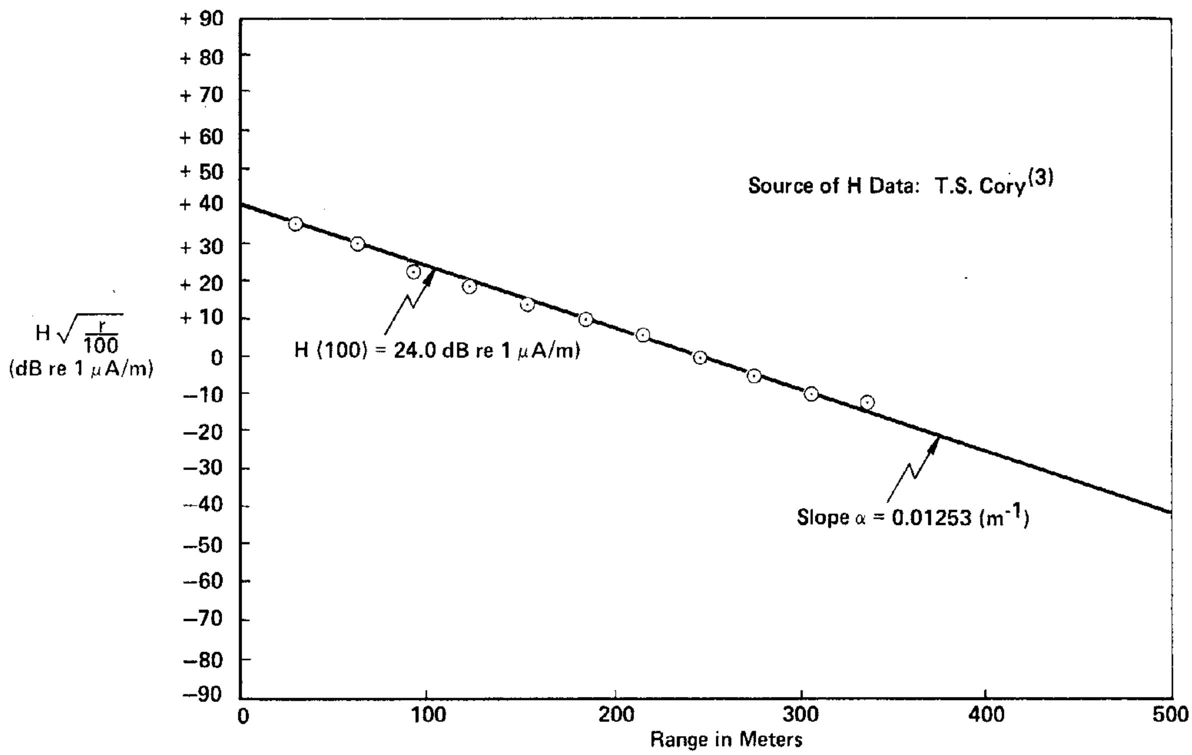


FIGURE 2
PLOT OF α VERSUS f BEHAVIOR
ROBINSON RUN COAL MINE

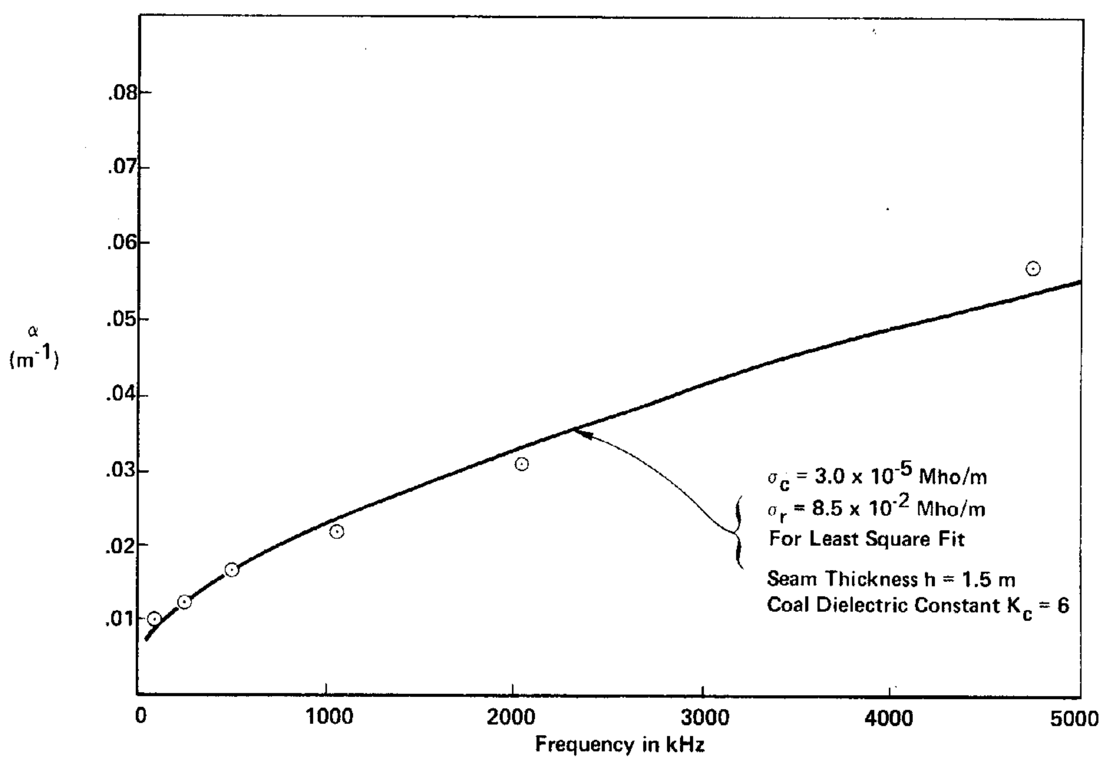
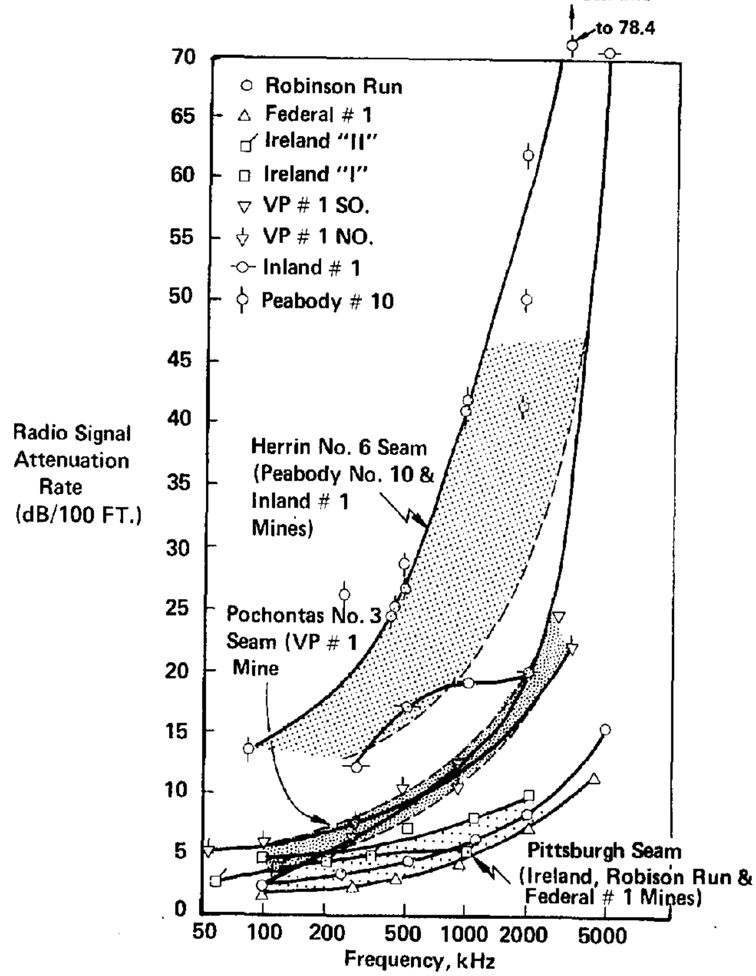


FIGURE 3
COMPOSITE PLOT OF SIGNAL ATTENUATION RATES IN dB/100 FT. FOR SIX MINES IN THREE DIFFERENT COAL SEAMS



COMPARISON OF CALCULATED AND EXPERIMENTAL VALUES OF THE MAGNETIC FIELD

From the values of σ_c and σ_r, we can determine α and β from (11) and (12) and then calculate $|H|$ versus r from (10) for each frequency. The results are shown in Figure 4 for the Robinson Run mine. The experimental curves are given in Figure 5. It is seen that the pattern of the intersecting experimental curves is well accounted for by the theory, although the theoretical values are in general somewhat too high.

The theoretical and experimental values have also been compared on a statistical basis to determine the goodness of fit between the simple three-layer theoretical model predicted results and the measured data. The ratio Δ = H (Theoretical)/H(Experimental) expressed in dB was analyzed statistically as a result of trends shown by plots of Δ versus frequency, at the standard range of 100 meters, for data taken from each of the mines. The Δ plots showed a small and relatively uniform disagreement, mainly on the high side, between about 100 and 1000 kHz, and progressively increasing disagreement, again on the high side, below 100 kHz and above 1000 kHz. The data at the higher frequencies also demonstrated considerable scatter relative to the predicted results.

TABLE I

CONDUCTIVITIES DERIVED FROM THE α VERSUS f PLOTS

Coal Mine	h (m)	K_c	σ_{coal} (Mho/m)	σ_{rock} (Mho/m)	Coal Seam
Robinson Run	1.5	6	0.3×10^{-4}	0.085	Pittsburgh
Federal No. 1	2	6	0.26×10^{-4}	0.084	Pittsburgh
Ireland "II"	2	6	1.0×10^{-4}	0.054	Pittsburgh
Ireland "I"	2	6	2.0×10^{-4}	1.09	Pittsburgh
			1.4×10^{-4}	0.3	Pittsburgh
Pocahontas No. 1					
3 South Area Entry A	1.37	6	3×10^{-5}	0.01	Pocahontas No. 3
3 South Area Entry B	1.37	6	3×10^{-5}	0.0077	Pocahontas No. 3
2 North No. 1 (Plow Area)	1.19	6	6×10^{-5}	0.017	Pocahontas No. 3
Inland No. 1	3	6	1.0×10^{-3}	0.22	Herrin No. 6
Peabody No. 10					
1 Main South 1st West 2nd North	2	6	4×10^{-3}	0.3	Herrin No. 6
1 South 5-1/2 East/ 1 South Jct.	2	6	2.5×10^{-3}	0.3	Herrin No. 6

The statistical analysis of Δ was restricted to data for the frequencies between 100 kHz and 1000 kHz at the reference distance of 100 meters. For frequencies outside the 100 to 1000 kHz range, the Δ plots indicated that the applicability of the simple three-layer model was breaking down. One hundred meters was chosen as the reference distance because it allowed most of the data from six mines in both low and high loss coal seams to be included, while avoiding near field effects in most cases. The ratio Δ in dB between theory and data in the 100 to 1000 kHz band was analyzed by:

- Computing the Sample Mean $\bar{\Delta} = \frac{\Sigma\Delta}{n}$ for each seam and the variance of Δ for the total sample population as a function of frequency (there were too few samples within each seam to compute a standard deviation on a seam basis).

- Computing the Sample Grand Mean for all mines in all three seams, and 95% Confidence Intervals for both the population and Sample Grand Mean as a function of frequency. These results are plotted in Figure 6.

- Testing the hypothesis that the theoretical model fits the data at the 95% confidence level for the frequencies between 100 kHz and 1000 kHz.

FIGURE 4
THEORETICAL CURVES OF H_ϕ VERSUS r WITH FREQUENCY AS A PARAMETER BASED ON σ_c AND σ_r DERIVED FROM DATA
ROBINSON RUN MINE

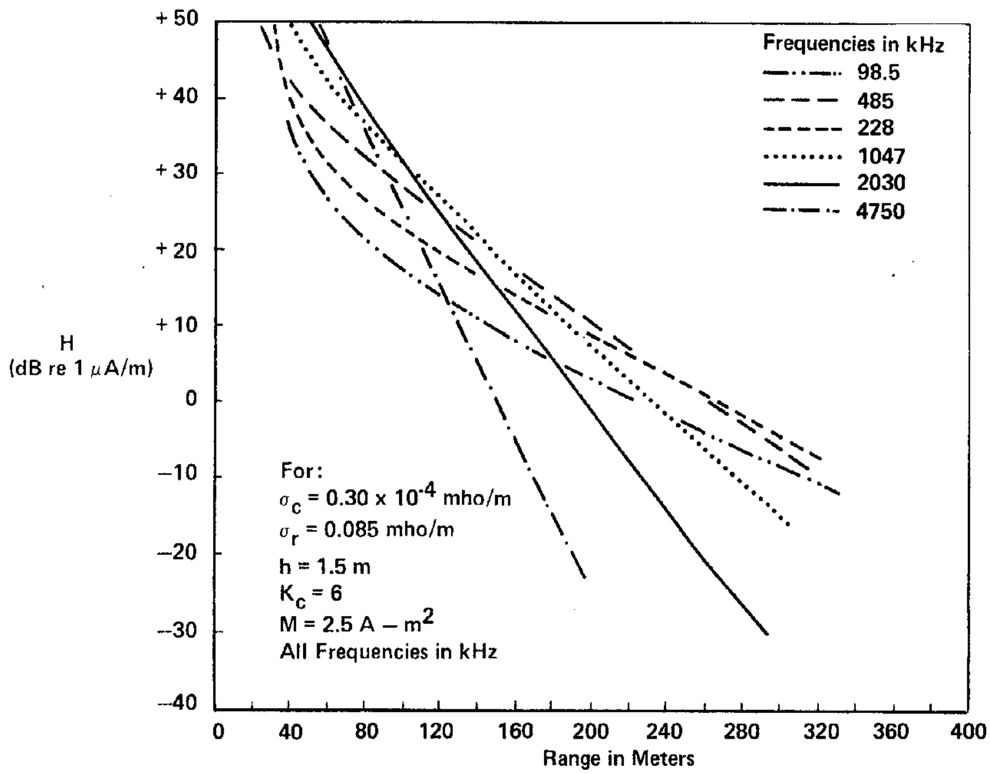
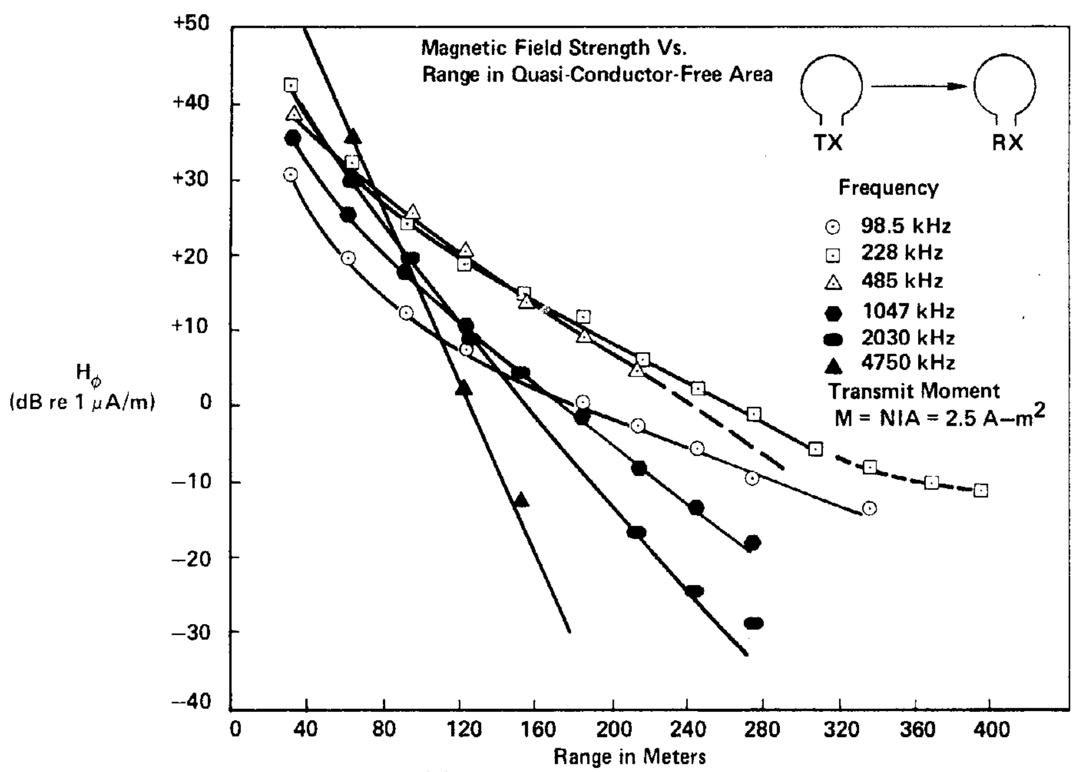
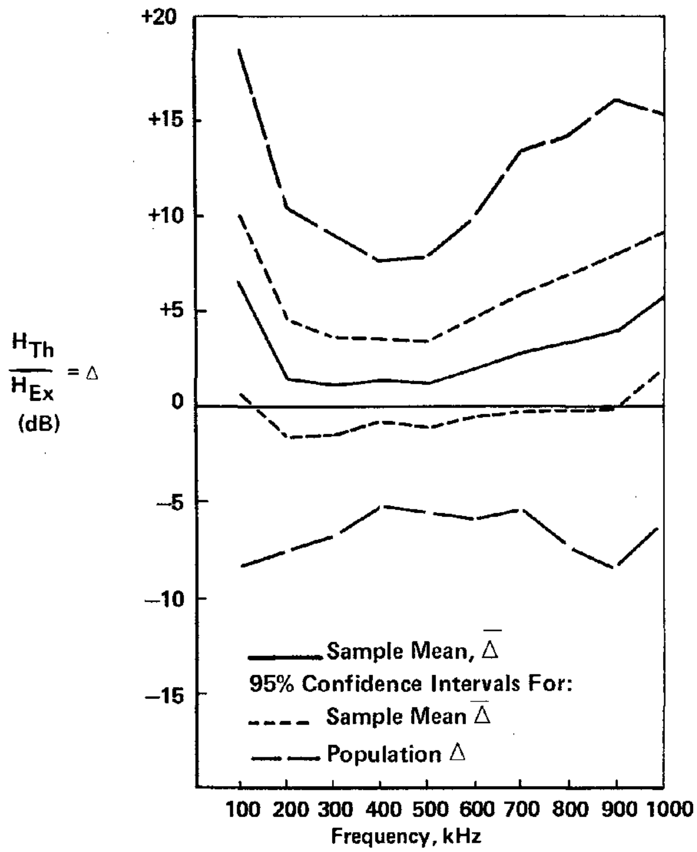


FIGURE 5
EXPERIMENTAL H VERSUS r DATA
ROBINSON RUN MINE



Source: T.S. Cory(3).

FIGURE 6
STATISTICAL SUMMARY OF
H (THEORY)/H(EXPER) RATIOS IN dB



From this analysis, the following observations and conclusions were made. The differences between the model and the experimental data were similar for all coal seams. The shapes of the curves for each seam average followed the same pattern and did not differ from one another by more than 1 or 2 dB, with the exception of the Herrin #6 seam at frequencies below 400 kHz. (At these frequencies, data for only one mine in the Herrin #6 seam was available.) This homogeneity implies that pooling all the data to compute a Grand Sample Mean for all mines is a valid technique, and that the results from any analysis can be applied back to all the mines uniformly.

The plot of the sample $\bar{\Delta}$ and its 95% Confidence Intervals in Figure 6 is a visual hypothesis test (t-test) that the model fits the experimental data, against the alternative hypothesis that the model does not fit the data. The analysis showed that between the frequencies of 200 kHz to 900 kHz, there is *no significant* difference between the model and the actual data. In this band of frequencies the average ratio, Δ, ranges from 1.1 dB to 3.9 dB. The range from 200 kHz to 600 kHz appears to be the best range for the model; both the variance and average Δ between the model and actual data being smallest, and positive, within this frequency range.

CONCLUSIONS

Based on the physical and statistical analyses applied to the data available to date, it can be concluded that:

- MF band radio wave attenuation rates experienced in coal mine conductor-free areas are highly dependent on the coal seam in which a mine is located; and that this attenuation rate versus frequency can be determined in a simple way from measurements of H vs r.

- The simple three-layer model fits the experimental data in the 200 kHz to 900 kHz frequency band at the 100 meter reference distance, with the best agreement occurring between 200 and 600 kHz. This is the band that also promises to provide the most favorable performance for portable radio communications in coal mines.[1,4]

- The Sample Grand Mean, $\bar{\Delta}$, of the ratio H(T)/H(E) expressed in dB may be used to represent the Means for individual coal seams.

- Practical MF system performance estimates may be made in the 200 to 900 kHz band by applying the simple three-layer model to seams in which the attenuation rates have been measured, and using the appropriate $\bar{\Delta}$ to correct the predicted field values.

A more comprehensive theoretical development, analysis of data, and presentation of results can be found in reference 5. Data to date have come primarily from high-coal seams. Data from four additional mines in low-coal seams will be analyzed in the near future to assess the applicability of the model in thinner coal seams, and as a more rigorous test of the theory.

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2. P.R. Bannister, "Approximate Results for the Mutual Electromagnetic Coupling of Loops over a Homogeneous Ground," Naval Underwater System Center Report #NL-3029, New London Laboratory, 23 November 1970, NTIS No. AD717351.

3. T.S. Cory, "Propagation of EM Signals in Underground Mines," Collins Radio Group, Rockwell International Corporation, Final Report, Bureau of Mines Contract H0366028, April 18, 1977.

4. Lagace, R.L. and Emslie, A.G., "Antenna Technology for Medium Frequency Portable Radio Communication s in Coal Mines," Proceedings of U.S. Bureau of Mines Guided Wave EM Workshop, Session III, March 1978.

5. Lagace, R.L., and Emslie, A.G., "Modelling and Data Analysis of In-Mine Electromagnetic Wave Propagation," Arthur D. Little, Inc., Interim Report, Task Order No. 4, Bureau of Mines Contract No. H0346045, to be published, Spring 1978.

COMMENT BY J.R. WAIT

Possibly, a more systematic method for handling the excitation problem would be to work directly with the (non quasi-static) integral representation for the fields of the dipole within the seam*. The residues of the poles of the integrand lead to both the excitation and the attenuation of the discrete modes. The dominant mode has the form given by equations (1) and (2). There is no need to invoke the complex image representation which may be suspect at the lowest frequencies where the seam thickness is comparable with or less than the skin depth of the adjacent rock.

*e.g. see J.R. Wait, "Electromagnetic Waves in Stratified Media", Pergamon Press, 2nd Edition, 1970.

ANTENNA TECHNOLOGY FOR MEDIUM FREQUENCY PORTABLE RADIO COMMUNICATION IN COAL MINES

R.L. Lagace and A.G. Emslie

Arthur D. Little, Inc.
Cambridge, MA 02140

ABSTRACT

Some of the basic principles and limitations of electrically small antennas operating in both free space and conducting media are reviewed. Specific antenna types considered for portable manpack and base station medium frequency radio communication applications in coal mines are: conventional whip antennas, active whip antennas, conventional air-core loop bandolier antennas, ferrite loaded loops, specially resonated multi-turn air-core loops, vertical planar air-core loops, horizontal-wire mode exciters, and vertical-rod mode exciters.

The antennas are assessed with respect to their ability to produce the largest usable signal at the desired range within the constraints of maximizing the efficiency of the total transmitter-antenna system while staying within practical limits of size, weight, convenience, intrinsic safety, and ruggedness of the overall system for use by roving miners. It is concluded that conventional air-core bandolier loop antennas, and perhaps smaller ferrite-loaded loop antennas, are the most suitable choices for roving miner portable radio applications at frequencies below about 1 MHz. For base station installations, horizontal-wire and vertical-rod mode exciters may also offer comparable or better performance than vertical planar air-core loops, but in-mine measurements are needed to resolve the matter.

INTRODUCTION

The U.S. Bureau of Mines would like to achieve a portable-to-portable range of about 1350 ft., especially in conductor-free areas of typical U.S. room and pillar coal mines, namely areas where electrical conductors such as power cables, trolley lines, and communication lines are absent. In the presence of these conductors, achieving the desired communication range is usually not a problem. The 1350 ft. conductor-free range corresponds to that required for two miners to communicate between the leftmost and rightmost entries (spaced about 600 ft. apart) of a seven-entry mine development while also being separated by a longitudinal distance of about 1000 ft. along the entries. This requires an area-coverage capability that is not adversely affected by the presence of the largely rectilinear grid of tunnels in the coal seam. Though extremely useful in several mine applications, UHF radio coverage does suffer from high losses when propagating around corners and through permanent ventilation stoppings. The frequency band from 50 kHz to 1 MHz was selected for investigation because radio waves in this band propagate through the coal seam as if the network of mine tunnels was absent.

The desire is to obtain a small, lightweight, unobtrusive, antenna for a personal radio that a roving miner can use to achieve the desired communication range goals in the LF to MF band of 50 kHz to 1 MHz. The antenna's impedance characteristics should also be relatively unaffected by the antenna's proximity to the miner's body, or structures found in underground coal mines. The antenna-transmitter system should provide the largest signal strength compatible with range requirements and battery and intrinsic safety constraints. This paper presents a summary assessment of antenna technology and mode coupling techniques related to this application.

ELECTRICALLY SMALL ANTENNAS

Free space wavelengths for operating frequencies from 50 kHz to 1000 kHz take on values of 6,000 to 300 meters respectively. Comparing these wavelengths to a practical dimension d = 0.5 meters for a personal antenna worn by a roving miner, we get $8 \times 10^{-5} < \frac{d}{\lambda} < 2 \times 10^{-3}$, which immediately places such an antenna well inside the category of "electrically small" antennas, namely, antennas having $d < \frac{\lambda}{10}$. The more the size of an antenna is decreased with respect to wavelength, the more it behaves as a high-Q, lumped-element, energy storage device, and as a poor radiator of power to the surrounding space. These limitations are fundamental, and are directly attributable to the antenna's size compared to wavelength, as illustrated in a classic paper by Chu [1] in 1948, and also by Harrington[2] and Wheeler.[3,4]

This work was supported by the U.S. Department of the Interior, Bureau of Mines, Pittsburgh Mining and Safety Research Center, under USBM Contract H0346045.

Radiation Resistance and Efficiency

Electrically small, lossless, whip-type dipole antennas behave as capacitors, and loop-type antennas behave as inductors. Their input impedances are highly reactive, and have only a very small resistive part defined as the radiation resistance R_r – a resistance which is a measure of the power radiated by an antenna into a lossless medium, and which rapidly decreases in value as the antenna size decreases relative to the wavelength.

For practical antenna dimensions of 0.5 meter diameter for a 10-turn loop, a 0.25 meter effective height for a dipole, and a 1 ampere input current to each from a portable radio operating at the MF frequency of 300 kHz, we find that the radiation resistances and radiated powers are $R_{rd} = 5.5 \times 10^{-6}$ ohm, $P_{re} = 2.75$ μw, and $R_{r\ell} = 1.5 \times 10^{-9}$ ohm, $P_{r\ell} =$ .0007 μw, for the dipole and loop antennas respectively. (The comparative difficulty of generating this current in each antenna with a voltage source is ignored for the present.) When compared with the resistive losses associated with practical antennas, these radiation resistances become insignificant. This accounts for the extremely low efficiencies attainable from man-carried antennas at these low operating frequencies.

For example, the efficiency of an antenna acting alone is given by

$$E = \frac{P_r}{P_r + P_d} = \frac{R_r}{R_r + R_d} \tag{1}$$

where P_r is the power radiated and P_d is the power dissipated in the ohmic resistance R_d associated with the antenna conductors. Therefore, if the 10-turn, 0.5 meter diameter loop is made of #16 AWG copper wire, its ohmic resistance will be 0.73 ohm including skin and proximity effects, thereby limiting the antenna efficiency to the negligible value of only 2×10^{-7}%. This is the best that can be done. The efficiency of the transmitting system is, of course, further degraded by the resistance associated with the source impedance, and with any tuning and coupling circuitry that may be required to drive such a highly-reactive, high-Q antenna from a practical power source. (see comments on pg. 180, Ed.).

Q, Power Factor, and Effective Volume

Figure 1 presents a curve of the behavior of the intrinsic quality factor Q_r of electrically small antennas, based on the work of Chu.[1] The quantity "a" represents the radius of the minimum size sphere within which the antenna can be enclosed. This Q_r is the ratio of ω times the peak electromagnetic energy stored in the near field by the antenna to the average power radiated into the far field. Figure 1 shows how rapidly Q_r increases with decreasing antenna size compared to wavelength, vividly illustrating the high energy storage and narrow intrinsic bandwidth characteristics of electrically small antennas.

The radiation power factor pf_r and the concept of antenna effective volume V′ introduced by Wheeler[3] are also convenient and powerful ways of defining the intrinsic limitations of small antennas. The efficiency can also be expressed as

$$E = \frac{pf_r}{pf_r + pf_\ell} = \frac{Q_{\ell r}}{Q_r} , \tag{2}$$

where

$$pf_r = \frac{R_r}{\omega L} = R_r \omega C = \frac{1}{Q_r} = \frac{2}{9}\frac{V'}{V_s} , \tag{3}$$

and

$$pf_\ell = \frac{R_\ell}{\omega L} = R_\ell \omega C = \frac{1}{Q_\ell} . \tag{4}$$

L and C are the intrinsic inductance and capacitance of the loop and dipole antennas respectively, pf_ℓ is the external circuit power factor, R_ℓ is the ohmic resistance of the antenna and tuning elements, Q_ℓ is the Q due to these ohmic losses alone, and $Q_{\ell r}$ is the Q which includes both ohmic and radiation losses. V′ is the physical volume of the antenna multiplied by a shape factor greater than unity, k_e for electric dipole-type antennas and k_m for magnetic dipole-type (loop) antennas. V_s is the volume of the "radian sphere," a sphere of radius equal to one radian length, namely $\frac{\lambda}{2\pi}$, within which most of the near field stored energy associated with the far field radiated power of electrically small ($d < \frac{\lambda}{10}$) antennas of either type is confined.[1,5] In addition, pf_ℓ will be further increased by the real part of the source impedance seen by the tuned antenna circuit, and pf_r will be decreased by the efficiency of coupling to the tuned circuit.

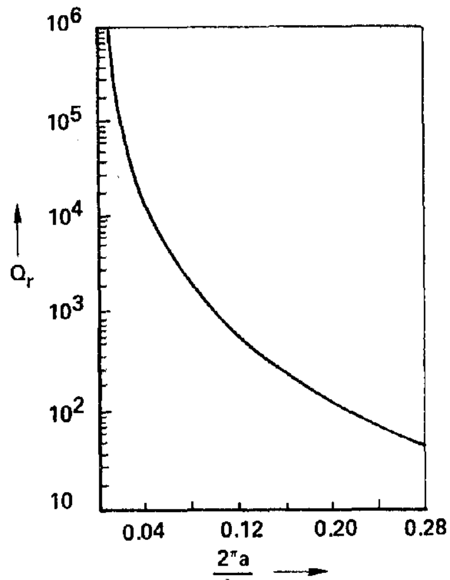
Figure 2 depicts comparative examples[4] of the effective volumes of several common dipole and loop antenna types. Figure 3 illustrates how well the volumetric comparison method allows a quick grasp of the fundamental limitations imposed by antenna size alone on the radiation power factor, the Q, and ultimately on the efficiency. Applying this effective volume method to either a thin electric dipole or flat air core loop of maximum dimensions d = 0.5 m gives a pf_r and Q_r on the order of 10^{-9} and 10^9, respectively. Taking the ratio of Q_r to a practical tuning circuit $Q_{\ell r}$ of 100 (i.e., having a bandwidth of 3 kHz at 300 kHz) results in a negligible radiation efficiency on the order of 10^{-7}%, as computed on the basis of radiation resistance.

Application Implications

The fundamental limitations imposed on electrically small antennas are severe. Thus, it is not surprising that portable two-way radio systems for most conventional applications on the surface are designed to operate at frequencies primarily in the VHF and UHF bands, and to a much lesser extent in the HF band, bands in which wavelengths are less than, or comparable to, the size of a person and thus the typical portable whip-type antenna. Even at HF, antenna problems are difficult for portable units,[6,7,8,9] and below HF the problems get rapidly worse, even for fixed station transmitters. To achieve reasonable efficiencies with electrically small antennas, overall system Q must be kept high. Below HF this is translated into the ability to construct large, low loss antenna structures tuned by large low loss circuit elements [10,11,12]. The tall radio tower antenna structures used by AM broadcast stations in the MF band are familiar examples.

The portable MF wireless mine application requires at least about 3 kHz of bandwidth for a single-sideband voice system and about 12 kHz for a conventional narrowband FM

FIGURE 1
BEHAVIOR OF ANTENNA INTRINSIC Q_r



Source: Ref. 27

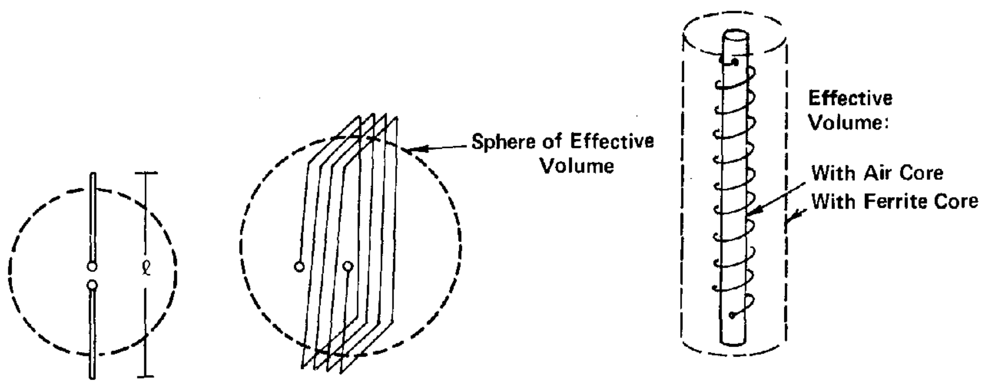
system. Thus, the portable transmit antenna size constraint, together with the system bandwidth required, make a goal of high radiation efficiency not only unattainable, but also undesirable for the LF and MF radio bands from 50 kHz to 1000 kHz of interest. Thus, other factors take on a much greater importance than efficiency in forming a basis for antenna selection.

IMPACT OF CONDUCTING MEDIA

Basic Limitations

When electrically small antennas are immersed in a conducting medium as in the case of the MF wireless mine application, additional factors come into play which dominate the choice of methods and antennas to optimize radio system performance. First and foremost is the severe signal attenuation introduced by the resistive losses experienced by the waves in the conducting medium at MF. The conductivity of the medium introduces power dissipation external to the antenna conductors. This reduces both the radiated and reactive fields and powers that would be present in a free space medium, thereby greatly reducing the attainable communication range. Furthermore, the free space definitions of radiation resistance and radiated power are no longer meaningful measures of antenna efficiency or performance. The so-called radiated power is no longer a fixed value independent of distance in the antenna far field, and the radiation resistance is no longer a measure of the input power coupled to or consumed by the medium external to the antenna.

FIGURE 2
EFFECTIVE VOLUMES OF SEVERAL COMMON ANTENNA TYPES

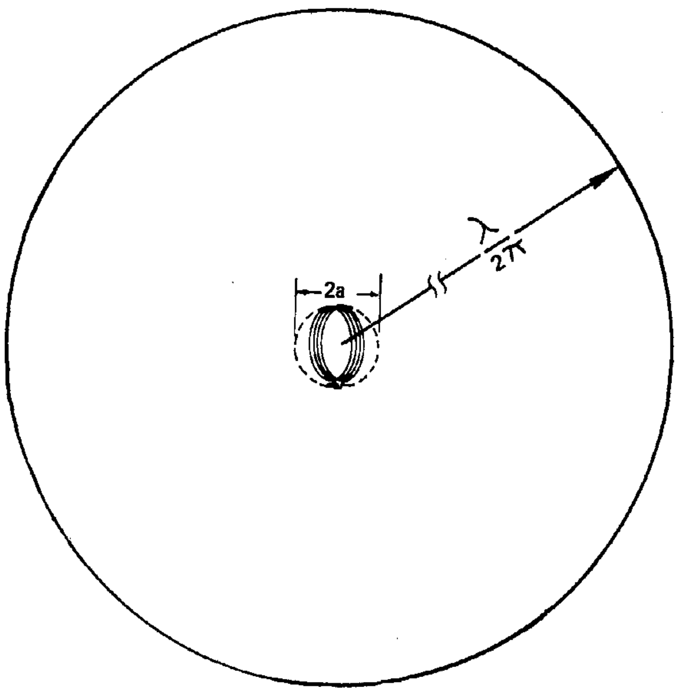


(a) Dipole (b) Loop (c) Air And Ferrite Core Loops

Source: Ref. 4

Limitations imposed by signal attenuation and representative mine layouts have led to the relatively modest range goal of 1350 feet for a portable, wireless radio system. This range falls well within the antenna near field region in free space, and by and large within the antenna induction field region in a coal mine conducting medium. Thus, the choice of antenna for the portable wireless mine application will not be based on its radiation efficiency, but on:

FIGURE 3
EFFECTIVE VOLUME OF ELECTRICALLY SMALL ANTENNA RELATIVE TO VOLUME OF A RADIAN SPHERE



- the *overall* power utilization efficiency of the total transmitter-antenna *system* in producing the largest usable signal strength at the desired range, and

- the overall practicality of its implementation with regard to intrinsic safety, and overall system size, weight, convenience, battery life, and mine worthiness for the roving miner application.

Loop Versus Dipole Antennas

A simple performance comparison illustrates how loops become superior to whip antennas in conducting media. Namely, we compare the loop-to-loop and dipole-to-dipole responses to the principal field components(13) of electrically small loop and dipole transmit antennas embedded in a conducting medium in which the condition $\sigma >> \omega\epsilon$ is satisfied. The responses apply for a loop fully insulated from the medium, and a dipole insulated along its length but grounded to the medium at each end of the dipole. In practice, this type of dipole is, of course, not suitable for a portable manpack radio unit.

Taking the ratio of the open circuit output voltage induced in the receive antenna of each pair of like antennas (i.e., dipole-dipole, loop-loop), we get the comparison factor

$$F_{d,\ell} = \frac{I_d h^2}{I_\ell N^2 \omega\mu\sigma} \tag{5}$$

Substituting into (5) the representative value of dipole length h = 0.5 m, loop area A = 0.2 m^2 based on a 0.5 m loop diameter, N = 10 loop turns, $\mu = \mu_o = 4\pi \times 10^{-7}$ H/m, medium conductivity $\sigma = 10^{-1}$ Mho/m, and frequency f = 300 kHz, $F_{d,\ell}$ reduces to $F_{d,\ell} = 0.1\ (I_d/I_\ell)$, or $F_{d,\ell} = 0.1$ for equal antenna currents. If both antennas are instead driven by the same voltage, the resultant performance factor $F_{d,\ell}$ is given by:

$$F_{d,\ell} = \frac{h^2}{N^2 \omega\mu\sigma} \frac{Z_\ell}{AZ_d} \tag{6}$$

The resonant impedance Z_ℓ of a tuned loop having a bandwidth of 12 kHz is about 10 ohms resistive, whereas the practical grounded dipole source is faced with the spreading resistance of its ground connections (if ideal), which is about 20 ohms in a medium having $\sigma = 10^{-1}$ Mho/m. Substituting these values into (6), including a 1 m increase in dipole effective length caused by the ground rod terminations, we get $F_{d,\ell} \approx 0.1$, the same value as for an equal current drive. Since the medium conductivity σ also appears in the numerator of $F_{d,\ell}$ when Z_d is the spreading resistance of ground rods, this performance factor remains relatively insensitive to the conductivity of the medium within its region of applicability, thus giving a clear performance advantage to the loop in a uniform conducting medium. Non-ideal grounding connections will, of course, increase this performance advantage.

The advantage of the loop in the MF band becomes even greater when compared to a whip-type dipole antenna,[7] which would be the practical choice instead of a grounded dipole for a roving miner's portable radio. This whip would be a completely insulated antenna not grounded to the medium at each end. The input impedance of this antenna worn as part of a manpack by a miner standing in a tunnel will be a large and highly variable capacitive reactance, instead of the modest resistance of the grounded dipole. The capacitance represented by a short whip worn on a person is very low (approximately 10 pf), and subject to large variations[6,7] due to "stray" capacitances such as those introduced by the whip's variable poximity to, and position on, the body and mine tunnel environment. The practical problems associated with tuning and driving such antennas become intolerable (viz the experience at HF with 6-foot whips).[6] Whips also fail to meet the objectives of smallness, ruggedness, and convenience of wear for use in the mine environment.

An active whip transmitting configuration reported [14,15] to offer certain performance advantages over that of a conventional series-tuned whip antenna in the MF band was also assessed,[16] namely the fed-emitter base loop (FEBL) transmitting antenna configuration. This active antenna configuration does not overcome the considerable disadvantages of whip type antennas for the MF mine application, and moreover it is significantly more wasteful of power.

In summary, the considerable problems associated with whip antennas for use in portable manpack applications at and below HF are not new. Thus, it is not surprising that whip antennas are placed at a considerable disadvantage compared to loops for MF band wireless radio mine applications.

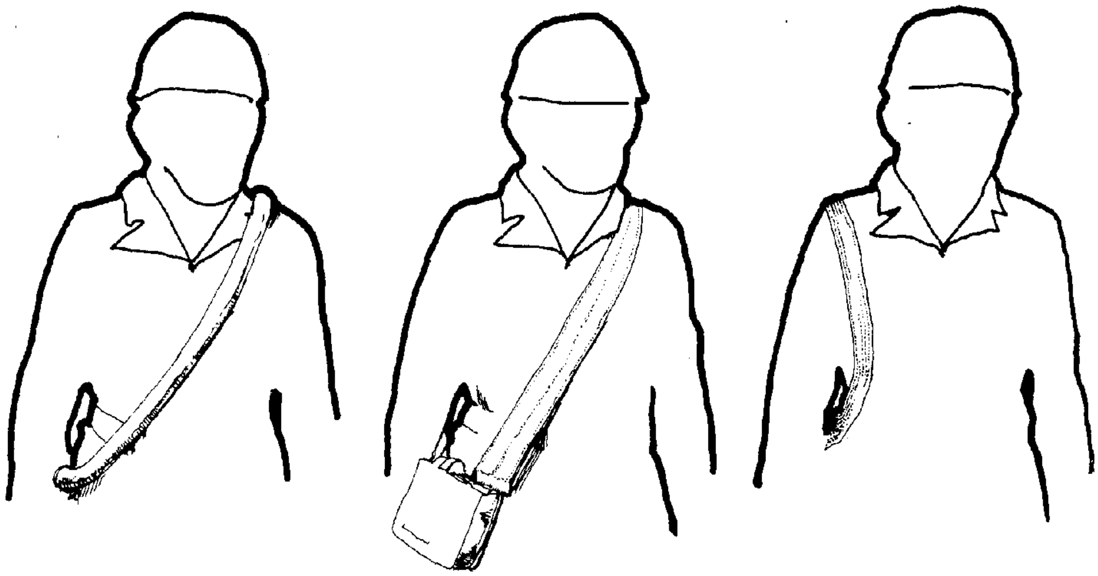
LOOP ANTENNAS

Air-Core Loops

Conventional air-core loop antennas represent one of the most suitable and effective choices for the MF wireless radio roving miner application. Thus, it is not surprising that the South Africans[17,18] and the British[19,20] have adopted the use of such antennas for rescue-type, portable MF band communications in mining and fire fighting applications respectively. Similar developments are also occurring within the U.S. coal mining community.[18,21,22] However, the state-of-the-art is such that, even with such loops, a completely wireless range goal of 1350 feet in U.S. coal mines is likely to be attainable only in the most favorable mine environments.[23,24,25] On the other hand, ranges on the order of 1 – 3 km appear possible[16,17,19,26] when such loop antennas are employed in the vicinity of mine electrical conductors.

The conventional loop antenna has a large number of positive attributes for roving miner applications. At MF frequencies, it is an inductively reactive device whose magnetic fields and input impedance remain for all practical purposes, unaffected by its location on and closeness to the body of a person, or by the local mine environment. The stray capacitance problems associated with whip antennas are avoided. The moderate inductive reactance of a typical loop configuration having 10 turns, can be easily tuned by readily available practical capacitors in the MF band. The loop can be readily formed into a lightweight, rigid or flexible, bandolier shape. This allows the area of the loop, and therefore the transmit magnetic moment, to be maximized within a form factor that can be tailored to protect the loop from damage while providing the least inconvenience to the miner wearing it. Some examples of current designs are presented in Figure 4.

FIGURE 4
SKETCHES OF SEVERAL BANDOLIER LOOP ANTENNA CONFIGURATIONS



a. Rigid Oval Tubular Type b. Flexible Strap Type c. Garment Type

The rigid tubular elliptically shaped banoliers (Figure 4a) used by the 335 kHz South African radio and the 520 kHz Collins Radio prototype unit provide the advantage of a fixed area, and therefore fixed transmit moment and loop inductance, but the disadvantage of somewhat greater awkwardness than flexible bandolier configuration such as the strap-type bandoliers (Figure 4b) used by Plessey in its 140 kHz Inductorfone TGR.1 unit and the garment-type (Figure 4c) used by Lee Engineering Division of Consolidation Coal Co. in one of its 425 kHz prototype units. The garment type can take the form of turns (typically stranded wire) sewn in the desired shape into a miner's coveralls or into a specially designed jacket or pullover worn by the radio-equipped person. The loop orientation which results from wearing a bandolier type of antenna (i.e., in a nearly vertical plane) also closely approximates the ideal orientation for coupling to the quasi-TEM mode of propagation in the coal seam waveguide.[23, 24]

A specific subclass of electrically small air-core loop antennas named MTL's (multi-turn loops)[27, 28, 29, 30] by Ohio State University investigators were also assessed.[16] Although these MTL antennas may offer size and/or performance advantages for some applications in the HF and VHF bands, they are inappropriate for MF band portable radio mine applications.

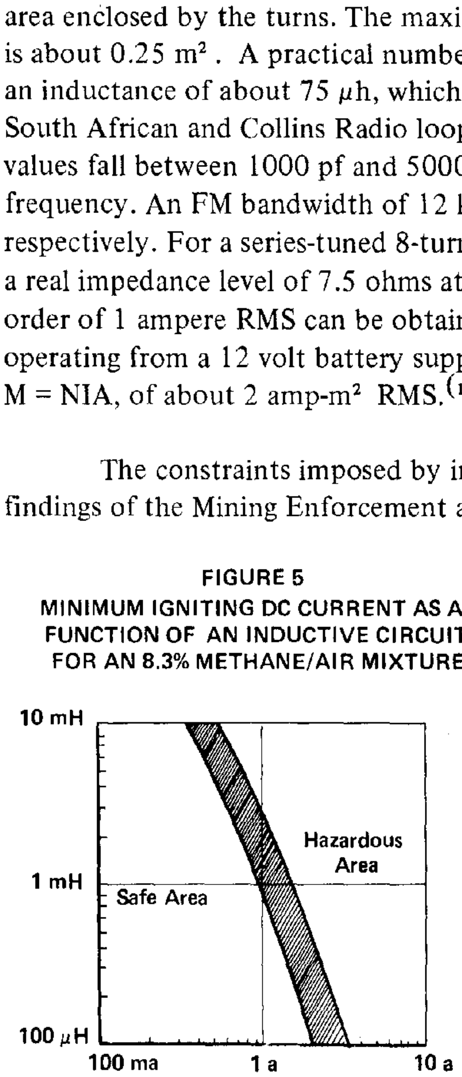
Design Objectives and Limitations

The first design objective is to generate the largest practical antenna magnetic moment within the constraints imposed by intrinsic safety, reasonableness of size and wearability, and practicality of tuning and matching elements. The second and equally important design objective is to produce this magnetic moment by the most efficient use of the energy stored in the battery of the portable unit. The signal attenuation rates and electromagnetic noise encountered in typical coal mines is so severe that attempts to achieve small-to-moderate increases in transmit magnetic moment, at the expense of overall transmitter efficiency or power dissipation, are generally not warranted. In a roving miner portable radio application, it is better to have a smaller, lighter-weight, longer-lasting unit that has a somewhat shorter range, than another unit which has 100 – 200 feet greater range under some highly favorable mine conditions, but is too heavy and bulky for the miner to wear, or has too short an operating time between battery recharges.

The transmit moment of a conventional air-core loop is maximized by maximizing the product M = NIA of the loop, where N is the number of turns, I is the loop current, and A is the area enclosed by the turns. The maximum reasonable loop area A for a roving miner application is about 0.25 m^2. A practical number of turns N is in the 5 to 15 turn range. Seven turns gives an inductance of about 75 μh, which is close to the measured values of 60 μh and 82 μh for the South African and Collins Radio loop antennas, respectively. The associated tuning capacitor values fall between 1000 pf and 5000 pf, depending on the inductance value and operating frequency. An FM bandwidth of 12 kHz in the 300 to 500 kHz band requires a Q of 25 to 42 respectively. For a series-tuned 8-turn, 0.25 m^2, loop of nominal 100 μh inductance, this means a real impedance level of 7.5 ohms at the operating frequency. Thus, a loop current I on the order of 1 ampere RMS can be obtained in an efficient manner from a Class D type transmitter(16) operating from a 12 volt battery supply. This allows the generation of a practical transmit moment M = NIA, of about 2 amp-m^2 RMS.(16)

The constraints imposed by intrinsic safety are shown in Figure 5. These are based on the findings of the Mining Enforcement and Safety Administration (MESA) for DC inductive circuits operating in explosive methane gas atmospheres. For example, it reveals that for an inductance of 100 μh, a current of 1 ampere is already close to the band defining the zone of maximum safe current level. In AC applications, this safe level must represent the peak current(33) level. So the practically attainable 1 ampere RMS level, compared to the allowed 2 amps peak, includes a modest safety factor. Significant gains in transmit moment cannot be obtained from increases in current at this inductance. Nor can they be obtained by increasing or decreasing the number of turns.(16) The rf voltages across the tuning capacitor (and the loop) at an operating frequency of 500 kHz fall well within the safe area of the intrinsic safety plot for capacitors.(31) However, it should be realized that the MESA intrinsic safety curves are based on DC circuit experience. The subject becomes much more complex(33) for AC applications, so the DC-based curves serve only as approximate design guidelines for the MF band radio application. In the end, the intrinsic safety of any MF wireless portable radio will be verified only after the radio, including the antenna, has successfully completed methane ignition tests conducted by MESA in its special test facility.

FIGURE 5
MINIMUM IGNITING DC CURRENT AS A FUNCTION OF AN INDUCTIVE CIRCUIT FOR AN 8.3% METHANE/AIR MIXTURE

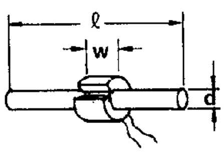


In summary, practical values of magnetic moment M for a portable MF band unit are not likely to exceed the values of about 2.5 amp-m^2 peak available with present equipment. In highly favorable mine attenuation and noise environments, this may result in achieving the range goal of 1350 feet. In fair-to-poor mine environments, ranges of about 200 feet will occur. However, although the completely wireless communication range will remain hopelessly small in these situations, effective ranges of 1 – 3 km may be attainable when in the vicinity of long runs of mine electrical conductors. Thus, the major objective for MF mine wireless portable radios for miners should be to obtain the most reasonable value of magnetic moment consistent with a transmitter/antenna circuit design that provides the most efficient use of the battery's energy supply and the smallest portable unit size and weight.

Ferrite-Core Loops

In many applications, loops loaded with magnetic core materials offer equal or improved performance in a considerably reduced cross-sectional area and volume over that of an air-core loop.[10] A typical configuration for many ferromagnetic core loop antenna applications is shown in Figure 6.[10, 11] Feasibility calculations[16] show that transmit moments equivalent to those obtainable with air-core bandolier loops are possible with small ferrite-core loops. Such loops could be conveniently mounted horizontally to the radio unit itself or to the miner's belt on which the cap lamp battery and self-rescuer are attached.

FIGURE 6
TYPICAL CONFIGURATION FOR LOOPS LOADED WITH FERROMAGNETIC CORES



Source: Ref. 10

The calculations were performed for a 10 in. long, 1 in. diameter rod (ℓ/d = 10) of manganese-zinc (MnZn) ferrite material typically used for antenna rods in the LF and MF bands from about 100 kHz to 2 MHz. This material typically has a saturation flux density B_{sat} of about 0.3 Tesla and an intrinsic initial permeability μ of 500 to 1000[34, 35]. It was found that a single-layer winding of about 28 turns over a two inch length of the 10 in. by 1 in. ferrite rod should provide the desired peak moment of 2.5 A-m² with inductance and peak current levels of about 80 μh and 1.8 amperes, respectively.

As in the case of the air-core bandolier loop, the maximum intrinsically-safe moment is constrained to be about 2.5 amp-m² peak. This value is well below the maximum attainable moment of M_{sat} = 31 A–m² peak of the rod, thereby suggesting that the desired moment of 2.5 A-m² could be achieved with an even smaller volume ferrite-core loop antenna. The favorable performance of ferrite loaded loops as receiving antennas[10, 11, 36] make their use doubly attractive. Consequently, effort should be directed toward verifying this predicted performance. This can best be accomplished experimentally, as will the future optimization of any practical design.

FIXED INSTALLATIONS

Some alternative antenna types and radio wave coupling methods suited to MF radio fixed installations such as base stations or repeaters are treated below.

Vertical Planar Loops

Air-core planar loops, series tuned, provide a practical choice for the MF portable radio units. They also provide the most convenient way of generating a large magnetic moment for a fixed radio installation. The planar loop combines the desirable attributes of high magnetic moment, well-defined and stable impedance characteristics, and ease of installation and maintenance. A practical loop configuration for a fixed installation is a rectangular air-core loop, series tuned, and formed in a vertical plane so that it can be "hung" or mounted on the wall or rib of the mine tunnel as in Figure 7. The wall-mounted loop presently used for fixed installations of the 500 kHz FM mine wireless prototype units in high-coal seams is a one-turn, 2 meter high by 7 meter long loop having an NA = 12 m^2 and an inductance of 27 μh. Its approximately 2 ohm tuned input impedance is driven with an intrinsically safe peak current of about 1 ampere to produce a peak magnetic moment of M = 14 A-m^2.

The vertical orientation of the planar antenna is the most favorable for exciting the desired quasi-TEM mode having horizontal magnetic field H_ϕ and vertical electric field E_z in coal seams.(23) Maximum signal strength occurs at positions in the plane of the loop ($\phi = 0°$), and minimum signal strength at locations in the vertical plane perpendicular to, and passing through the center of, the loop ($\phi = 90°$), according to the cosine angular dependence of a magnetic dipole (loop) source. The angular dependence of the mode excitation presents a nonuniform area coverage problem that is more serious than the one experienced by a miner wearing a bandolier loop antenna. A miner can always rotate his body position to maximize his received signal-to-noise ratio. The same result can be obtained for fixed installations by orienting a second vertical planar loop perpendicular to the first loop, and driving it with a current equal to, and 90° out of phase with, the current in the first loop as in Figure 8. This turnstyle antenna configuration with its electrically rotating antenna pattern provides omnidirectional coverage in the horizontal plane of the coal seam.

FIGURE 7
WALL-MOUNTED FIXED STATION LOOP IN COAL MINE TUNNEL

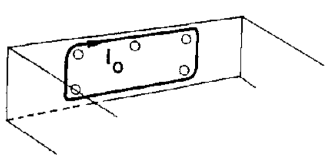
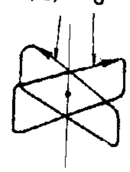
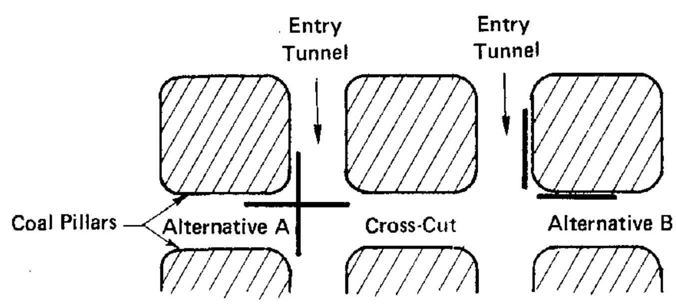


FIGURE 8
IDEAL TURNSTYLE-TYPE ANTENNA CONFIGURATION



Ideally, the two loops should have a common center as shown in Figure 8 so as to ensure a common electrical phase center for the waves and minimize mutual coupling between the two loops. However, installing two such perpendicular vertical loops spanning the intersection of an entry tunnel and a cross-cut as shown in Figure 9, alternative A, may not be practical at intersections used by haulage vehicles unless special precautions are taken to protect the loop conductors. The alternative B installation shown in Figure 9 avoids the need for such protection, by mounting the loops on the perpendicular entry and cross-cut walls of a coal pillar. The noncoincident phase centers should not create any significant pattern distortions in the LF-MF band, but mutual coupling between the loops may have to be accounted for when tuning and driving the loops.

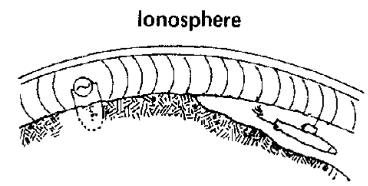
FIGURE 9
FIXED-STATION TURNSTYLE LOOP ANTENNA CONFIGURATIONS



Horizontal Grounded Long Wire

Mode excitation techniques or antennas(37, 38, 39) similar to those used for the Navy's proposed worldwide ELF Sanguine communications system may also be applicable for wireless radio fixed installations in coal mines. Recent and continuing analyses (23, 24, 25) and measurements(26) have shown that coal seams bounded above and below by sedimentary rock behave as lossy parallel plate waveguides with respect to the propagation of radio waves in the LF and MF frequency bands. At ELF, the conducting ionosphere and surface of the earth form the upper and lower boundaries of a similar parallel plate earth-ionosphere waveguide as depicted in Figure 10, within which transmitted ELF waves travel around the world with very low attenuation to be received by submerged submarines. Perhaps the most important difference between the two waveguides is that the coal seam waveguide is filled with a lossy medium, the coal itself, in addition to being bounded by lossy media above and below, thus introducing shunt loss in addition to the series loss of the "plates" in both waveguides. The spherical shell nature of the earth-ionosphere waveguide introduces only a different geometrical spreading loss factor. However, neither of these differences, the shunt loss nor the spherical shell, changes the basic nature of the quasi-TEM mode of propagation in the waveguide. Thus, the excitation methods for one waveguide should be largely applicable to the other.

FIGURE 10
PROPAGATION AT ELF IN THE EARTH-IONOSPHERE WAVEGUIDE



Source: Ref. 37

The Sanguine investigators found that the most effective, feasible transmitting antenna configuration for their application was a parallel array of grounded, center-driven long wire antennas. A single long horizontal wire is the simplest form of such an antenna. However, to achieve the required current moment Iℓ requires an excessively long antenna (and thus land having the desired conductivity) and/or unreasonably high currents. High currents lead to two consequences that are undesirable for both the Sanguine and wireless mine applications, namely excessive power dissipation and environmentally nonpermissible field or current levels. A parallel array greatly reduces the maximum linear dimensions of the antenna, and distributes the required current moment over an area, which in turn allows the use of several parallel wires, each carrying a fraction of the total current required. Furthermore, it was found that a spacing between horizontal wires of about 2 skin depths provides a favorable compromise between overall system performance, power dissipation, and cost for the intended application.

Thus, one can visualize a wireless fixed station mine application that employs one or more horizontal long wire antennas strung along the roof of a mine main tunnel or crosscut and grounded to the rock above the coal seam by roof bolts or other means as portrayed in Figures 11 and 12. Recent estimates place the conductivity of roof rock in coal mines within the range

FIGURE 11
GROUNDED LONG WIRE ANTENNA INSTALLED IN A COAL MINE TUNNEL

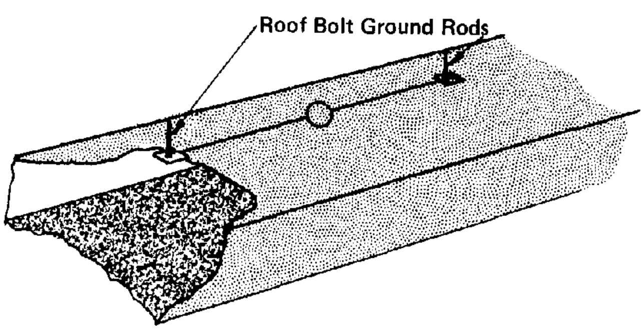
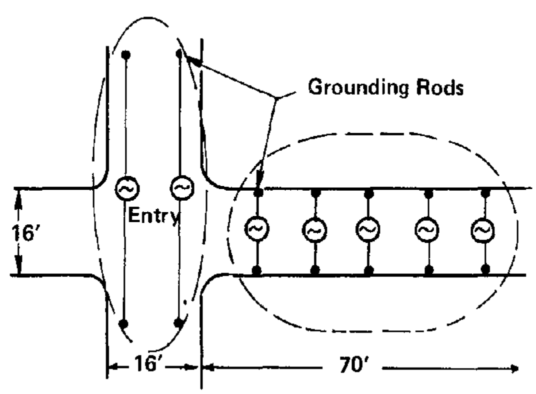


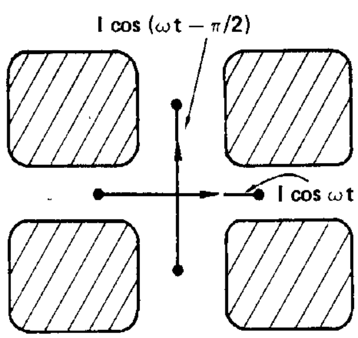
FIGURE 12
LONG WIRE ARRAY CONFIGURATIONS IN A MINE ENTRY AND A CROSSCUT



of 0.01 to 1.0 mho/m with a value of about 0.1 mho/m being typical. At an operating frequency of 300 kHz, this produces a skin depth in the rock, δ_r, of about 2 meters. Thus, the Sanguine 2δ separation criteria would allow at least two, and perhaps three, parallel long wires to be placed in a typical mine tunnel. Another possibility would involve a similar or greater number of considerably shorter long wire antennas installed in a crosscut tunnel between two entries as in Figure 12.

The long wire antenna exhibits the same $\cos\phi$ dipole pattern angular dependence as a vertical planar loop. Thus, omnidirectional area coverage can be obtained in the same manner as with loops, by employing a second long wire antenna perpendicular to the first and driven 90° out of phase with it in turnstyle fashion shown in Figure 13. The long wire antenna is better suited than a loop to turnstyle installations across the intersections of mine entry tunnels and crosscuts, because it does not require a floor-mounted wire which is generally difficult to protect from damage in well-travelled intersections.

FIGURE 13
LONG WIRE TURNSTYLE CONFIGURATION IN AN ENTRY/CROSSCUT INTERSECTION



A first-order estimate of the performance of a single, terminated long wire compared to that of a conventional rectangular vertical planar loop antenna can be made by computing the effective magnetic moment of the "loop" formed by the long wire and its return current path in the roof. When the antenna is long compared to the skin depth in the rock, the return current flow spreads out into the roof to about one skin depth from the wire.[37, 38] Thus, the long wire and its return current path can be approximately represented by a rectangular loop of length ℓ, effective height $\delta_r / \sqrt{2}$,[38] and effective area $A_e = \ell\delta_r/\sqrt{2}$.

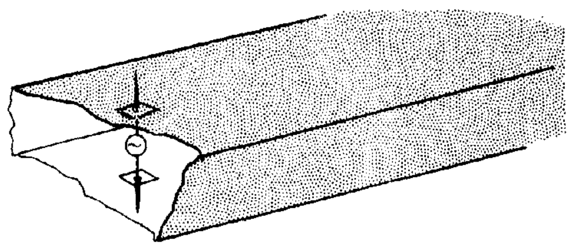
An operating frequency of 300 kHz in rock having a conductivity $\sigma = 0.1$ mho/m, gives a δ_r value of about 2 meters which results in an effective loop height of 1.4 m. The most effective long wire antenna length, ℓ, may range from slightly longer than the waveguide (i.e., seam) height as in the Sanguine case, to a length that is on the order of the effective skin depth, δ, of the coal seam waveguide, where δ is largely determined by the conductivity of the coal as opposed to that of the rock. The effective skin depth δ of lossy coal seam waveguides at 300 kHz has been found[25] to vary from about 10 m in high loss seams to as large as 100 m in low loss seams. Maximum wire lengths, ℓ, between 10 and 100 m can therefore be expected. Thus, the effective loop areas of grounded long wires are seen to compare favorably with that of a 2 m x 7 m rectangular fixed station loop.

Yet to be determined is whether the resultant waveguide coupling factor for the long wire "loop" embedded in the rock is more or less favorable than indicated by the above simplified analysis, whether termination impedances comparable to the impedances of tuned loops can be practically achieved and easily maintained over time, and to what extent the signal strength remains proportional to wire length as the length is increased. Both analytical and experimental efforts will be required to resolve these issues.

Vertical Grounded Wires

The Sanguine ELF radio system unfortunately could not use what is perhaps the ideal antenna to excite the quasi-TEM mode in the earth-ionosphere waveguide, namely a vertical current source rising out of the conducting earth and terminating in the conducting ionosphere. Although such an antenna is clearly not feasible for the Sanguine application, it does become a very real possibility for use in the coal seam waveguide. In its simplest form it could consist of a vertical center-driven wire, or a steel roof support, grounded at each end to two vertical roof bolts or other type of grounding connection, one in the roof and one in the bottom (floor), in a mine entry or crosscut tunnel as depicted in Figure 14. It also has the advantage of producing omnidirectional coverage in the plane of the coal seam, without the addition of a second unit.

FIGURE 14
GROUNDED VERTICAL WIRE ANTENNA
INSTALLED IN A COAL MINE TUNNEL



A first order estimate of the performance of this vertical wire/rod terminated source has been made and compared[16] to that of a representative vertical planar/loop. The vertical monopole-type source current generates a ϕ-directed magnetic field component that is independent of ϕ. Solution of the wave equation for this source distribution gives the following expression for the magnitude squared of H_ϕ:

$$|H_\phi|_V^2 = I^2 \frac{(\alpha^2+\beta^2)^{1/2}}{8\pi} \frac{e^{-2\alpha r}}{r} \quad (7)$$

where $r \gg \delta$, δ is the effective skin depth in the coal seam waveguide, I is the source current, α is the attenuation constant, and β is the phase constant. Ideal matching of the source current distribution to that required in the rock by the quasi-TEM mode is assumed. The corresponding field expression for a dipole source current such as that of a rectangular planar loop is:

$$|H_\phi|_L^2 = \frac{M^2}{(h+\delta_r)^2+\delta_r^2} \frac{(\alpha+\beta)^{3/2}}{8\pi} \frac{e^{-2\alpha r}}{r} \quad (8)$$

for $r \gg \delta$, where M is the source loop magnetic moment, δ_r is the skin depth in the rock, and h is the height of the coal seam. Forming the ratio of these two field quantities gives the expression:

$$R_{V/L} = \frac{I^2}{M^2} \frac{(h+\delta_r)^2+\delta_r^2}{(\alpha^2+\beta^2)} \quad . \quad (9)$$

The ratio $R_{V/L}$ was computed for installations operating at a frequency of 250 kHz in a representative high-loss coal seam having a seam height of 3 m and a low loss seam of 1.5 m height. A current of 1 ampere was used to drive both the vertical wire and the vertical planar loop (a 2 m x 7 m one-turn loop of $M = 14$ A-m^2). In the high loss seam the ratio $R_{V/L} = 43$, represents a 16 dB advantage over the loop. In the low-loss seam the ratio $R_{V/L} = 132$, represents a 21 dB advantage. In the high-loss seam, a 16 dB advantage represents only about a 130 ft. increase in the receiver-noise-limited communications range, whereas a 21 dB advantage in the low-loss seam represents a more substantial increase in operating range of about 630 ft. at 250 kHz.

The spreading resistance of a 2 m long by 2.5 cm diameter ground rod (about the size of a roof bolt) making ideal contact along its length to rock of uniform conductivity 0.1 mho/m, is about 4 ohm, giving a total of 8 ohms for two terminations. When compared to the 2 m by 7 m vertical planar loop tuned impedance of about 2 ohms resistive, we find that the signal strength advantages of 16 dB and 21 dB are substantially reduced to 4 dB and 9 dB (corresponding to only 33 ft and 270 ft increases) in high and low loss seams, respectively, for a voltage source drive.

Need For In-Mine Measurements

The most effective fixed installation configurations from the standpoint of power drain, communication range, ease of installation and maintenance, compatibility of fixed station units with portable units, etc., will finally be determined after analyzing the results of a simple set of performance measurements in mines. Attention should be given to: determining the dependence of received signal strength on the length of the vertical planar loop and horizontal long wire mode exciters, checking the favorable theoretical results predicted for the vertical wire mode exciters, and determining the dependence of both signal strength and antenna input impedance on the spacing between parallel mode exciters. More importantly, the levels and variability of the impedances of roof bolt and other types of grounding terminations for vertical and horizontal wire mode exciters should be given particular attention.

The quasi-TEM coal seam mode of propagation, the seam thickness, and the mode effective skin depth may impose limits on the increase in signal strength obtainable by increasing the length of either the vertical planar loop or horizontal long wire mode exciters. Namely, there may be a readily identifiable point of diminishing return associated with continuous increases in length.

The elaborate ground termination techniques available to a large, permanent, well-controlled Sanguine antenna installation will, of course, not be practical for a semi-permanent and largely uncontrolled mine antenna installation. As a result, ground termination impedances achievable in a mine may be so large and/or variable as to make the horizontal long wire and/or vertical wire mode exciters impractical for mines. Although the above computed 8 ohms is reasonable, it may not be achievable in practice. In fact, impedances on the order of 100 ohms are more common with conventional roof bolts. At present, quantitative knowledge on how roof bolt and other types of grounding terminations work, vary over time, should be installed, and could be improved from an electrical standpoint, is not available. If and when this information becomes available, as a result of carefully performed in-mine measurements, a more definitive assessment of the feasibility of the long horizontal wire and vertical rod mode exciters will be possible.

CONCLUSION

The choice of a specific portable antenna should be based on the overall efficiency and practicality achievable by the total transmitter-antenna *system* in producing the largest usable signal at the desired ranges, within the practical constraints of system overall size, weight, battery life, convenience of use, intrinsic safety, and ruggedness for use by roving miners. Conventional air-core bandolier loop antennas, and perhaps smaller ferrite-loaded loop antennas, are the most suitable and reasonable choices for roving miner portable radio applications at frequencies below about 1 MHz. Furthermore, a transmit moment of about 2.5 amp-m^2 (peak) appears to be a practical upper bound for intrinsically safe portable units. For fixed installations, horizontal-wire and vertical-rod mode exciters may also offer comparable or better performance than vertical planar air-core loops, but in-mine measurements will be needed to resolve the matter.

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COMMENT BY R. GABILLARD

An effective way to get good electrical contacts in mines is to drill a hole in the ceiling, wall, or floor (about 1 meter depth) and to force in it an ordinary non-corrosive iron tube wrapped with rags (one must add salt to the rags). The moisture of rocks make the rags always wet and the electric contact is insured on all the surface of the iron tube. Earth resistance as low as 10 Ω may be obtained.

COUPLING OF THE COAL-SEAM MODE TO A CABLE IN A TUNNEL AT MEDIUM RADIO FREQUENCIES

R.L. Lagace and A.G. Emslie

Arthur D. Little, Inc.
Cambridge, MA 02140

ABSTRACT

The coal-seam mode is a transmission-line mode of propagation of medium frequency radio waves in a low conductivity coal seam which is bounded above and below by relatively high conductivity rock, with the electric field vertical and the magnetic field horizontal. When the mode is excited by a vertical loop antenna, the magnetic field falls off with distance with cylindrical spreading and exponential decay factors. However, when a long conductor such as a power cable is present in a nearby tunnel, the field is found experimentally, after a certain distance, to level out to a relatively insignificant decay rate. This effect is clearly due to excitation by the coal-seam mode of a low-attenuation mode guided by the cable. The field of this mode, at points along a measurement path parallel to the cable replaces that of the original coal-seam mode. The magnetic field of the cable-guided mode for a given current in the cable is calculated by image theory using a Fourier series to represent the infinite set of images. The voltage induced in a loop located at some horizontal distance away from the cable is next calculated from the magnetic field at that point. The principle of reciprocity is then used to find the current induced in the cable when the loop acts as the transmitter. The theoretical coupling to the cable is compared with some experimental data taken in a mine.

INTRODUCTION

The lowest order coal-seam mode is a parallel-plate TEM transmission-line type of propagation[1] of medium frequency radio waves in a conducting coal seam which is bounded above and below by more conductive rock, with the electric field vertical and the magnetic field horizontal. This mode can be excited by a vertical loop antenna placed in a tunnel in the coal seam. In a conductor-free region of the coal mine the magnetic field falls off with distance with a cylindrical spreading factor multiplied by an exponential loss factor. However, when a conductor, such as a power cable, is present in a parallel tunnel, the field is found experimentally[2] to level off to a much lower rate of decay after a certain fairly well defined distance.

The effect can be attributed to coupling of the coal-seam mode to a low attenuation mode guided by the cable. Experiment shows[2] that a slowly decaying current is indeed present in the cable. The purpose of this paper is to investigate the nature of the coupling. The method is to start with a current in the cable, calculate the magnetic field produced by this current at the location of the loop antenna, determine the voltage induced in the loop by this field, and finally use the reciprocity principle to determine the current induced in the cable when the loop acts as a transmitter.

MAGNETIC FIELD DUE TO CURRENT IN CABLE

Figure 1 shows the cable at a distance y = d from the center of the coal seam. The cable carries a current I which produces an infinite set of images[3] with current

$$I_n = (-1)^n I \qquad (1)$$

located at the (complex) positions:

$$y_n = n\,(h + \delta_r - i\delta_r) + (-1)^n d \qquad (2)$$

where $\delta_r = (\pi\mu_0 f\sigma)^{-1/2}$ is the skin depth in the rock. The images, including the cable itself, form a periodic distribution of current of period:

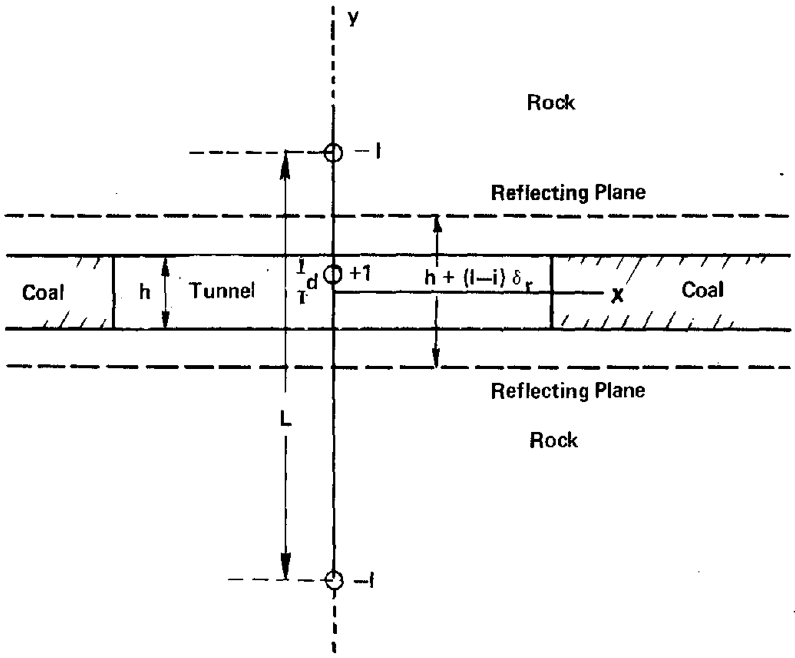
$$L = 2(h + d - i\delta_r) \qquad (3)$$

To calculate the magnetic field due to the set of images it is convenient to represent the discrete currents by a Fourier series:

$$I(y) = \sum_{1}^{\infty} \left(A_p \cos \frac{2\pi py}{L} + B_p \sin \frac{2\pi py}{L} \right) \qquad (4)$$

This work was supported by the U.S. Department of the Interior, Bureau of Mines, Pittsburgh Mining and Safety Research Center, under USBM Contract HO346045.

FIGURE 1
CABLE WITH CURRENT (+ I) IN
COAL MINE TUNNEL WITH IMAGES (– I) IN ROCK



which contains no constant term since the total current in cable and rock is zero. For the currents defined by (1) and (2) the Fourier series takes the form:

$$I(y) = \frac{4I}{L}\left(\cos\frac{2\pi d}{L}\cos\frac{2\pi y}{L} + \sin\frac{4\pi d}{L}\sin\frac{4\pi y}{L}\right.$$

$$+ \cos\frac{6\pi d}{L}\cos\frac{6\pi y}{L} + \sin\frac{8\pi d}{L}\sin\frac{8\pi y}{L} \tag{5}$$

$$\left. + - - - - - \right)$$

The magnetic fields corresponding to the first harmonic of the yz current sheet, for a wave propagating along the cable, are readily found to be:

$$H_x = \frac{ik_y}{k_x}\frac{2I_o}{L}\cos k_y d e^{-ik_x x} e^{-ik_z z} \sin k_y y \tag{6}$$

$$H_y = \frac{2I_o}{L}\cos k_y d\, e^{-ik_x x} e^{-ik_z z} \cos k_y y \tag{7}$$

$$H_z = 0 \tag{8}$$

$$k_y = \frac{2\pi}{L} \tag{9}$$

where I_o is the current in the cable at z = 0.

For the second harmonic, the field is:

$$H_x = -\frac{ik_y}{k_x}\frac{2I_o}{L}\sin k_y d e^{-ik_x x} e^{-ik_z z}\cos k_y y \tag{10}$$

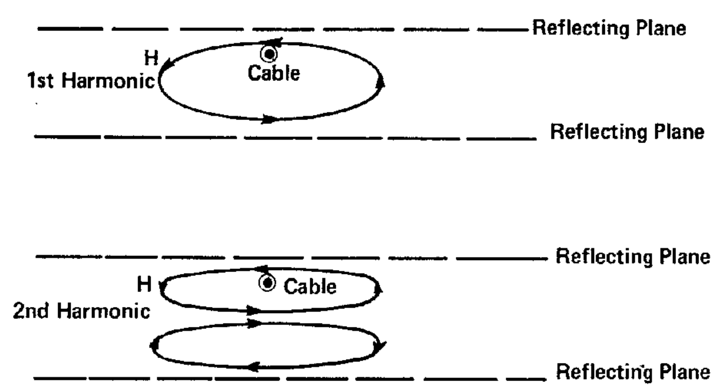
$$H_y = \frac{2I_o}{L}\sin k_y d e^{-ik_x x} e^{-ik_z z}\sin k_y y \quad . \tag{11}$$

$$H_z = 0 \tag{12}$$

$$k_y = \frac{4\pi}{L} \tag{13}$$

The magnetic field lines for the two harmonics are shown in Figure 2.

FIGURE 2
MAGNETIC FIELD LINES FOR TWO HARMONICS



The k's are related by the formula:

$$k_x^2 + k_y^2 + k_z^2 = k_o^2 \tag{14}$$

where k_o is the free-space propagation constant and k_z is the z-directed propagation constant for the cable-guided wave. For medium frequencies, k_z is not very different from k_o[4,5] ($\sim .75k_o$) and both are small in magnitude compared with k_y. Therefore, Equation (14) can be written to a good approximation as:

$$k_x \cong -ik_y \, . \tag{15}$$

If we make the further approximation that $|k_y d| \ll 1$, and let y = 0, we obtain the simple results:

$$\left|\frac{H_x}{I_o}\right|_1 = 0 \tag{16}$$

$$\left|\frac{H_y}{I_o}\right|_1 = \frac{1}{\sqrt{(h+\delta_r)^2 + \delta_r^2}} \, e^{-\frac{\pi (h+\delta_r) x}{(h+\delta_r)^2 + \delta_r^2}} \tag{17}$$

$$\left|\frac{H_x}{I_o}\right|_2 = \frac{2\pi d}{(h+\delta_r)^2 + \delta_r^2} \, e^{-\frac{2\pi (h+\delta_r) x}{(h+\delta_r)^2 + \delta_r^2}} \tag{18}$$

$$\left|\frac{H_y}{I_o}\right|_2 = 0 \tag{19}$$

where the subscripts 1 and 2 refer to first and second harmonics, respectively.

EXCITATION OF THE CABLE BY A TRANSMITTING LOOP ANTENNA

By the principle of reciprocity, we can use the above results to find the current I_c induced in the cable by a loop antenna situated at a distance x from the cable. The result is:

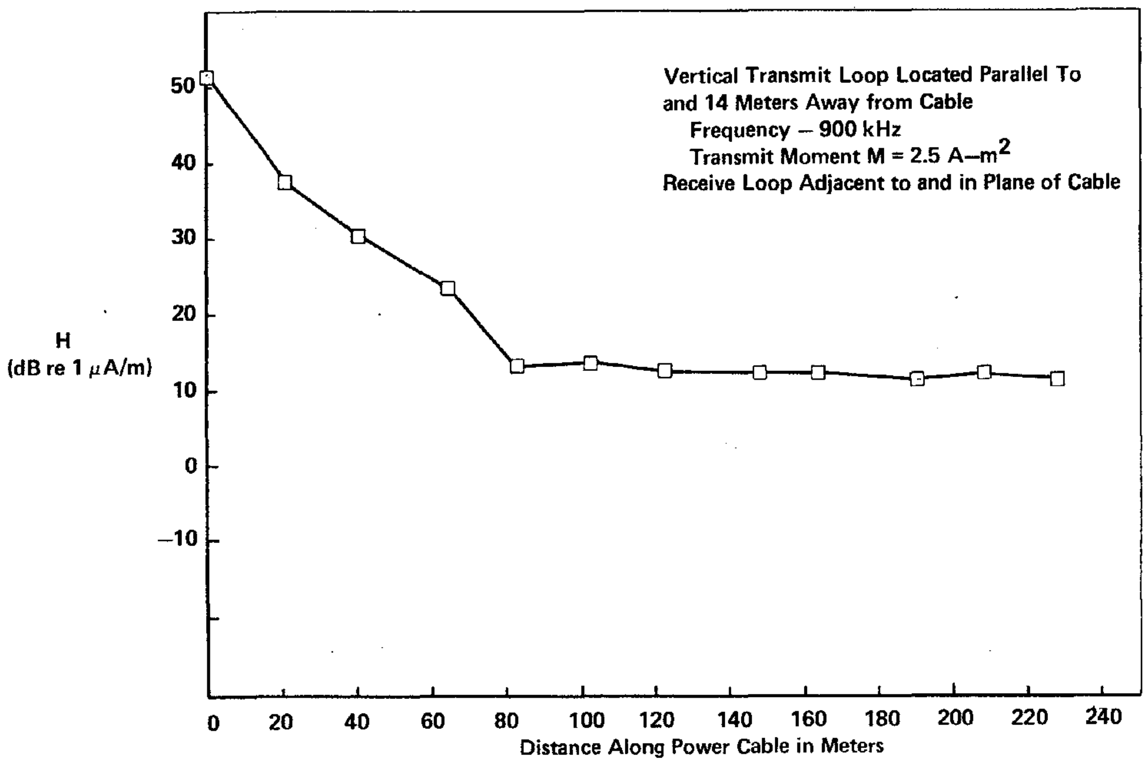
$$I_c = \frac{\omega \mu_o M}{2Z_o} \frac{H}{I_o} \tag{20}$$

where M is the magnetic moment of the loop, Z_o is the characteristic impedance of the cable transmission line, and (H/I_o) is the appropriate ratio from Equations (16 – 19).

Measurements by T.S. Cory[2] in the Helvetia Coal Co. Margaret No. 11 mine with a vertical transmit loop parallel to, and placed at a distance of 14 meters from, a power cable are shown in Figure 3. The first part of the magnetic field plot represents the decay of the incident field of the coal-seam mode. Beyond about 80 m along the cable, the measured field is due to the current induced in the cable. This current is estimated from the data, and from the size of the receiving loop located adjacent to and in the plane of the cable, to be about 6×10^{-6} A. From other measurements taken in a conductor-free area of the same mine, we can infer that the conductivity of the rock, σ_r, is in the range 0.01 to 0.02 Mho/m. For σ_r = 0.01 Mho/m, the skin depth in the rock is $\delta_r = 5.3$ m.

If the transmitting loop is oriented vertically, so as to generate the TEM coal-seam mode with the magnetic field horizontal, it is seen from (16) and (18) that coupling to the cable occurs only via the second and higher-order even harmonics. Using the above value of skin depth, δ_r, and the experimental values d = 0.45 m, h = 1.3 m, and x = 14 m, theoretical estimates of the current induced in the cable are obtained from Equation (18) for the second harmonic. Equation (18) gives $|H_x/I_o|_2 = 1.2 \times 10^{-5}\ m^{-1}$. Equation (20), with M = 2.5 A - m^2 and $Z_o = 300\ \Omega$, then gives $I_c = 1.4 \times 10^{-7}$ A. Thus, the predicted current value for a vertically oriented transmit loop is about 50 times less than the experimentally determined value.

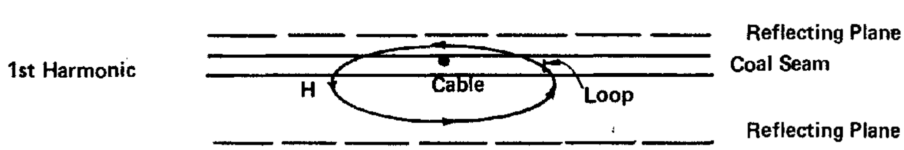
FIGURE 3
MAGNETIC FIELD VS. DISTANCE
ALONG POWER CABLE
MARGARET NO. 11 MINE, FREEPORT SEAM



This disagreement implies that the excitation of the cable-guided mode is not brought about by the TEM coal-seam mode. Rather, it is brought about by a higher-order mode, below cutoff in the coal seam at MF frequencies, which has circular lines of E about a vertical axis and an H field with both radial and vertical components. This type of mode is generated by a horizontally oriented transmitting loop which couples to the cable via the first harmonic which produces a nonzero y-component of magnetic field. Equation (17), which applies for this case, gives $|H_y/I_0|_1 = 2.06 \times 10^{-3}$. Equation (20) then gives $I_c = 6.1 \times 10^{-5}$ A, which is 10 times larger than the experimental value obtained for a vertical antenna. An accidental tilt of the antenna of about 6 degrees away from the vertical would therefore make theory and experiment agree.

An alternative explanation of the coupling of a vertical loop to the cable is that the rock conductivity above the coal seam is appreciably different from that below the seam. Under these conditions the two reflecting planes are not symmetrically located with respect to the coal seam. In that case, as shown in Figure 4, a vertical loop now couples to the first harmonic of the cable field. A vertical displacement of the loop with respect to the center of the tunnel, with symmetrically placed reflecting planes, also gives coupling to the first harmonic.

FIGURE 4
VERTICAL LOOP COUPLING TO 1ST HARMONIC WHEN COAL
SEAM NOT CENTERED BETWEEN REFLECTING PLANES



COMMUNICATION BETWEEN HORIZONTAL LOOP ANTENNAS IN THE PRESENCE OF CONDUCTORS

The foregoing analysis suggests that horizontally oriented loop antennas may provide an efficient communication system in areas where conductors such as power lines, trolley lines, and rails are present. The vertical component of magnetic field at a distance x_2 from a cable due to a horizontal transmitting loop at a distance x_1 from the same cable is from Equation (17) (used twice) and Equation (20),

$$H = \left(\frac{\omega\mu_0 M}{2Z_0}\right)\left\{\frac{1}{(h+\delta_r)^2+\delta_r^2}\right\} e^{-\frac{\pi(h+\delta_r)(x_1+x_2)}{(h+\delta_r)^2+\delta_r^2}} \tag{21}$$

Figures 5 and 6 show graphs of H versus $x_1 + x_2$ for various frequencies, for rock conductivities of 0.01 and 0.1 Mho/m. The horizontal bar on each line indicates the intrinsic receiver noise level at that frequency for a 12 kHz bandwidth. It is seen that the maximum value of $x_1 + x_2$ depends strongly on the conductivity and is 38 m for σ_r = 0.01 Mho/m, but only 18 m for σ_r = 0.1 Mho/m. The optimum frequencies are 300 kHz and 100 kHz, respectively. No allowance has been made for loss along the cable itself. The results are not valid for very small values of either x_1 or x_2.

FIGURE 5
VERTICAL MAGNETIC FIELD AT DISTANCE X_2 FROM CABLE
DUE TO HORIZONTAL TRANSMIT LOOP AT DISTANCE X_1
(Rock Conductivity σ_r = 0.01 Mho/m)

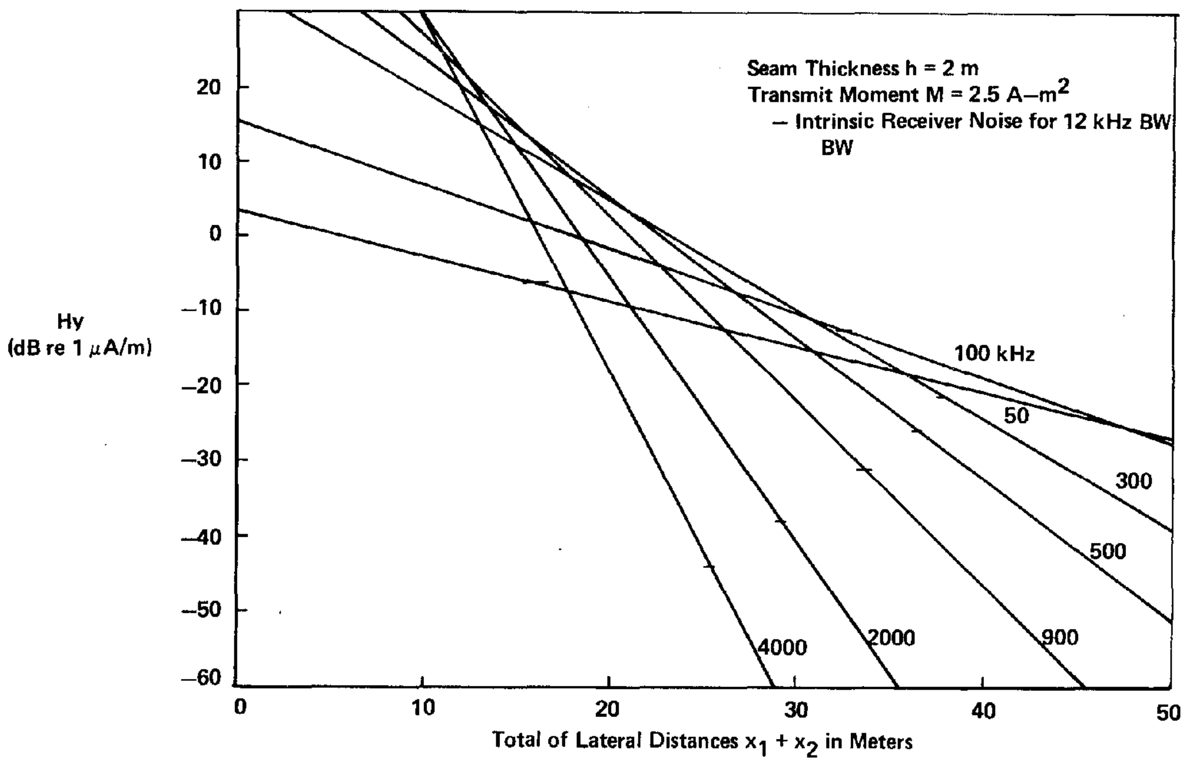
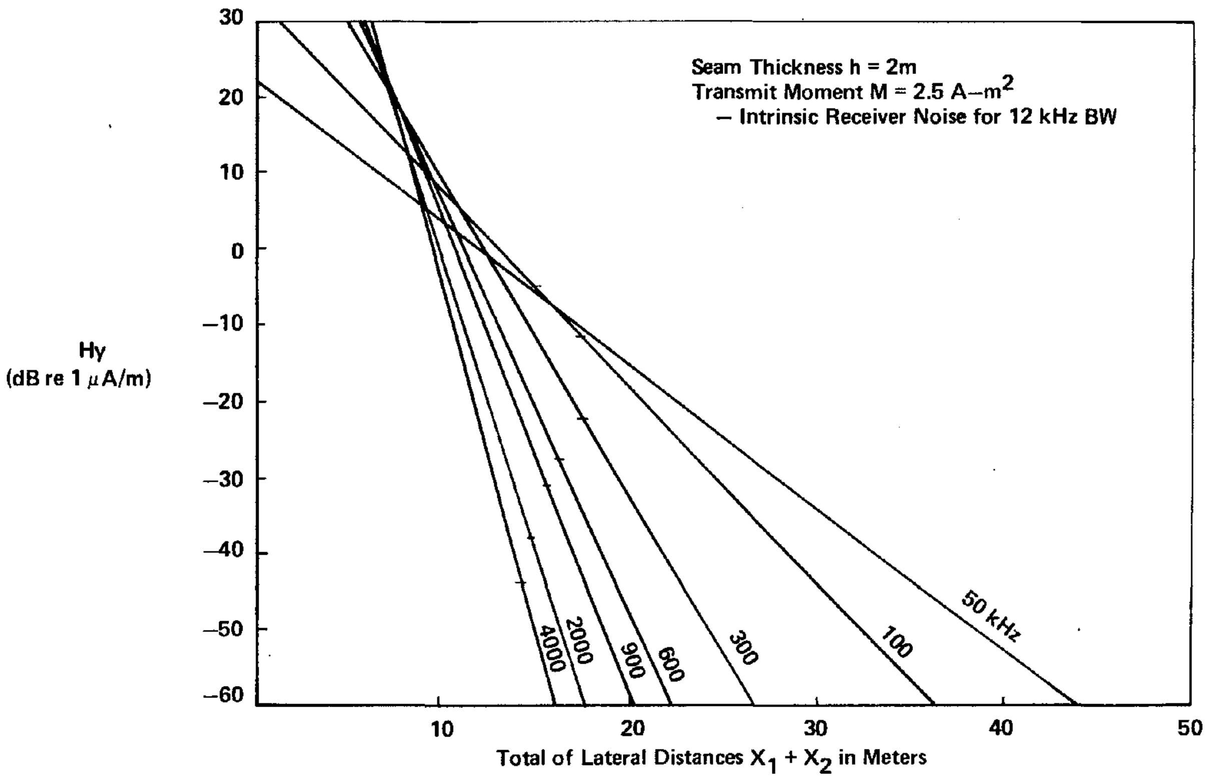


FIGURE 6
VERTICAL MAGNETIC FIELD AT DISTANCE X_2 FROM CABLE DUE TO HORIZONTAL TRANSMIT LOOP AT DISTANCE X_1
(Rock Conductivity σ_r = 0.1 Mho/m)



Figures 7 and 8 show similar plots, derived from (17) and (20), for the field at the cable itself, (as measured by a square coil of edge 0.67 m held with one edge alongside the cable) due to a horizontal transmitter loop at a distance x from the cable. The maximum distances for the two values of rock conductivities are 57 m and 22.5 m, and they occur at 100 kHz and 50 kHz, respectively.

CONCLUSION

A theoretical model for predicting the coupling behavior between loop antennas and conducting cables in coal mine tunnels is currently under development, and the results are comparing favorably with the limited amount of experimental in-mine data taken to date. Further refinements of this theoretical model, together with comparisons with more comprehensive data from two additional coal mines, will be performed in the near future.

FIGURE 7
MAGNETIC FIELD AT THE CABLE DUE TO
HORIZONTAL TRANSMIT LOOP AT DISTANCE X FROM THE CABLE
(Rock Conductivity σ_r = 0.01 Mho/m)

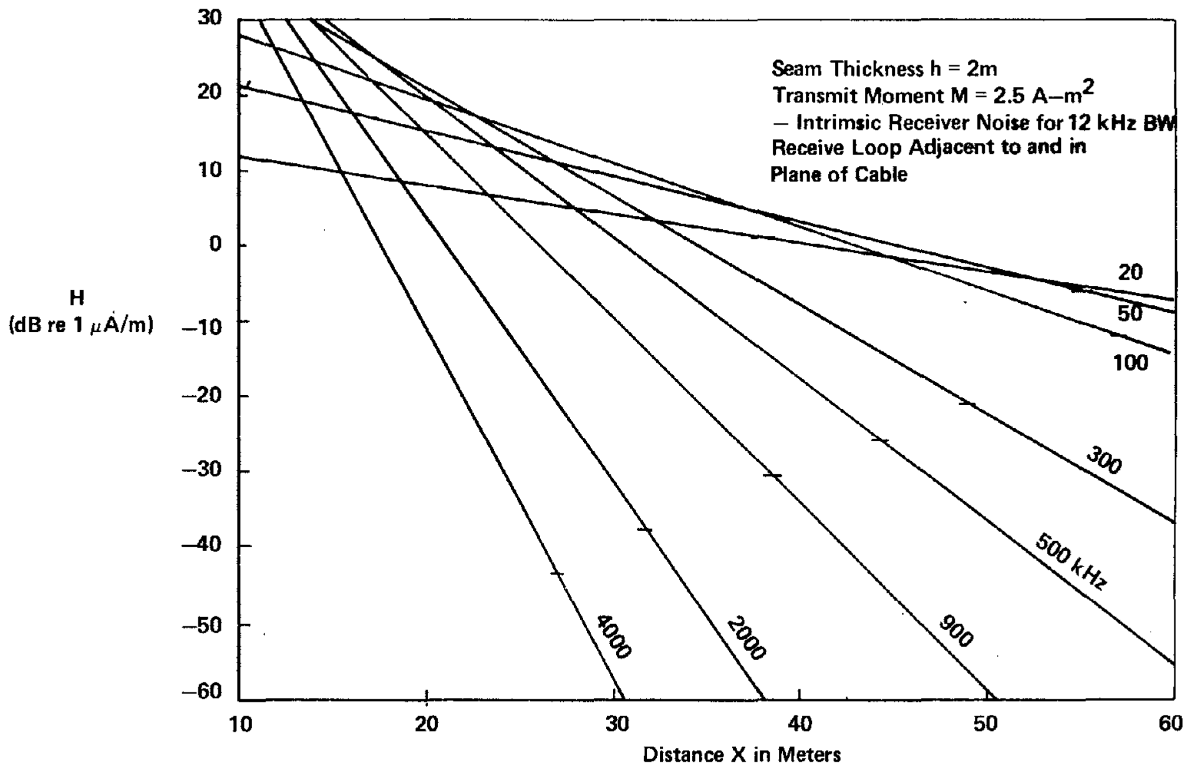
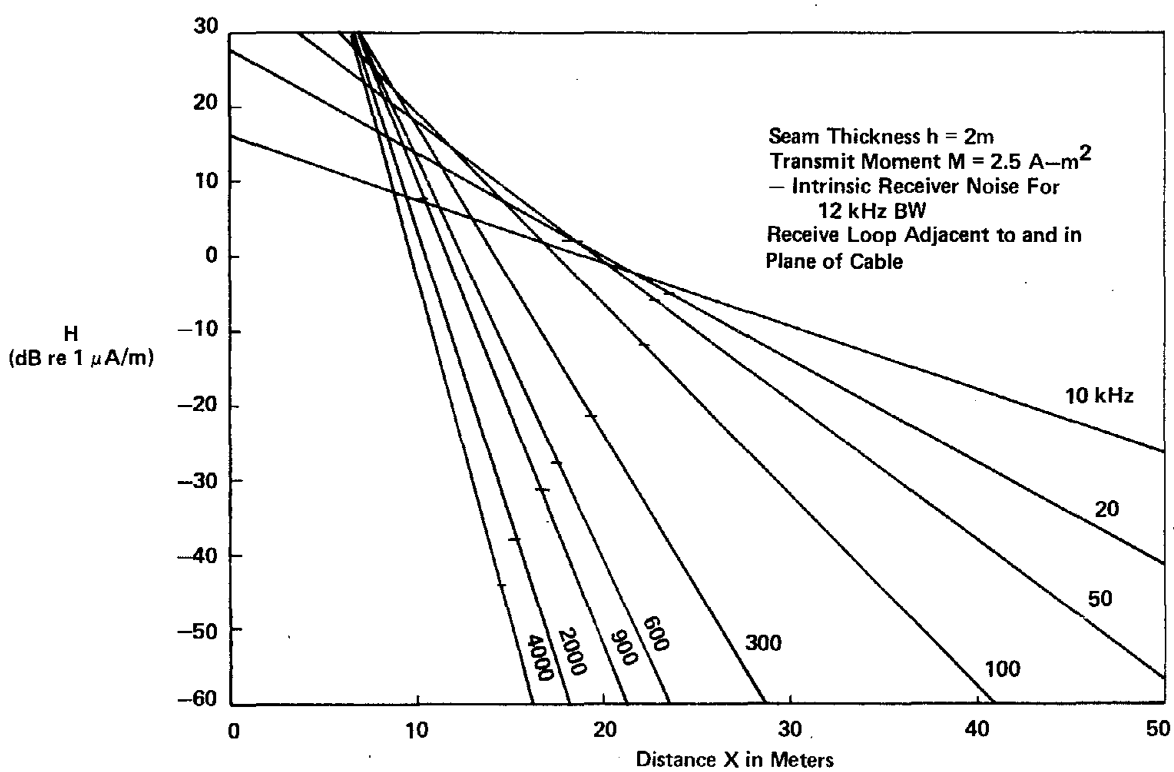


FIGURE 8
MAGNETIC FIELD AT THE CABLE DUE TO
HORIZONTAL TRANSMIT LOOP AT DISTANCE X FROM THE CABLE
(Rock Conductivity σ_r = 0.1 Mho/m)



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2. T.S. Cory, "Electromagnetic Noise and Propagation in Low-Coal Mines, Helvetia Coal Co. Margaret No. 11 Mine," Summary Data Report 1 to Collins Group, Rockwell International Corporation, under Bureau of Mines Contract HO377053.

3. P.R. Bannister, "Approximate Results for the Mutual Electromagnetic Coupling of Loops over a Homogeneous Ground," Naval Underwater System Center Report #NL-3029, New London Laboratory, 23 November 1970, NTIS No. AD717351.

4. J.R. Wait and D.A. Hill, "Guided Electromagnetic Waves Along an Axial Conductor in a Circular Tunnel," IEEE Trans. Antennas and Propagation, AP-22(4), pp627-630, July 1974.

5. D.A. Hill and J.R. Wait, "Pulse Transmission Along Cables in Circular Tunnels," Proceedings of U.S. Bureau of Mines Guided Wave EM Workshop, Transmission Session I, March 1978.

U.S. DEPARTMENT OF COMMERCE
Office of Telecommunications
INSTITUTE FOR TELECOMMUNICATION SCIENCES
Boulder, Colorado 80302
Room 242, RB 1

10 October 1977

COMMENTS BY J.R. WAIT

(The first two pages of an earlier letter from J.R. Wait is included here since it bears on the previous two papers and the ensuing discussions).

Mr. Robert L. Lagace
Research & Development Division
Arthur D. Little, Inc.
Acorn Park
Cambridge, MA 02140

Dear Bob:

With great interest I read your report on "Transmit Antennas for Portable VLF to MF Wireless Mine Communication", ADL-77229, Task C, May 1977 coauthored with D.A. Curtis, J.D. Foulkes and J.L. Rothery (USBM Contract No. H0346045). I can see you have drawn on a great deal of information from many sources to put this together. I can appreciate the great deal of work that went into it.

As you know, I have had a long interest in the basic aspects of antennas immersed in conducting media. It is possible you saw my five chapters in the two volume book "Antenna Theory" edited by Zucker and Collin and published by McGraw-Hill in 1969. The chapter 24 discussed some of the same topics found in the first part of your report. As you indicate, the neglect of displacement currents is usually justified for rock conductivities greater than 10^{-2}mhos/m and for frequencies less than about 1 MHz. However, when these conditions are violated or not well satisfied, the displacement currents can allow for modest increases of range.

The three layer model that you covered in your paper in the April 1976 issue of Radio Science has, of course, been used in related studies such as propagation in the earth-crust waveguide. In my note in the same issue, I pointed out that the earlier sub-surface layer formulations could be adapted directly to this coal seam situation without having to introduce "effective widths" of the seam waveguide. In any case, would not your b_e be complex? Even then it is only an approximation, particularly when the actual coal seam width is comparable with or less than the skin depth δ_r in the adjacent rock. Ramsey Decker should now have a general computer program available for the general layer formulations so detailed comparative calculations should be possible.

Your discussion of electrically small antennas was particularly interesting. However, in talking about efficiency, one of the important aspects for antennas near or within the earth is the component of the input resistance due to the eddy currents excited in the earth by the induction fields. Don Watt and I refer to this as the H-field loss. It greatly dominates R_d the loss associated with the antenna conductors. To minimize this loss, huge ground systems consisting of tons of copper wire were laid down at Cutler Maine (NAA). This is a project I was closely involved in as you can see in Don Watt's book and his references. At this time (in the middle Fifties), we also became aware of the so-called E-field loss which (viewed in hindsight) is obviously the effect of ohmic dissipation with current flowing vertically across the air ground interface. In a sense, it is additive to the H-field loss. I believe this E-field or electrostatic loss is the major component of the loss for an antenna located in a mine tunnel even for frequencies as high as a few MHz. Dave Hill and I discussed this point in our paper also in the April 1976 special issue. More recently, we had considered analytical models for this effect in tunnels that also contain axial conductors (you should have received copies of all of our Preliminary Reports on this topic)*.

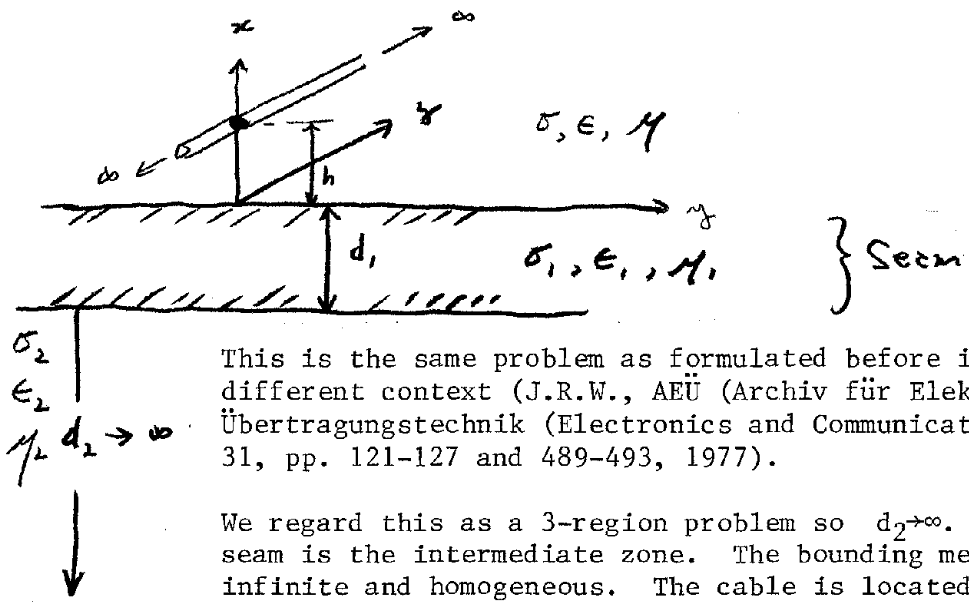
While I can see the logic in defining a comparison factor $F_{d,\ell}$ as you have done on page 47, I believe the conclusions could be modified when you bring in the different input resistance values for loops and whips. If anything, this will make the loop look even more attractive. Sometime ago (i.e. 1952, Proc. IRE paper), I made a comparison of insulated magnetic dipoles (loop) and electric dipoles (whips) in a conducting medium. The material is also covered in chapter 24 of the Collin/Zucker book. As indicated, the basic difference is that the near field of a magnetic dipole is largely reactive for a conducting medium whereas for an electric dipole the near field is largely resistive. For a given volume to locate the antenna, the unqualified conclusion is that small insulated loops are vastly superior to small electric dipoles as far as power effeciencies are concerned.

I thought your discussion of the relevancy of the project Sanguine work was interesting. In this connection, you might be interested to know that M.L. Burrows has just published a book on ELF antennas. It should be available shortly from the Peter Peregrinus Co. Ltd. in England. You referred to the 1974 special 200 page issue of the IEEE Comm. Trans. that contained many different approaches to the Sanguine problem. Of course, the Navy could not use vertical grounded wires but actually this type of excitation had been suggested for the earth-crust waveguide where the deep drill hole was to contain the vertical radiating structure (e.g. see the 1971 A.G.U. Monograph No. 14 edited by J.G. Heacock of the ONR). The earth-crust waveguide project has largely petered out due to the lack of success in finding highly resistive sub-surface layers. I will tell you all about this sometime if you are interested.

*J.R. Wait and D.A. Hill, "The Impedance of Dipoles in a Circular Tunnel with an Axial Conductor", Preliminary Report issued on 11 April 1977, also see April 1978 issue of IEEE Transactions on Geoscience Electronics.

FURTHER COMMENT BY J.R. WAIT:

THE COAL SEAM MODE(S) FOR A PARALLEL CABLE - A POSSIBLE THEORETICAL APPROACH



This is the same problem as formulated before in a slightly different context (J.R.W., AEÜ (Archiv für Elektronik und Übertragungstechnik (Electronics and Communication), Band 31, pp. 121-127 and 489-493, 1977).

We regard this as a 3-region problem so $d_2 \to \infty$. The coal seam is the intermediate zone. The bounding media are semi-infinite and homogeneous. The cable is located at a distance h (i.e. x=h) from the slab in one of the boundary media. Later we can let $h \to 0$. The formulation for the case when h is negative would be similar but we exclude this case here. The mode equation for the discrete modes can be written $T(\beta) = 0$ where $i\beta$ is the axial (z directed) propagation constant) and where

$$T(\beta) = 4\pi(\sigma+i\varepsilon\omega)Z_a(\beta) - (k^2-\beta^2)\{2K_o[i(k^2-\beta^2)^{\frac{1}{2}}c] + \int_{-\infty}^{+\infty} R_a(\lambda,\beta)e^{-2uh}u^{-1}d\lambda\}$$

$u = (\lambda^2 + \beta^2 - k^2)^{\frac{1}{2}}$, where the time factor is $\exp(i\omega t)$. $Z_a(\beta)$ is the series impedance of the cable per unit length; it may be spatially dispersive. $ik = [(i\mu\omega)(\sigma+i\varepsilon\omega)]^{\frac{1}{2}}$ is the (intrinsic) propagation constant in the upper region. If this is free space then, of course, $k = (\varepsilon_o\mu_o)^{\frac{1}{2}}\omega$.

In the preceding,

$$R_a(\lambda,\beta) = \frac{u^2\beta^2R_e + k^2\lambda^2R_m}{(k^2-\beta^2)(\lambda^2+\beta^2)}$$

$$R_e = \frac{K - Z_1}{K + Z_1}, \quad R_m = \frac{N - Y_1}{N + Y_1}, \quad Z_1 = K_1\frac{K_2 + K_1\tanh u_1d_1}{K_1 + K_2\tanh u_1d_1},$$

$$Y_1 = N_1\frac{N_2 + N_1\tanh u_1d_1}{N_1 + N_2\tanh u_1d_1}, \quad u_1 = (\lambda^2+\beta^2+\gamma_1^2)^{\frac{1}{2}}, \quad \gamma_1^2 = i\mu_1\omega(\sigma_1+i\varepsilon_1\omega),$$

$$N = \frac{u}{i\mu\omega}, \quad N_1 = \frac{u_1}{i\mu_1\omega}, \quad K = \frac{u}{\sigma + i\varepsilon\omega}, \quad K_1 = \frac{u_1}{\sigma_1 + i\varepsilon_1\omega},$$

$$N_2 = \frac{u_2}{i\mu_2\omega}, \quad K_2 = \frac{u_2}{\sigma_2 + i\varepsilon_2\omega}, \quad u_2 = (\lambda^2+\beta^2+\gamma_2^2)^{\frac{1}{2}}, \quad \text{and}$$

$$\gamma_2^2 = i\mu_2\omega(\sigma_2+i\varepsilon_2\omega)$$

Special cases of this solution (e.g. for homogeneous half-space) are given in J.R. Wait, Can. J. Phys., Vol. 50, 2402-2409, 1972; R.G. Olsen and D.C. Chang, Radio Science, Vol. 11, 875-884, 1976; E.F. Kuester and D.C. Chang, RADC Tech. Rept., No. 1, Electromagnetic Lab. CU, 1976, etc.

Here we need to examine the modal solutions which can be classified as coal seam modes and cable modes. It is true the dominant cable mode should be the one of lowest attenuation at the longer distances from the source, particularly if the cable is insulated. The excitation could also be treated explicitly by examining the residue of the cable and seam modes as in the 1977 AEÜ papers cited above. (The later one was distributed to workshop participants).

Wireless Radio Transmission at Medium Frequency in Underground Coal Mines - Summary of Measurements and Expected System Propagational Effects

By

Terry S. Cory, P.E.

ABSTRACT-

Wireless radio transmission range at medium frequency in underground coal mines has been demonstrated to exceed expectations through homogenious earth. Modeling of the cylindrical spreading of a quasi-TEM mode between parallel conducting planes has been found to predict observed results in many situations. This paper reviews the measurement research to date and describes the data base of mf field strengths in several high-production coal seams. The measurement technique is described along with a summary of typical measurement environments. Typical curves of measured field strength vs frequency and range for each seam and typical field strength maps illustrating area coverage in working mine topologies are given. Observed field strength variabilities with mine topology and along mine wiring conductors are briefly presented and discussed. A method of converting the field strength data to estimated maximum communication range for assumed values of transmit NIA, mine noise, and receiver sensitivity is described. Summary curves of this range illustrating important wireless transmission effects both in quasi-conductor-free and in conductor proximity areas of underground coal mines are presented.

I HISTORICAL BACKGROUND & RESULTS SUMMARY

In 1974, the U.S. Bureau of Mines PMSRC Communications Group @ Bruceton perceived a need for mine-wireless communications in coal mines for use in sections, forward of the section phone. The maximum transmission range expected to be required in these forward and largely conductor-free areas was 1300 feet (400 meters), providing for the diagonal path across a developing set of main headings of up to 12 entries wide.

Theoretical calculations assuming homogenious earth transmission, untested at that time, predicted communication ranges far short of 400 meters; again using unsubstantiated estimates for conductivity and dielectric constant. In August of 1974, a team of PMSRC personnel using newly procurred ECAM (South African made) AM portable radio transceivers (10-20 watt range) obtained a range of 1400 feet (427 meters) at the operating frequency of 335 KHz in a conductor-free area of Consolidation Coal Ireland Mine near Moundsville, W. Va. Based on this data, the Bureau entered into a contract (H034067) with Collins Radio Co. to define and develop prototype mine wireless portable and base station radios for coal mines. A portion of this contract involved in-mine measurements at medium frequency to permit selection of the operating frequency for these radios and to further confirm this previous test data.

Measurements of magnetic field intensity vs range and frequency by the author (via second tier subcontract with Collins through Spectra Associates Inc., of Cedar Rapids, Iowa) in January and February of 1975 in Consolidation Coal's Ireland and Rose Valley mines in the Pittsburgh seam. These results confirmed the previously observed communication range.

The results also exposed the optimum orientation of antennas to be coplanar and HMD (vertically oriented loop plane), with VMD orientations producing much weaker coupling. In concert with Arthur D. Little Inc.

the "waveguide" propagation mode at medium frequency was identified with ADL developing a theoretical model for evaluation of the measurements. Collins used the measured results in a system analysis, identifying the decrease in transmission loss with increase in frequency up to a frequency of about 500 KHz. Due to restricted capabilities of the transmitting measurement set-up, measurements at 1000 KHz and above could not be made. Collins selected a frequency of 520 KHz for use in their prototype radios; judged as being near optimum, yet just below the AM broadcast band.

Encouraged by the results of the testing, the Bureau entered into a contract with Collins (H0366028) specifically to perform similar measurements in several seams in six mines with ADL performing data evaluations and computer modeling under separate contract to the Bureau. The measurements were again performed by the author under a second tier subcontract to Collins via Spectra Associates, Inc.. The six mines consisted of Consolidation Coal's Ireland and #95 mines and Eastern Associated Coal's Federal #1 mine in the Pittsburgh seam, Inland Steel Coal Mine #1 and Peabody Coal Mine #10 in the Herrin #6 seam, and Island Creek's V.P. #1 mine in the Pocahontas #3 seam. The measuring equipment complement was improved to provide a dynamic measurement range coincident with the range for the ECAM and Collins prototype radios, and to increase the measurable frequency range to 4-5 MHz. Emphasis was placed on characterizing quasi-conductor-free propagation with the collection of conductor-proximity data being adjunctive except for area coverage magnetic field strength mapping of operating working section panels. Whenever possible, the ECAM and Collins prototype radios were taken into the test mine and used in conjuction with the field strength testing. By the time the measurements under this contract were completed, ADL had carefully estimated the value of the complex propagation constant from their three-layer computer model and, hence, estimates of coal seam and overburden/underburden conductivities were available. In particular, the predicted values of seam conductivity were low; of the order of 10^{-4} mho/meter.

The results presented in this paper are based on the data obtained under the above measurement contract H0366028.

In preparing the final report [1] for the above contract, the most useful data summary was determined to consist of maximum communication range curves vs frequency for each mine/seam embodying reasonable system operating assumptions. This summary analysis has shown that there is an optimum operating frequency for each mine/seam in conductor-free areas. At that point, it was felt to be beneficial to estimate the maximum communication range in the presence of conductors even though little quantitative data was available to substantiate specific computed results. Using the ADL derived attenuation and phase constants, scatter gain and conductor attenuation estimates were made using a simple model. Three measured data samples were available to compare with computed scatter gains; using these samples, agreement was within 3 dB. The optimum frequency phenomenon was again observed from the maximum communication range vs frequency summaries assuming conductor proximity for separation between at least one of the radios and the conductor(s) of at least one entry. The optimum frequencies so determined were higher than those for conductor-free operation, increasing with further separation of the radios from the conductor(s) up to about three entries in the case of the Pittsburgh seam. For close coupling (radios and conductor(s) in the same entry) the optimum frequency was not determined; the computed results exhibiting an increase in communication range with decrease in frequency down to the lowest frequency at which data for this situation was available (100 KHz). The attenuation rates used in the data reduction are in agreement with ADL estimates of 1-2 dB /km @ 100 KHz (2.2 dB/km @ 100 KHz computed for the Pittsburgh seam).

The measurements performed on contract H0366028 were largely in high coal (seam thicknesses of 60 inches or greater) and contained a fairly limited sampling of seams other than the Pittsburgh. To extend the measurements into low-coal and provide data characterizing a larger number of high production seams, the Bureau entered into a contract with Collins (H0377053) with measurements commencing in December, 1977. To date, measurements have been performed in four mines in three additional seams. These measurements have included systematically gathered samples in the presence of conductors to permit measured estimates of scatter gain. The mines tested include Helvetia Coal Margaret #11 mine and Upshur Coal's Adrian mine in the Upper Freeport seam, Bethlehem Steel Coal Mine #38D in the Lower Freeport seam, and Bethlehem Steel Coal Mine #31 in the Lower Kittanning seam. These new results will be forthcoming with preliminary indications that the communication ranges will be comparable to those of the Herrin #6 seam in all cases.

The results of measurements from contract H0366028 are summarized in Table I.

II MEASUREMENT SYSTEM DESCRIPTION

The measuring equipment was packaged for easy transportation and deployment in the mine. The equipment arrangement is illustrated in the block diagram of Figure I.

The transmitting equipment exclusive of antennas and battery packs was housed in two lightweight aluminum suitcases. One suitcase was dedicated to the HP-651B signal generator. The battery packs consisted of two groupings of four-each Gates 4.5 amp hr sealed lead acid cells, one wired to provide 12 VDC for the inverter and the other wired to provide 25 VDC for the ENI power amplifier. The ENI amplifier was broadband over the frequency range and required no tuning. The two antennas, a four-turn antenna for use above 1000KHz and a seven-turn antenna for use below 1000 KHz were L-section matched to the amplifier output impedance of 50 ohms at discrete frequencies corresponding to roughly 100, 230, 470, 900, 1850, and 4000 KHz. Later measurements at 100 KHz used a Pyott Boone portable trolleyphone operating into a loop antenna terminated with a 25 ohm resistor.

TABLE 1

SUMMARY OF RESULTS

- EXTENDED PROPAGATION RANGE (WRT HOMOGENIOUS EARTH) FOR PORTABLE RADIOS

  UP TO: 500 METERS IN PITTSBURGH SEAM
  300 METERS IN POCAHONTAS #3 SEAM
  140 METERS IN HERRIN #6 SEAM

- MOST MEASUREMENTS WERE SET-NOISE-LIMITED

  OPTIMUM FREQUENCY EXISTS FOR MAXIMUM COMMUNICATION RANGE IN CONDUCTOR-FREE-AREAS

  FOR PITTSBURGH SEAM: 400 KHZ SET-NOISE-LIMIT
  800 KHZ MEDIAN-MINE-NOISE-LIMIT

- SIGNIFICANT ENHANCEMENT IN COMMUNICATION RANGE EXISTS WHEN TRANSMITTER IS LOCATED WITHIN 2-3 ENTRIES OF LONG CONDUCTOR STRINGS; RECEIVER IN SAME ENTRY AS CONDUCTORS

  FOR PITTSBURGH SEAM: 1070 METERS, 1 ENTRY AWAY MEDIAN-MINE-NOISE
  620 METERS, 2 " " " " "
  360 METERS, 3 " " " " "

- ESTIMATES OF OPTIMUM FREQUENCY FOR CONDUCTOR-PROXIMITY SITUATIONS ARE GENERALLY HIGHER THAN FOR CONDUCTOR-FREE AREAS BASED ON SIMPLE MODEL ANALYSIS

  FOR PITTSBURGH SEAM: 700-800 KHZ 1 ENTRY AWAY MEDIAN-MINE-NOISE
  1050 KHZ 2 " " " " "
  2000 KHZ 3 " " " " "

- AREA COVERAGE OF AN ENTIRE 1100 METER BY 120 METER WORKING SECTION PANEL CAN BE PROVIDED FROM A TRANSMITTER LOCATED NEAR A BELTWAY HEAD OR TAIL PIECE WITH OPTIMUM FREQUENCY OF ABOUT 1 MHZ

TABLE 2

MEASUREMENT SYSTEM PARAMETERS

O FREQUENCY RANGE	83 - 4750 KHZ
O TRANSMIT NIA	0.44 - 2.04
NA	1.17 ABOVE 1000 KHZ
	2.04 BELOW 1000 KHZ
RESULTS NORMALIZED TO NIA = 2.5	
O RECEIVE SYSTEM SENSITIVITY OF NM-12 & NM-25	-25 DB > 1 uA/M @ 100 KHZ
	-37 DB > 1 uA/M @ 1000KHZ
	-47 DB > 1 uA/M @ 4750KHZ
O RECEIVE ANTENNA	0.73 M^2 SINGLE-TURN LOOP, COARSE TUNED

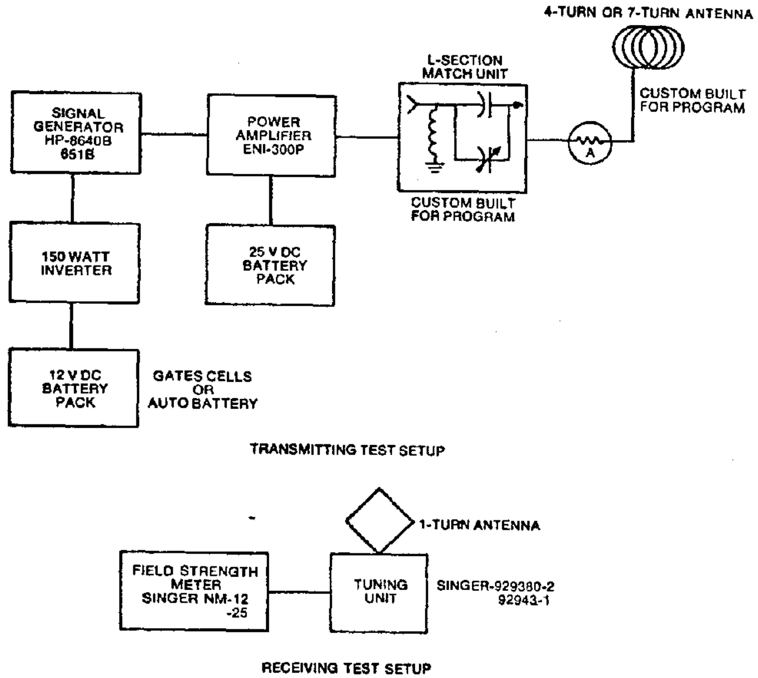


FIGURE 1

Measurement Test Equipment Configurations.

The receiving equipment consisted of two-each Singer NM-12's (up to 250 KHz) and two-each Singer NM-25's (above 250 KHz). These field strength meters were used with coaxial one-turn loop antennas with coarse tuning in the base sections.

The measurement system parameters are summarized in Table 2.

III PRESENTATION & EVALUATION OF TYPICAL RESULTS

Measured magnetic field strength data typifying quasi-conductor-free areas in the Pittsburgh, Pocahontas #3, and Herrin #6 seams are shown in Figures 2,3, and 4 respectively which illustrate frequency and range dependance. At close range in the Pittsburgh and Pocahontas #3 seams, the field strength "excitation" levels increase with frequency over the measurement range. In the Herrin #6 seam, these close-in levels increase with frequency up to about 1-2 MHz and then decrease. This latter tendancy has also been observed more recently in the new low-coal results. All results show an increase in range attenuation with frequency. The rate of this attenuation increase with frequency largely determines the optimum frequency per seam.

Some variation in measured field strength occurs with range according to the topological path taken. The attenuation with range at a given frequency is generally a little higher when the path is through or adjacent to a large coal block than it is along an entry having adjacent parallel entries on each side. Generally, the variation increases with frequency. Variations are illustrated in Figures 3a and 3b for paths along adjacent parallel entries with the transmitter located in the particular entry being traversed at the time. Not shown are paths oblique to the usual entry/crosscut grid. For these latter paths, in addition to the above observations, a periodic variation with range is superimposed on the attenuation curves with range; this variation being more pronounced at higher frequencies.

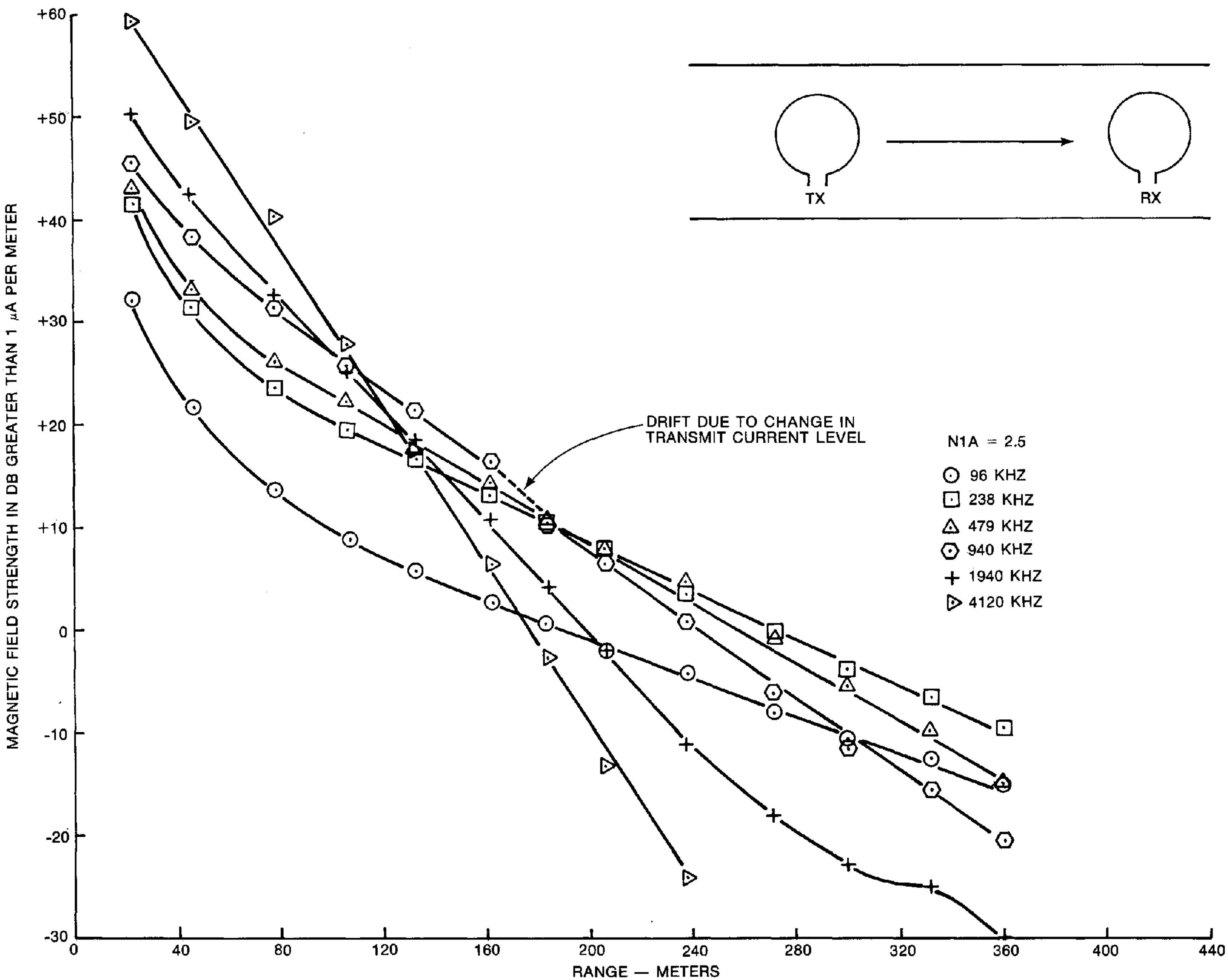


Figure . 2 Magnetic Field Strength Vs Range in Quasi-Conductor-Free 8 Main North Area of Eastern Associated Coal Federal No. 1 Mine (Eddy Portal) for Coplanar Orientation of Antennas.

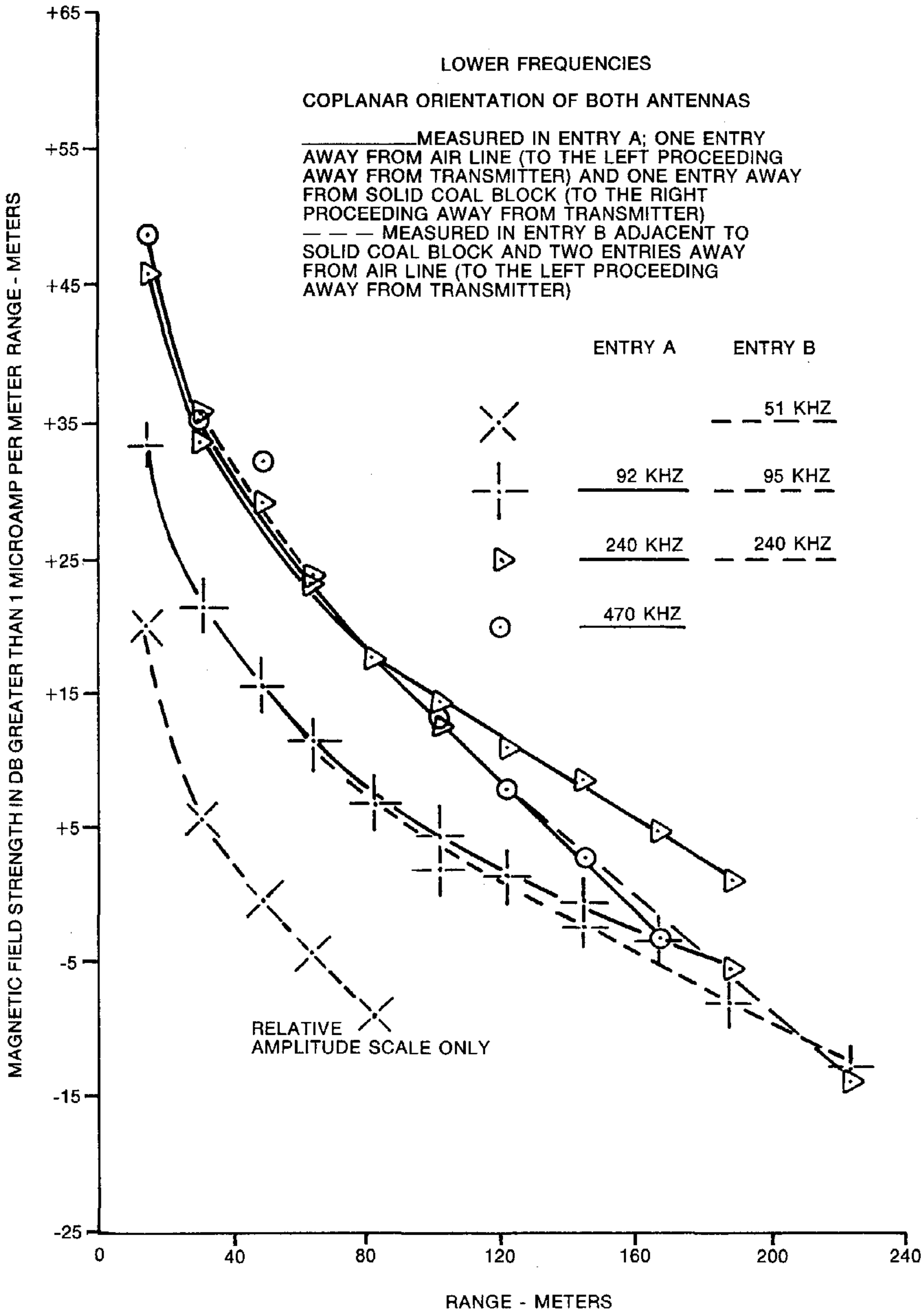


FIGURE 3a

Magnetic Field Strength Vs Range in 3-South Area of Island Creek Coal Co. Virginia Pocahontas No. 1 Mine Comparing Two Paths Along Adjacent Entries (TX Located in Measurement Entry for Both Paths) NIA = 2.5.

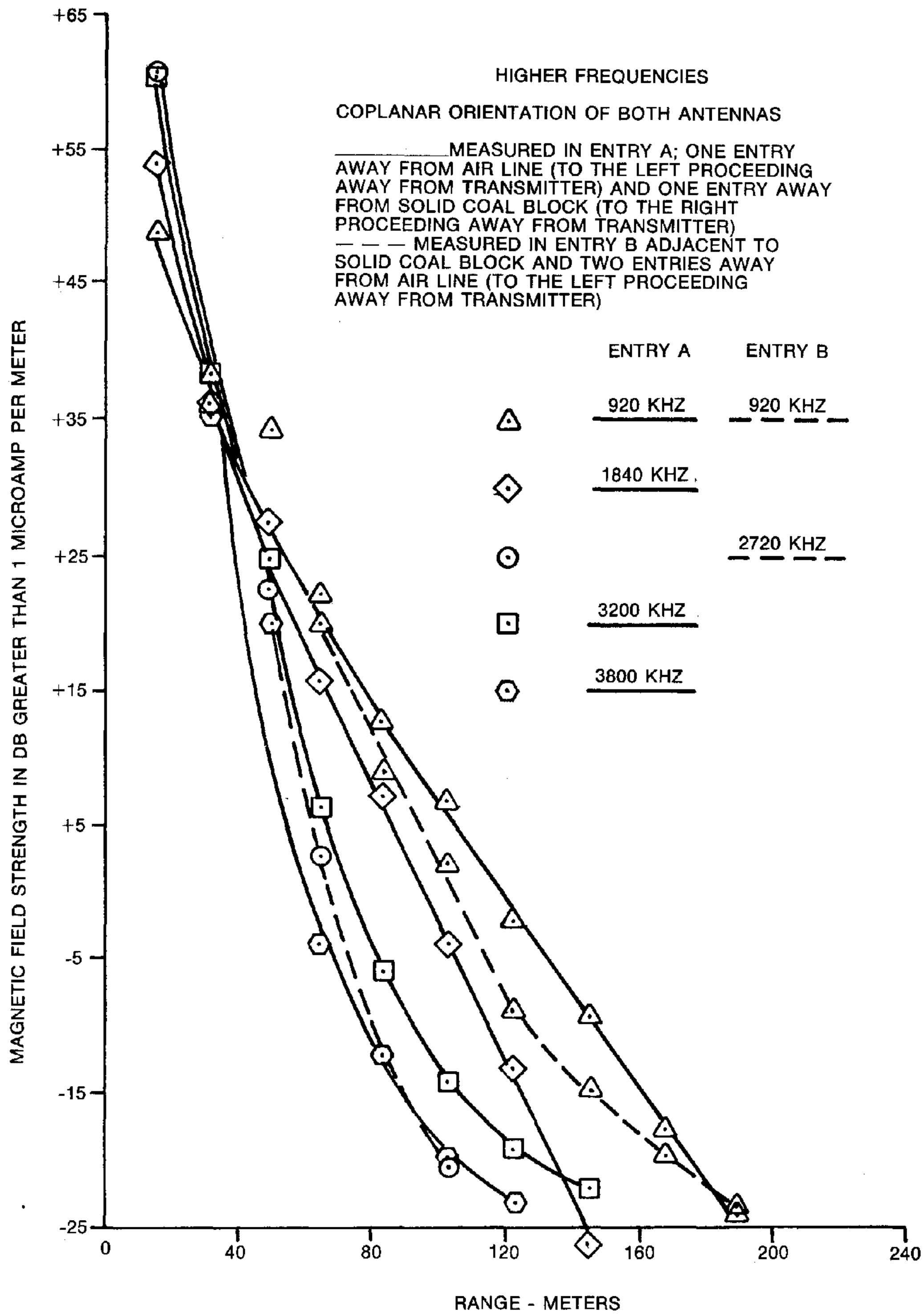


FIGURE 3b

Magnetic Field Strength Vs Range in 3-South Area of Island Creek Coal Co. Virginia Pocahontas No. 1 Mine Comparing Two Paths Along Adjacent Entries (TX Located in Measurement Entry for Both Paths) NIA = 2.5.

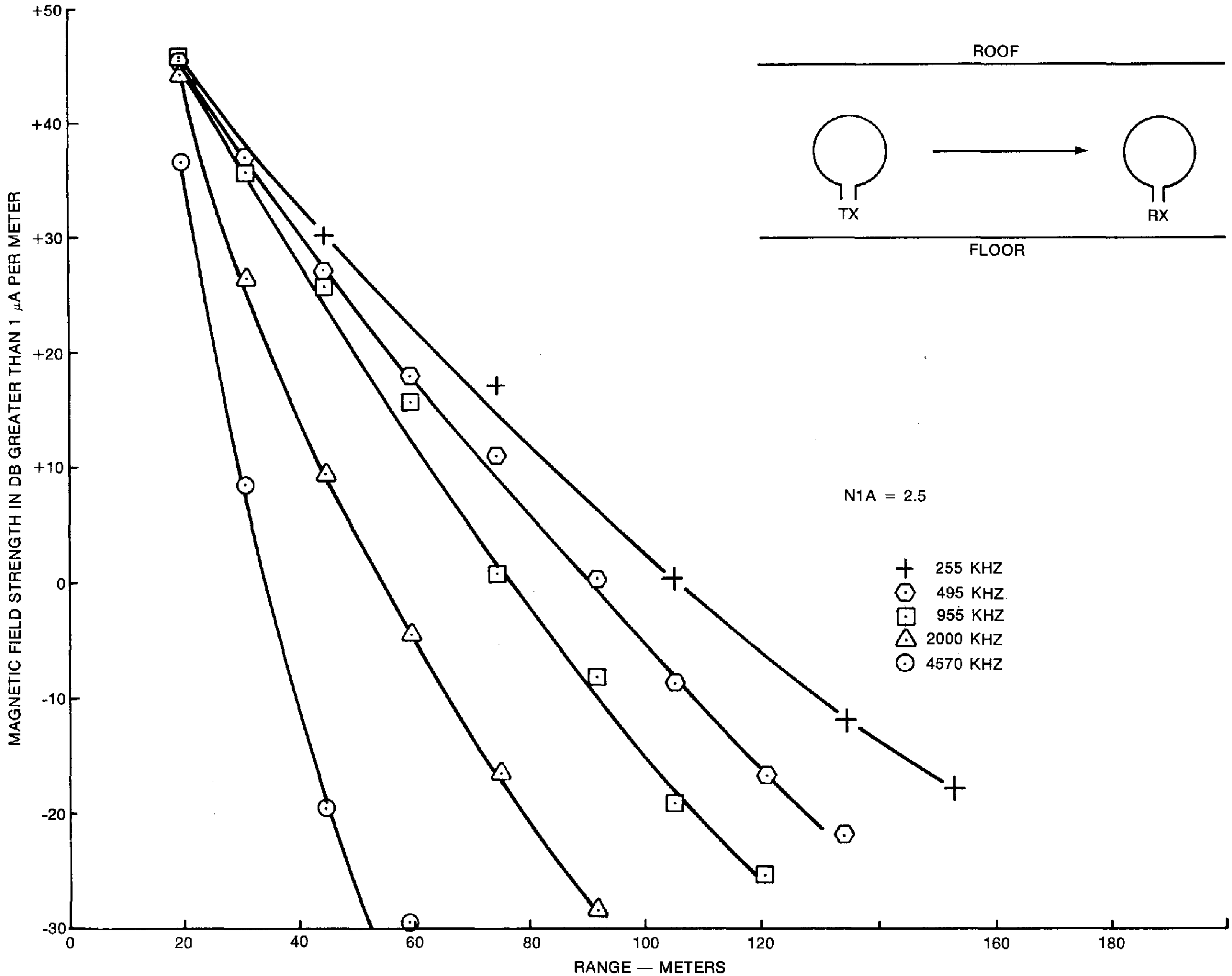


Figure 4 Magnetic Field Strength Vs Range in Quasi-Conductor-Free Area of 1 Main East for Coplanar HMD Orientation of Transmit and Receive Antennas.

Maximum communication range in quasi-conductor-free areas showing frequency dependance for set-noise-limited and median-mine-noise-limited operation in the three seams is shown in Figure 5. These calculations were made using the measured field strength data plus system assumptions given in Table 3. Median mine noise is that defined by Decker(2) from the NBS study(3). When more than one entry away from conductors, the MF field strength measurements were always set-noise-limited. Checks in major haulageways during this program have confirmed median mine noise although amplitude probability distributions of this noise with time are really required for accurate system performance prediction. The maximum communication range curves of Figure 5 represent an average over the available mine samples per seam. The variations experienced over the individual Pittsburgh seam data samples were 470-525 meters in maximum communication range and 300-500 KHz in optimum frequency.

Maximum communication range estimates using a simple scatter gain model to account for propagation in the presence of conductors are illustrated in Figure 6 for the Pittsburgh seam. The calculations supporting these estimates use typical quasi-conductor-free measured field strengths (specifically those given in Figures 2, 3, and 4) to define the field impinging on the conductor(s) plus calculated attenuation of the carrier current along the conductor(s) and calculated scatter gain to define the scattered field strength at the receiver. These maximum communication range estimates show an increase in optimum frequency with increase in the separation of one of the radios from the conductor. The scatter gain and conductor attenuation formulations and assumptions are illustrated in Tables 4 and 5. A comparison of scatter gain estimates derived from the measured data with computed scatter gains is shown in Table 6.

The least comprehensive treatment during the measurement program was given to determining maximum range along continuous (or contiguous) conductor strings when both radios were in the same entry as the

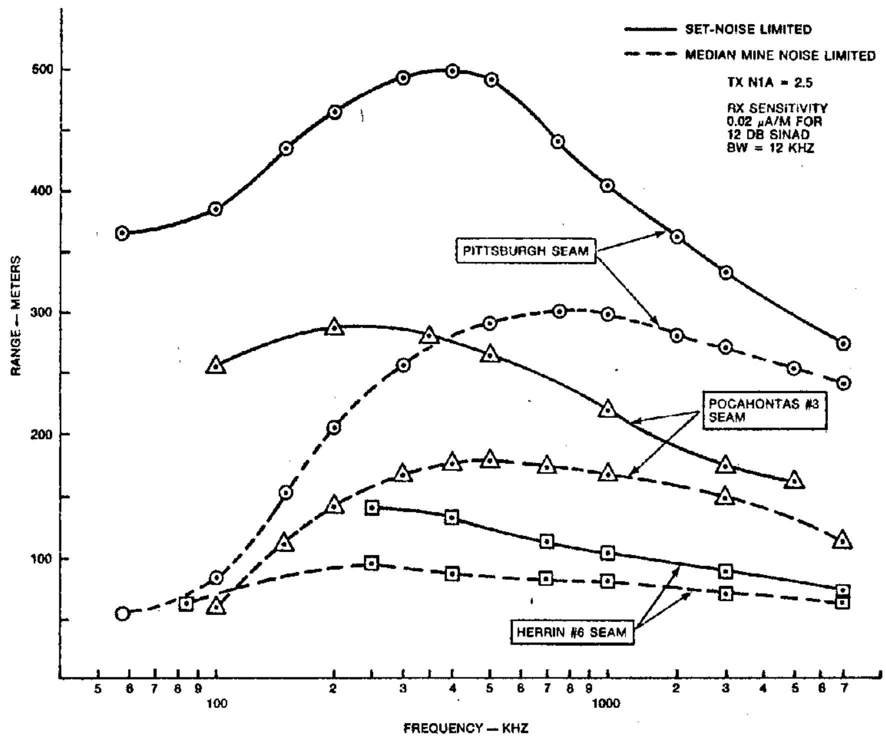


FIGURE 5

Maximum Communication Range Vs Operating Frequency In Conductor-Free Areas of Coal Mines - Geometric Average of All Data Sets.

TABLE 3

SYSTEM CALCULATION ASSUMPTIONS

- PORTABLE RADIOS
- TRANSMIT NIA = 2.5 , (20 WATTS INTO 0.22 M^2 ANTENNA, 7 TURNS, NA = 1.52) MATCHED TO 50 OHMS
- (A) SET-NOISE-LIMITED RECEPTION

  SENSITIVITY -34 DB > 1 μA/M @ 520 KHZ FOR 12 DB SINAD (HAMSHUR)

  VARIES AS 1/F

  (B) MEDIAN-MINE-NOISE-LIMITED RECEPTION

  H_N = 121.5 - 32.5 LOG_{10} F(HZ) IN DB > 1 μA/M/ $\sqrt{HZ}$

  13.5 DB S/N REQUIRED FOR 12 DB SINAD (HAMSHUR)

  NOTE: FOR 12 DB SINAD @ 520 KHZ FOR 2 DB NOISE FIGURE

HAMSHUR CALCULATION GIVES	-34 DB
SHIMBO CALCULATION GIVES	-37.6 DB
MEASURED DATA GIVES	-35.5 DB

  NEW ECAM SSB 0.2 μV FOR 15 DB S/N

  WHEN ADJUSTED FOR 10 DB S/N GIVES -34.9 DB
- RECEIVE ANTENNA LOCATED 2 METER FROM CONDUCTOR (WHEN APPLICABLE)
- TRANSMIT ANTENNA LOCATED 1, 2, OR 3 ENTRIES FROM CONDUCTORS (WHEN APPLICABLE) ENTRY SPACING = 30 METERS
- CONDUCTORS HAVE 50 OHM SURGE IMPEDANCE

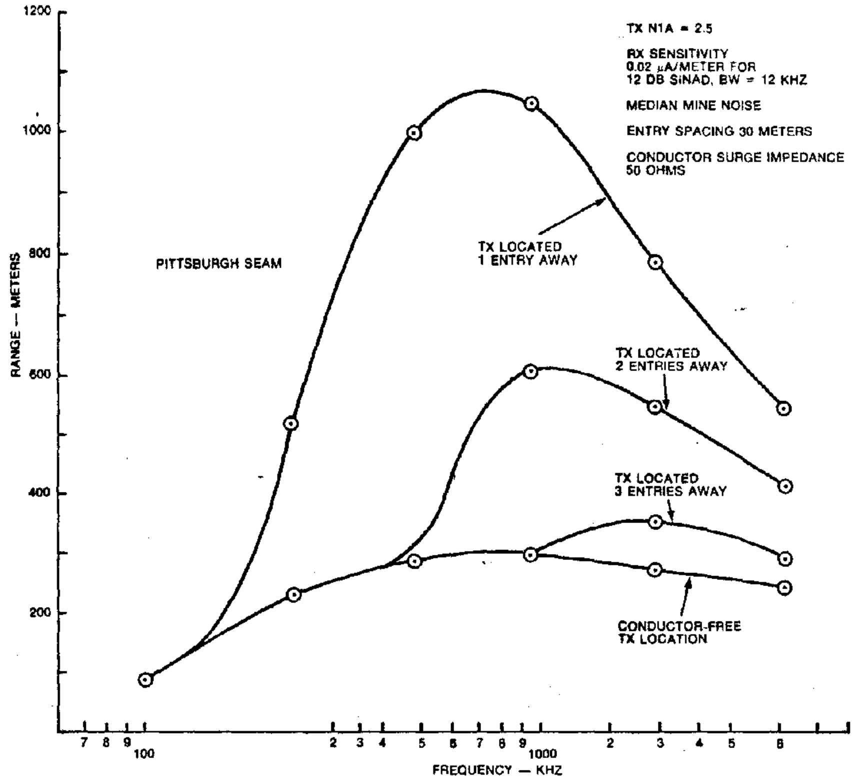


FIGURE 6

Maximum Communication Range Vs Operating Frequency in Proximity to Conductor's RX Antenna 2 Meters From Conductor(s), TX Located Remotely - Pittsburgh Seam.

TABLE 4

SCATTER GAIN FORMULATION

$$G_S = \frac{H_S}{H_M} = \frac{\text{FIELD STRENGTH @ RECEIVER WITH CONDUCTOR PRESENT}}{\text{FIELD STRENGTH @ RECEIVER IN ABSENCE OF CONDUCTOR}}$$

$$H_S = \frac{30\,(\alpha^2+\beta^2)^{\frac{1}{2}}\,(NIA)}{4\pi\, Z_{os}\, R_{TC}\, R_R}$$ FROM A SINGLE INFINITE CONDUCTOR IN A HOMOGENIOUS MEDIUM

$$H_M = \frac{(\alpha^2+\beta^2)^{3/4}\,(NIA)}{\sqrt{8\pi R_{TR}}\,(h+\delta_r)}$$ FROM ADL ANALYSIS MODEL

ASSUME $R_{TR} \approx R_{TC} = R$

$$G_S = \frac{11.97}{Z_{os}\,(h+\delta_r)\,\sqrt{R_T}\,R_R\,(\alpha^2+\beta^2)^{\frac{1}{4}}}$$

FROM EVALUATING ADL DATA FOR &

- PITTSBURGH SEAM $\quad G_S \approx \frac{9.53}{Z_{os}\sqrt{R_T}\,R_R}$
- HERRIN #6 SEAM $\quad G_S \approx \frac{7.0\, f_{MHZ}^{\frac{1}{2}}}{Z_{os}\sqrt{R_T}\,R_R}$
- POCAHONTAS #3 SEAM $\quad G_S \approx \frac{3.76\, f_{MHZ}^{.075}}{Z_{os}\sqrt{R_T}\,R_R}$

TABLE 5

CONDUCTOR ATTENUATION

- USING $$\frac{\alpha}{k} = \frac{\pi/8}{\left[\ln\frac{2a}{a_i}\left(\ln\frac{2a}{a_i} - \ln\frac{2.517}{\delta_r}\right)\right]^{\frac{1}{2}}}$$
  a_i is the conductor radius
  a is the conductor-to-conducting-surface spacing
- FOR THE PITTSBURGH SEAM, ASSUMING $\sigma_R = 0.085$ MHO/METER
  $\frac{\alpha}{k} = 0.14652$ @ 1 MHZ
- FOR THE HERRIN #6 SEAM, ASSUMING $\sigma_r = 0.22$ MHO/METER
  $\frac{\alpha}{k} = 0.16030$ @ 1 MHZ
- FOR THE POCAHONTAS #3 SEAM, ASSUMING $\sigma_r = 0.01$ MHO/METER
  $\frac{\alpha}{k} = 0.12518$ @ 1 MHZ

TABLE 6

COMPARISONS OF MEASURED & COMPUTED SCATTER GAINS

SEAM	FREQUENCY KHZ	CONDUCTOR TYPE	OHMS Z_{os}	METERS R_T	METERS R_R	MEAS SCATTER GAIN DB	CALC SCATTER GAIN DB
PITTSBURGH	935	WATER LINE	50	46	1	-28	-31
PITTSBURGH	236	WATER LINE BURIED	10	55	½	-13.5	-11.8
POCAHONTAS #3	520	WATER LINE	50	62.6	1	-35	-35.5

conductors. Although typical maximum ranges were never measured, communications was established using the ECAM radios over ranges of up to two miles; limited only by the inability to move farther away and not by signal attenuation.

The EM Signalling contract (H0366028) provided measurements supporting area coverage of working section panels of large extent from a transmitter located adjacent to the support cabling/power and control wiring of a section belt conveyor head or tail piece. Section panels of extent up to 1100 meters long and 120 meters wide were mapped including fields produced near the face (up to 200 meters forward of the nearest conductor) from a transmitter located at the head piece. The area coverage was displayed in the reduced data in terms of contours of magnetic field intensity plotted from a large number of discrete measurements made in the panel. A typical set of contours for a panel in the Ireland mine together with the mine topology are given in Figures 7 and 8. These mappings were made over a range of frequencies with a frequency of about 1 Mhz being optimum. Data gathered at 230 KHz in the Inland Steel Mine #1, although not shown, has confirmed that area coverage can be provided over the working areas of adjacent working sections panels spaced 120 meters apart and also over panels on opposite sides of a set of mains from a single transmit location near the center of one of the panels.

IV CONCLUSIONS

The measurements and data evaluations reported in this paper are supportive of the practical implementation of two-way radio communications in coal mines; both in sections and in major haulageways. The section communication embraces both transmission in areas of substantial size without conductors (communications over paths of 200-400 meters in the Pittsburgh seam and up to 150 meters in the

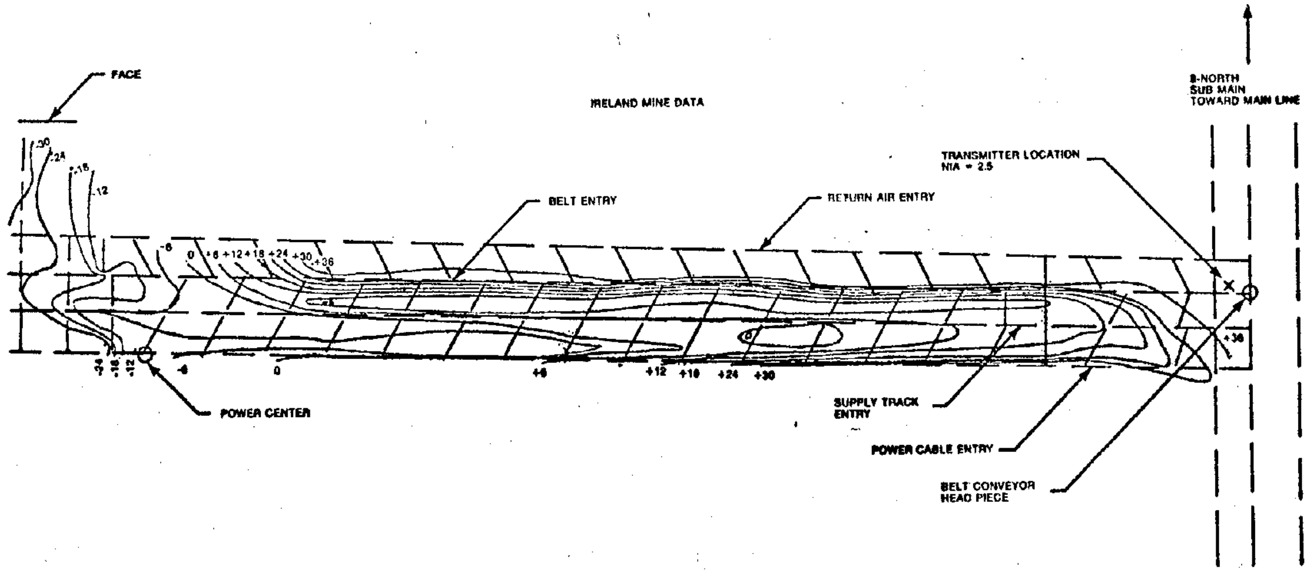


Figure 7 Magnetic Field Strength Coverage Map of 8-North 3-Right Working Section at 486 kHz With Transmitter Located Adjacent to the Conveyor Headpiece and With the **Transmit Antenna Loop Plane Oriented Parallel to the Entries — The Field at Any Point is the Maximum Field in dB Greater Than 1 µA Per Meter.**

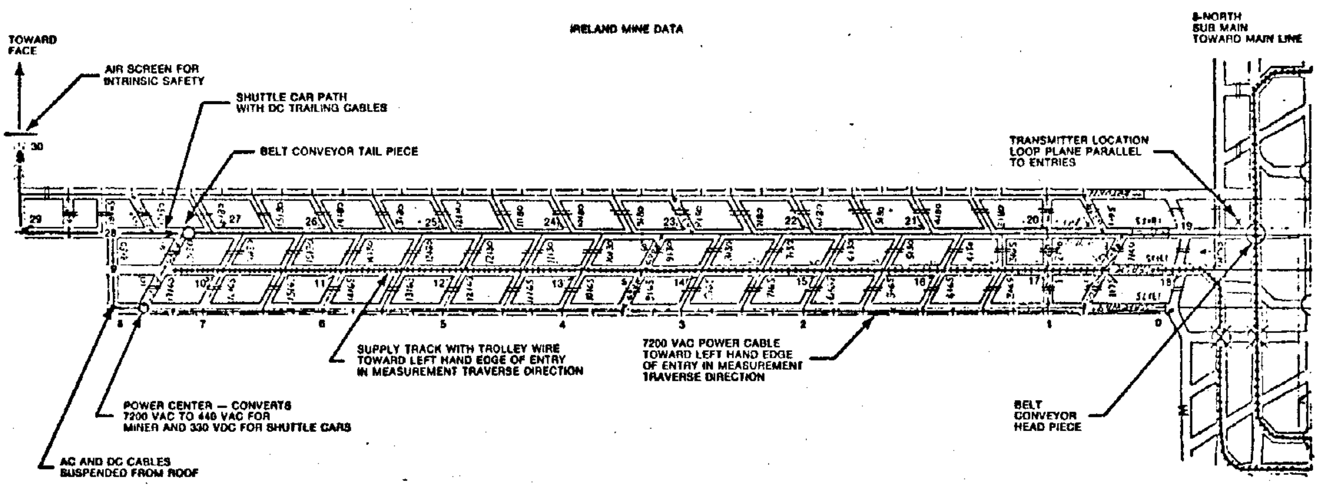


Figure 8 Topology of 8-North 3-Right Working Section Showing Measurement Traverse For Magnetic Field Strength Mapping.

Herrin #6 seam) and over the conductor carrying and conductor-free entries of a working section in area coverage. Haulageway communication has not been specifically investigated for maximum range, but attenuation past load sleds, AC power junction boxes, and transformers has been found to be less than 6 dB. Promise for MF wireless haulage communication is held at least where the mine wiring is integral.

The optimum frequency for achieving maximum communication range is expected to vary from mine to mine and from seam to seam and according to the anticipated use in the mine. The variations between mines within a given seam are sufficiently small that a near optimum single frequency would suffice for a given intended application.

The current propagation data base is sufficient to permit predictive system designs for many mines for applications in areas which are either conductor-free or else which are conductor dilute. Situations have yet to be found where the presence of conductors materially degrades the conductor-free performance. System designs specifically depending on coupled carrier current in conductors are not yet predictable with confidence; however, preliminary performance estimates can be made with the results identifying particular areas in which measurements should be made in order to define system requirements. Continued work is needed in defining improved modeling for general two-way comminications in coal mines via radio. This work includes the definition of the coupling between radio antennas in the same entry as mine wiring.

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(3) W. Bensema, M. Kandu, I.W.Adams, "Electromagnetic Noise in Robena #4 Coal Mine, NBS Technical Note 654, April 1974; and other technical notes in the same series, U.S.B.M. Contract

HARRISON R. COOPER SYSTEMS, INC.
AMF BOX 22014
SALT LAKE CITY, UTAH 84122
PHONE (801) 295-2345

MEDIUM FREQUENCY COMMUNICATION EXPERIMENTS

The purpose of this discussion is to make known to the interested parties information gained as a result of testing HRCS medium frequency communication system under operating conditions.

It is not the intent of this presentation to cover in detail the theory or design of the system, but rather to give to those interested a brief summary of the actual test and findings gained through the installation and use of the medium frequency system under typical conditions encountered in a working mine in operation.

This has proven to be a slow and difficult way to gain test data but has also been valuable to us in determining the necessary equipment required in order to make an m.f. system operate successfully under the widely ranging conditions found in the mining industry. A number of antenna configurations have been evaluated to determine a design of a practical system. The optimum degree of noise filtering and power level is being investigated, in order to determine the minimum transmission power suitable for effective signal propagation.Most of our testing has been confined to relative deep hard rock western mines. Some testing has been conducted in the lead mines of the central United States.

Presented by L.D. Whittington on behalf of H.R. Cooper

PASSIVE REFLECTORS AS A MEANS FOR EXTENDING UHF SIGNALS DOWN INTERSECTING CROSS CUTS IN MINES

Robert L. Chufo
Bureau of Mines
U. S. Department of the Interior
Pittsburgh, Pennsylvania

R. A. Isberg, P.E.
Comsul, Ltd.
San Francisco, California

Abstract

Tests of communication between hand held two watt UHF transceivers in a room and pillar limestone mine were satisfactory for several thousand feet through straight haulage ways but the range of communication at right angles to the haulage way into intersecting cross cuts was quite limited. It was evident that the radiation from the transceivers was not being reflected by the limestone pillars into the intersecting cross cuts. Two passive aluminum reflectors, each four feet square, were installed near the ceiling and positioned 45° with respect to the axis at an intersection of a main haulage way and a cross cut. The range of communications down the intersecting cross cut was significantly extended. Propagation measurements on 466 and 812 mHz in the same mine demonstrated that passive reflectors are a practical and inexpensive means for extending communications into intersecting cross cuts. The same formulae and nomograms which are used for microwave passive reflectors can be used for UHF reflector design.

Introduction

The Bureau of Mines has established a project to demonstrate means for improving underground communications in metal/non metal room and pillar mines. The Black River Mine, which is a large limestone mine near Butler, Kentucky, was selected for conducting propagation tests on 466 and 812 mHz.

The Black River Mine is approximately one-half mile in diameter and six hundred fifty feet under ground. The excavated area of the mine is a symmetrical pattern of cross cuts approximately forty feet wide and thirty-five feet high which are separated by limestone pillars approximately thirty-five feet square and seventy-five feet from center to center. Since most of the cross cuts are clear of obstructions, they can be used by mine vehicles so the capability for radio communications throughout the mine is desired.

Tests by mine personnel using 150 mHz transceivers indicated that the VHF communication range was limited to several hundred feet.

A series of preliminary tests using two watt UHF transceivers demonstrated that communication was satisfactory for at least seventeen hundred feet down a cross cut that is used as a main haulage way. However, the range of communication down right angle intersecting cross cuts was unreliable beyond a few hundred feet and was subject to multipath interference. This appeared to be due to the high degree of attenuation of the 466 mHz signals through the maze of limestone pillars and the lack of efficient reflecting surfaces on their sides.

Since the mine has thirty-seven east-west and forty-eight north-south cross cuts, the use of leaky transmission lines and in line repeaters was not considered to be an economical and practical solution.

A network of branching low loss transmission lines and numerous antennas fed by a 60 watt base station through power dividers was considered and it appears to be a possible solution. However, a solution avoiding extensive cable installation was desired so passive reflectors in conjunction with several low power repeaters and control stations were considered.

Passive reflectors are used extensively for extending microwave communication above 2000 mHz around obstacles and the theory of their design and operation is well documented and understood.[1,2]

A review of passive reflector design theory and nomograms indicated that passive reflectors would afford adequate gain at 466 mHz to compensate for the attenuation of signals around pillars into intersecting cross cuts. Figure 1 is a typical nomogram for computing reflector gain as a function of frequency.

Computation of Passive Reflector Gain

The gain of passive repeater flat billboard reflectors is computed as follows:

$$G = 22.2 + 40 \log_{10} f + 20 \log_{10} A + 20 \log_{10} \cos \alpha$$

G = 2 way gain of reflector in dB

F = Frequency in GHz

A = Area of passive reflector in square feet

α = 1/2 of the included angle between the incident and reflected paths.

Since the roof to floor heights in the Black River Mine vary between twenty-four and thirty-four feet, a four by four foot reflector was considered to be a practical size to permit adequate clearance to vehicles moving below it.

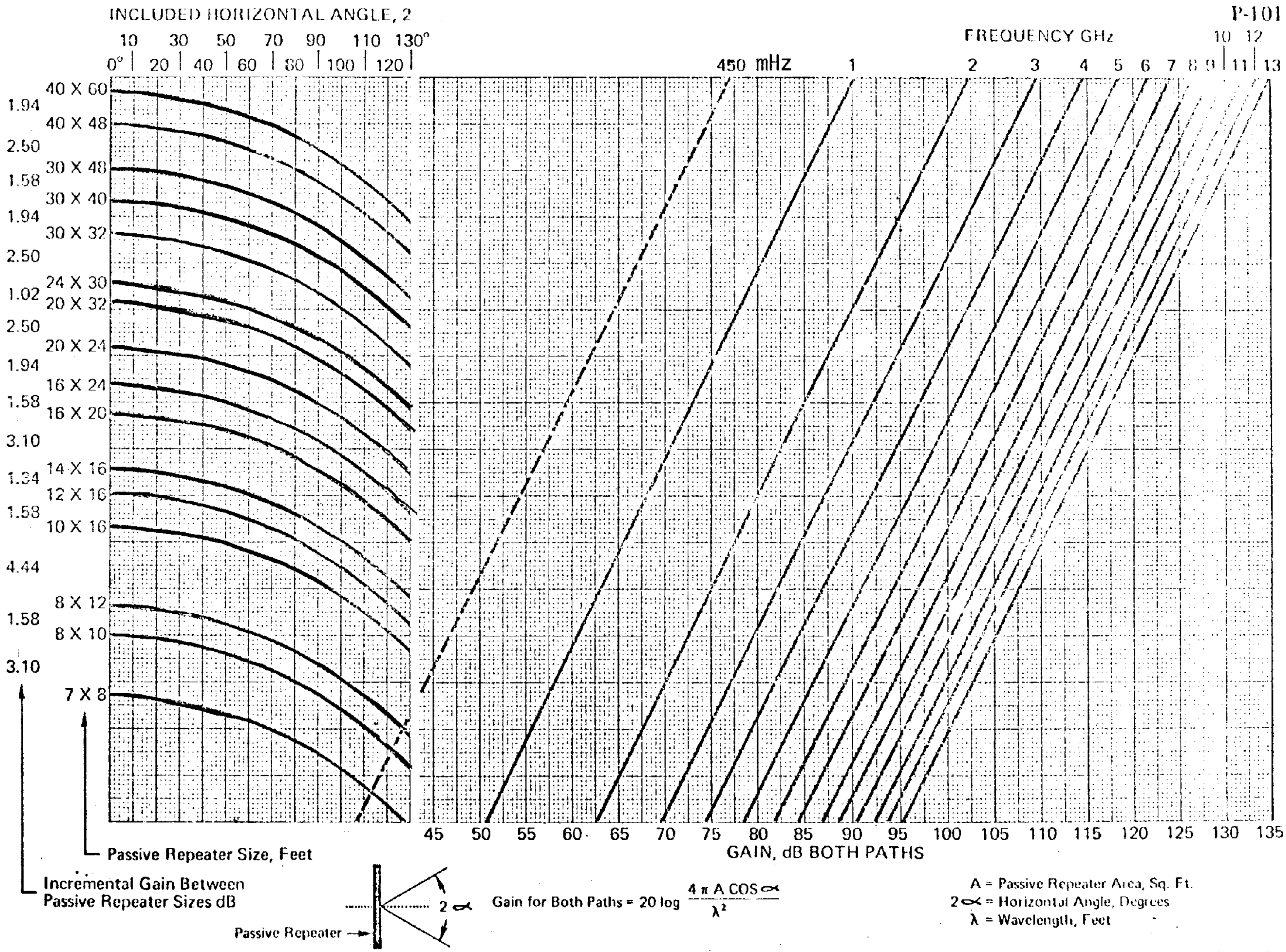

Figure 1 Passive Repeater Gain Chart

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Following is a computation of the gain of a 4' x 4' reflector:

$G_R = 22.2 + 40 \log 0.45 + 20 \log 16 + 20 \log \cos 45°$

$G_R = 22.2 + (-13.87) + 24.08 + (-3.01) = 29.4$ dB

G_R is increased 6.02 dB when A is doubled.

Computation of Attenuation Between Antennas

The free space attenuation between isotropic antennas in dB is equal to:

$A_i = 20 \log D \text{ miles} + 20 \log f \text{ mHz} + 36.6$

Since the distances in the mine are less than a mile, there are refraction losses in the rock and dipole antennas are used, the following formula for the attenuation between the transmitter output and receiver input was derived:

$A_T = 20 \log D \text{ feet} + 20 \log f \text{ mHz} - 37.85 + R_L + T_L - (G_{Tx} + G_{Rx})$

R_L = Refraction losses in cross cuts in dB per one hundred feet x D in hundreds of feet.

T_L = Transmission line losses to T_x and R_x antennas.

G_{Tx}, G_{Rx} = Antenna Gain respect to $\frac{\lambda}{2}$ dipole + 2.15 dB, which is the gain of a $\frac{\lambda}{2}$ dipole with respect to an isotropic antenna.

The attenuations of 466 and 812 mHz signals were measured in the 2200 foot slope of the Black River Mine and the results are plotted on Figure 2 along with the two free space attenuation curves. At 1000' the 466 mHz signal strength was 27 dB less than the free space value and R_L = 2.7 dB/100'. The 812 mHz signal strength was 14 dB less than the free space value and R_L = 1.4 dB/100'.

The cross section area of the cross cuts in the mine (35' x 40') is approximately seven times the 9.5' x 20' cross section of the slope and signal strength measurements indicated that the loss factor R_L in the cross cuts at 466 mHz did not exceed one dB per 100'.

The free space attenuation between two dipole antennas and a reflector is computed from the following formula:

$A_{Total} = A_1 + A_2 - G_R$

A_{Total} = Total attenuation between the two dipole antennas with passive reflector interposed.

A_1 = Path attenuation between antenna No. 1 and the passive reflector.

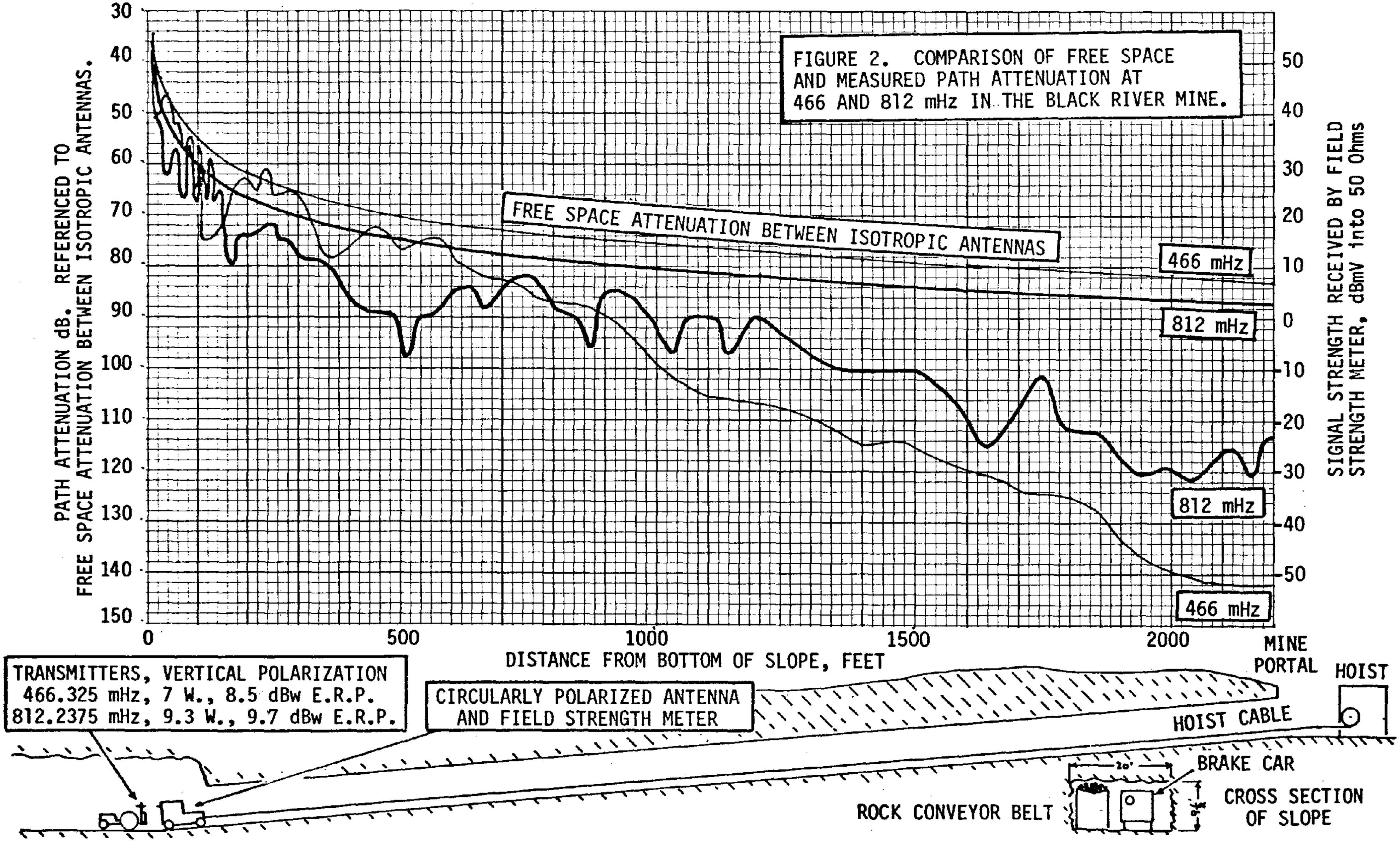

FIGURE 2. COMPARISON OF FREE SPACE AND MEASURED PATH ATTENUATION AT 466 AND 812 mHz IN THE BLACK RIVER MINE.

A_2 = Path attenuation between the passive repeater and antenna No. 2.

G_R = Gain of 4' x 4' passive reflector = 29.4 dB.

Referring to Figure 4, assume that a two watt 466 mHz transceiver No. 1 is operating in the center of the intersection of cross cuts 205S and 119 and a passive reflector is located 800' down 205S in the center of the intersection of cross cuts 205S and 130 with the reflected paths aligned along the axis of 130E and 130W. Also assume that a second transceiver is operating in 130E 400' from the reflector.

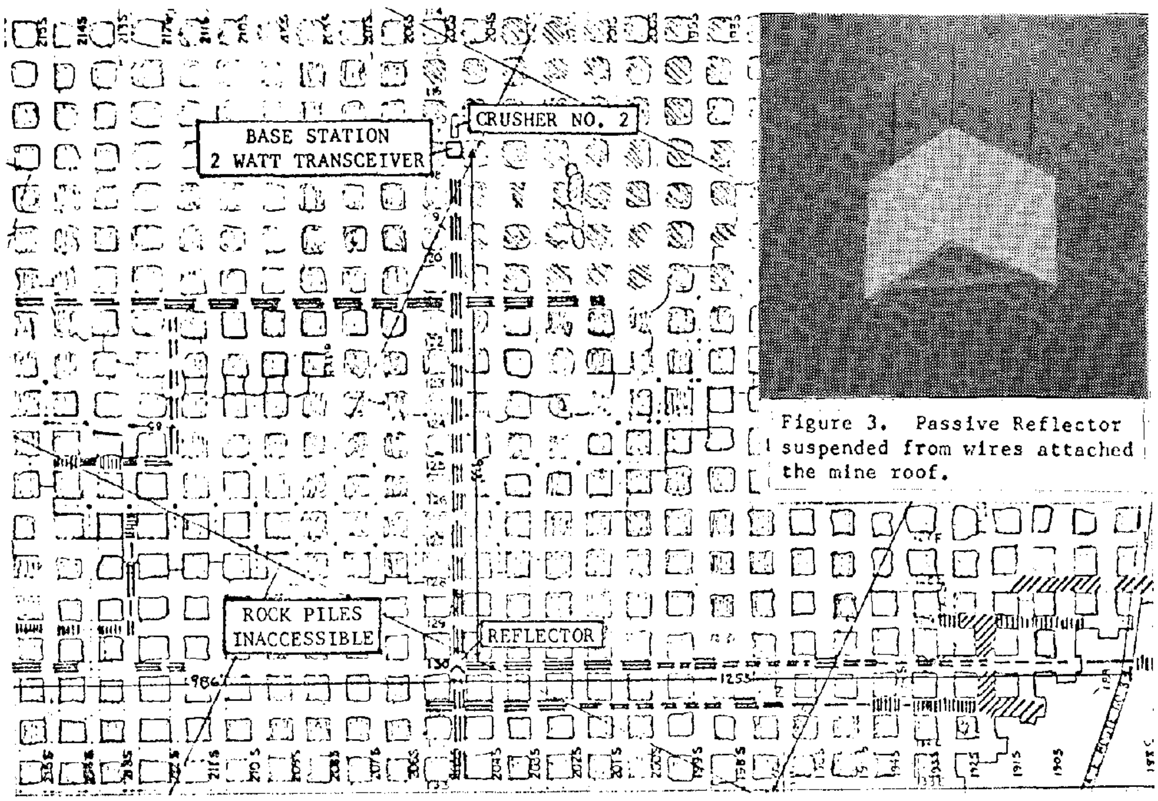

Figure 3. Passive Reflector suspended from wires attached the mine roof.

Figure 4. Portion of Black River Mine Map showing the location of 4' x 4' aluminum reflector hung 28' above the floor at the center of the intersection of cross cuts 205S and 130. The qualitative signal evaluations using two watt transceivers are shown on the map by dashed lines as follows:

Excellent signal, full quieting, no multipath.

Good signal, adequate quieting.

Satisfactory signal, some impairment.

Marginal, critical squelch or multipath objectionable.

Unsatisfactory, occasionally breaking squelch.

The estimated received signal power will be:

Free space attenuation of 800' path	A_1 = - 74 dB
Free space attenuation of 400' path	A_2 = - 68 dB
Gain of 4' x 4' reflector	G_R = 29 dB
Free space path attenuation	A_T = -113 dB
Refraction losses 1200' Estimated as 0.8 dB/100' at 450 mHz	R_L = - 10 dB
Total path attenuation for the 1200' Reflected path in mine	A_M = -123 dB
Transceiver output power	T_xP = 30 dBm
Derating factor for transceiver transmitting antenna re isotropic	G_{Tx} = - 10 dB
Derating factor for transceiver receiving antenna re isotropic	G_{Rx} = - 10 dB
Estimated received power	R_xP = -113 dBm

The estimated received power is -113 dBm which is adequate for 20 dB quieting of the transceiver.

Qualitative Tests to Verify Reflector Performance

In order to qualitatively verify the preceding compuations, two 4' x 4' flat aluminum passive reflectors were constructed and tested in the Black River Mine in October, 1977. The two reflectors were mounted on a wood frame, the included angle of the reflector was aligned along the axis of cross cut 205S facing the base station site at crusher No. 2. See Figures 3 and 4. The reflecting surfaces were at 45° angles with respect to the axis of cross cut 130 (E and W). A two watt transceiver operating on 466 mHz served as a base station near crusher No. 2 and two two watt transceivers were carried through the east-west cross cuts to qualitatively test the range and quality of communication in cross cut 130 (E and W) compared to the range and quality of communication in parallel cross cuts. The results of the test are shown on Figure 4.

Audio tape recordings were made of the voice communication heard from the loudspeakers at both the base and roving transceivers. These recordings were carefully evaluated and the transmissions were rated according to the table on Figure 4.

The test results indicated that the passive reflector definitely enhanced the two way signal strengths in cross cuts 130 E and W and also increased the useful range of communication in adjacent parallel cross cuts 131E, 129E and 128E.

Multipath distortion and the need for critical positioning of the transceiver for optimum reception was substantially reduced in cross cut 130. However, as the distance from the reflector in cross cut 130 was increased to approximately 1,000 feet the signal strength was not great enough to penetrate beyond the width of one pillar (40') into intersecting cross cuts on the north-south axis.

Signal Strength Measurements

A series of signal strength measurements were made in the Black River Mine on 466 and 812 mHz to obtain signal propagation data in a room and pillar metal/non-metal mine. These measurements included studies of vertical, horizontal and cross polarization in addition to measurements of reflector performance.

Control Center

The signal strength measuring equipment was located in a large trailer van near crusher No. 3. This location was near 115 volt A.C. power and it also was near the outer perimeter of the mine affording a large area of long intersecting cross cuts for the signal strength measurements. See Figure 8.

Signal strengths of 466 and 812 mHz mobile radio transmissions received by the base station receivers were simultaneously recorded on a two channel strip chart recorder. Modified signal quality modules from a receiver voting system were used for measuring signal strengths above carrier quieting. Since the modules measure audible noise the carriers were not modulated during the measurements. The receivers and voting modules were calibrated with signal generators and all readings were in microvolts input to the receiver antenna receptacle. An array of antennas was attached to a roof bolt at the center of intersecting cross cuts 190S and 111E. The antennas included 466 and 812 mHz omnidirectional 1/4 wave dipoles on ground planes and horizontal and vertically polarized Yagis which were directed down the two cross cuts.

Mobile Unit

The mobile radio equipment was mounted on a plywood assembly which was strapped to the lift platform on a utility tractor shown on Figure 5. This assembly included a lighted map shelf unit, and an audio tape recorder for recording voice transmissions received and transmitted as well as the field engineers and drivers comments. Both mobile radios operated with 13 to 17 watts output power into 1/4 wave ground plane antennas which could be positioned for either vertical or horizontal polarization. See Figure 6.

The special audio control unit was constructed by the Bureau of Mines technicians for controlling the modulation of both transmitters and for interfacing the audio tape recorder and two headset microphones which were worn by the tractor driver and mobile radio operator. The audio tape recordings made on the mobile unit were very helpful for identifying the regions where the mobile unit was not heard at the control center. At each intersection, the tractor driver would announce the mobile unit's location and the control center would respond. The signal strength measurements were made while the mobile unit was in motion and the median values were tabulated from the chart recordings.

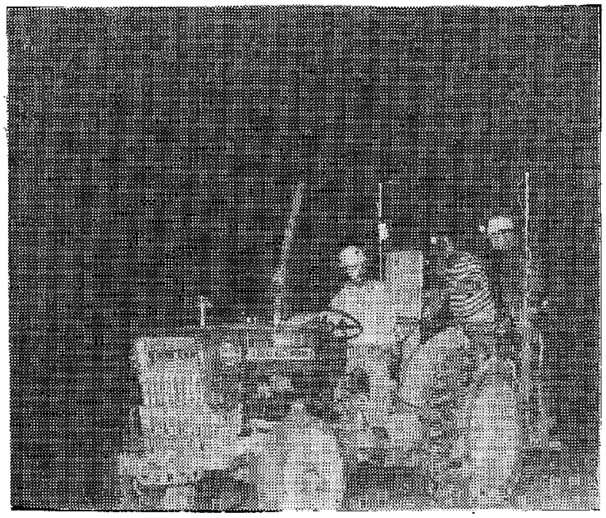

Figure 5. The mobile unit "Ben Hur" transmitted simultaneously on 466.325 and 812.2375 mHz while the signal strengths received at the control center were recorded. One-quarter wave whip antennas were mounted on ground planes on top of the poles. The field crew were Steve Wilson, driver/navigator, Don Parrish, operator and Ned Mountain, engineer.

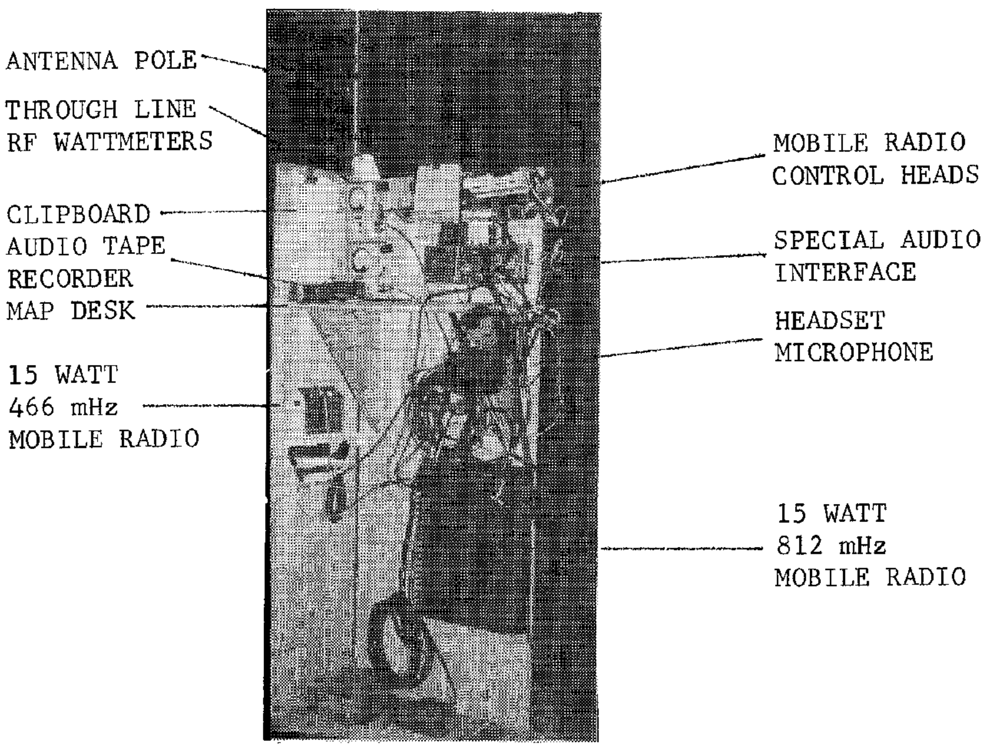

Figure 6. Mobile Radio Assembly that was mounted on lift platform on tractor for mobile signal strength measurements.

Signal Strength Measurements of Reflector Performance

Three additional 4' x 4' reflectors were constructed and were hung in major haulage way intersections. One reflector was installed 1425' from the control center's antenna at the intersection of 190S and 130E. This made it possible to test a double reflection because the first reflector had been installed at 130E and 205S.

One reflector was installed at the intersection of 111W and 208N, 1340' from the control center and the other reflector was installed at 111W and 212N, 1640' from the control center.

The signal strength tests confirmed that the reflectors definitely increased signal strengths and reduced multipath distortion in the intersecting cross cuts where they were installed. Typical signal strength measurements along cross cut 130 (E and W) and adjacent parallel cross cuts are tabulated on Figure 7.

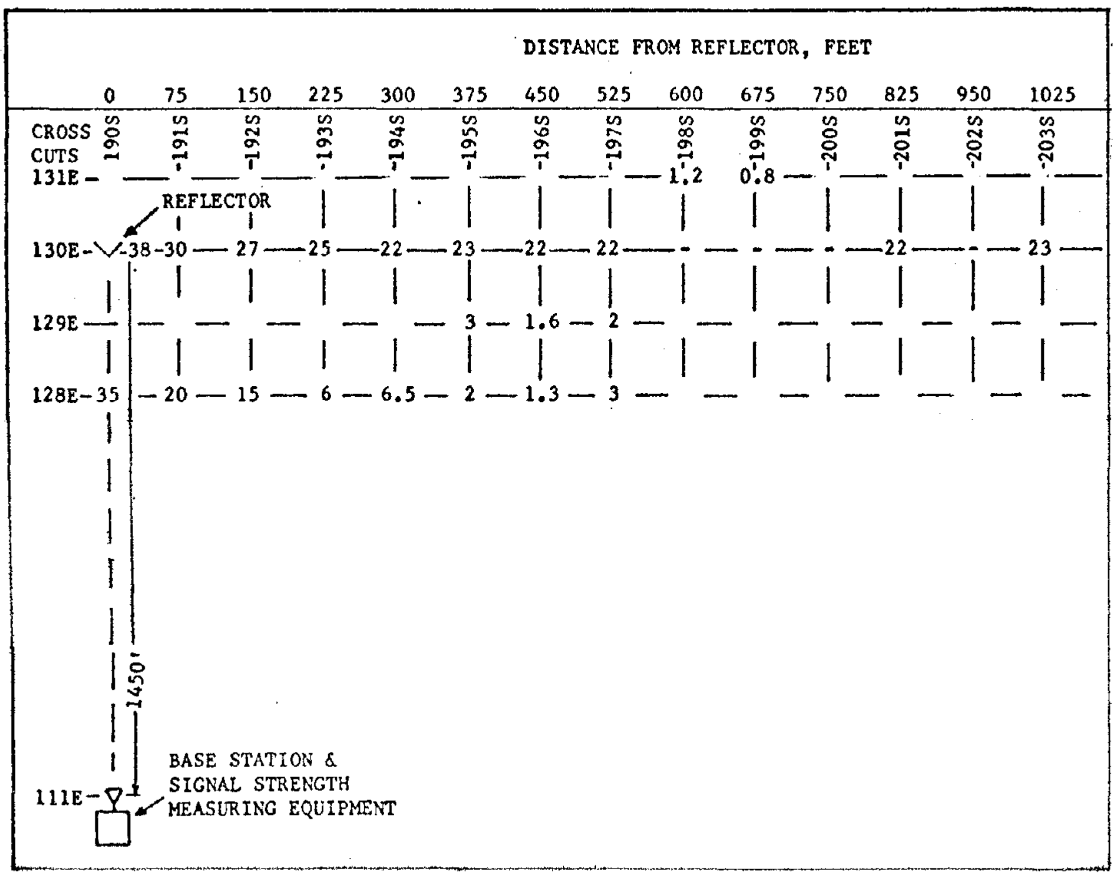

Figure 7. Median signal strengths in microvolts measured at base station receiver input receptacle from 466 mHz mobile transmitter along the reflected path.

JRE 8. PROPOSED 450 mHz RADIO SYSTEM FOR THE BLACK RIVER MINE USING TWO CONTROL STATIONS AND TWO BASE/MOBILE AY STATIONS IN THE HOIST HOUSE AND IN THE UNDERGROUND SHOP OFFICE AND ONE MOBILE RELAY STATION AT CRUSHER NO. 3. DISTRIBUTED ANTENNAS WILL SERVE MAIN HAULAGE WAYS WITH DIRECT RADIO PATHS, PASSIVE REFLECTORS WILL SERVE BRANCHING HAULAGE WAYS.

Proposed UHF Radio System for the Black River Mine

The proposed radio system would use two UHF two frequency simplex channels in the 450 mHz Special Industrial Radio Service. Two base/mobile relay stations and two control stations would be located in the hoist house at the slope's portal, two base/mobile relay stations and two control stations would be located in the underground shop office near the end of the slope and one mobile relay station would be located near crusher No. 3. Figure 8 is a map of the mine showing the proposed radio station and antenna locations and the direct and reflected radio propagation paths through the mine. The base/mobile relay and control stations in the hoist house will feed Yagi antennas located inside the portal. These antennas will provide two way communication with two watt transceivers in the slope and control of the base/mobile relay station in the shop office. Conversely, the base/mobile relay stations and control stations in the shop office will communicate with transceivers in the slope and can control the base/mobile relay station in the hoist house. Alarm conditions in the mine, such as low velocity air for ventilation, smoke, power outages and an emergency stop signal from the brake car to the hoist operator will be annunciated in the hoist house. The hoist operator will be able to operate sirens in the mine, operate paging loudspeakers in the mine and selectively call transceivers, control stations and mobile radios.

The base/mobile relay stations in the shop office will feed four antennas through two power dividers and low loss transmission lines. The antennas will be inverted omnidirectional 7.5 dB gain antennas attached to roof plates at the center of intersecting main haulageways which are labeled "direct radio path" on Figure 8. Ten 4' x 4' and three 4' x 8' reflectors will be installed where other haulageways intersect the direct radio paths. These reflectors are expected to increase the signal levels approximately 30 dB along the intersecting haulageways affording essentially total coverage of the active areas in the mine.

Conclusions

1. The tests of passive reflectors in the Black River Mining Company's limestone mine demonstrated that the range of 466 and 812 mHz communication can be readily extended into intersecting cross cuts which are approximately forty feet wide and thirty-five feet high.

2. The reflectors are effective with either horizontal or vertical polarization. Vertical polarization is most appropriate for the mobile units but the signals from the mobile unit were received best by cross polarized Yagi antennas.

3. By using a combination of distributed antennas and passive reflectors, it should be possible to establish a UHF radio system in the mine which will provide reliable communications in all of the major haulageways and work areas and in substantially all of the inactive excavated area. All equipment would be battery operated with float

chargers in order to operate during power outages. By using radio control stations, wire control lines will not be required.

Acknowledgments

We gratefully acknowledge the excellent cooperation and facilities furnished by the Black River Mining Company and for the assistance provided by Mr. A. J. Cigallio, Mr. Robert Keuhneman, Mr. Richard Wehrmeyer and their staff.

We also thank Mr. Joe Rupe and Mr. Loren Burkett of the Parma, Ohio, office of Motorola Communications and Electronics, Inc., for the use of the mobile radios and base station equipment.

Richard A. Watson and Carl W. Ganoe of the Bureau of Mines provided technical services and constructed the special assemblies used for the mobile and signal strength measuring equipment.

Special thanks are due the mobile crew; Don Parrish of Comsul, Ltd., operated the radio gear while Steve Wilson of the Black River Mine piloted their lurching chariot "Ben Hur" through the dark and foreboding cross cuts of the mine.

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COMMENT BY P. DELOGNE

A problem with leaky feeders is that they need to be strung at a certain distance of the wall, which increases the laying cost. On the other hand, one can use a non-leaky feeder with antennas connected to it; the cable laying cost is then reduced but the cost of the antennas is not negligible. As a compromise, we designed a coaxial cable with leaky sections acting as antennas; at 450 MHz the leaky section length is of the order of 1 meter and the spacing may be 100 meters. This system may be useful for this application.

REPLY BY R.L. CHUFO

Cables in the U.S. are installed without concern for spacing from the wall. Granted, since mine walls are not smooth, intimate wall contact is not possible. Tunnels in the U.S. are also not lined with metal sets and lock plates. When the cable is mounted to a messenger wire, the manufacturers mounting hardware, which spaces the cable 3/8" from the messenger wire, is generally used. We have not experienced significant excess cable attenuation or reduced radiation due to cable mounting methods.

The distributed antenna system I reported provided a high power density in the cross drifts which were over one thousand feet in length. I do not feel that the coaxial cable with leaky sections would provide the required lateral range, although this special cable would be useful for providing communications in haulageways where lateral range is of less importance.

WIRELESS COMMUNICATION IN THE TUNNELS AND UNDERGROUND STREETS

J. Chiba[*], T. Inaba[**], Y. Kuwamoto[***], O. Banno[****], R. Sato[*]

Abstract. At present, the increasing demand for underground streets and tunnels has generated a need for securing a communication system similar to those in deep mine shafts and tunnels.

In a disaster, i.e., fires, the lines of the wire communication system may be damaged. Therefore, a simple portable wireless device is necessary for the prevention of crimes and disasters, and for the preservation of public safety. In this paper we prove through experiments the possibilities of radio communication in tunnels.

1. INTRODUCTION

Every one has experienced that radio broadcasting cannot be heard in a tunnel. A communicating system in such a place is a convenience for daily activities, however, in emergencies it may become vital for survival. Thus, we see that great interest is focused on the tunnel and underground problem. Various authors (Wait and Hill [1][2][3], Delogne [4], Degauque and Demoulin [5] and Emslie and Lagace [6]) have studied tunnel propagation and recent research in this area has been promising (Murphy and Parkinson [7]). In a disaster, such as a fire, conventional wire communication systems may become unreliable, necessitating a wireless radio system.

This paper proves that communication is fully possible in tunnel. We had been searching for a place where radio wave noises were minimal or nonexistant in order to perform experimental studies. The inside of a long tunnel seemed to fulfill the necessary requirements. However, during the experiments, it was discovered that the tunnel itself was a transmission line of high pass type [8], similar to the circular waveguide.

The three main purposes for the present study were,

* Tohoku University, Japan
** Yagi Antenna Co., Ltd. Japan
*** Hitachi Ltd. Japan
**** Hitachi Denshi Ltd. Japan

first, to experimental prove the possibilities of radio communication in the tunnel second, to obtain attenuation constant, and third, obtain phase constant, theoretically and experimentally.

2. Experimental Communication and Results

Experimental Materials

A straight 1470 meter long tunnel constructed of ferroconcrete was employed.
The dimensions of the cross section are provided in Figure 1.
Yagi-Uda antennas or half-wavelength dipole antennas were used for both stations.
The height of the antenna was 3.6 meters above the ground (Figure 2).

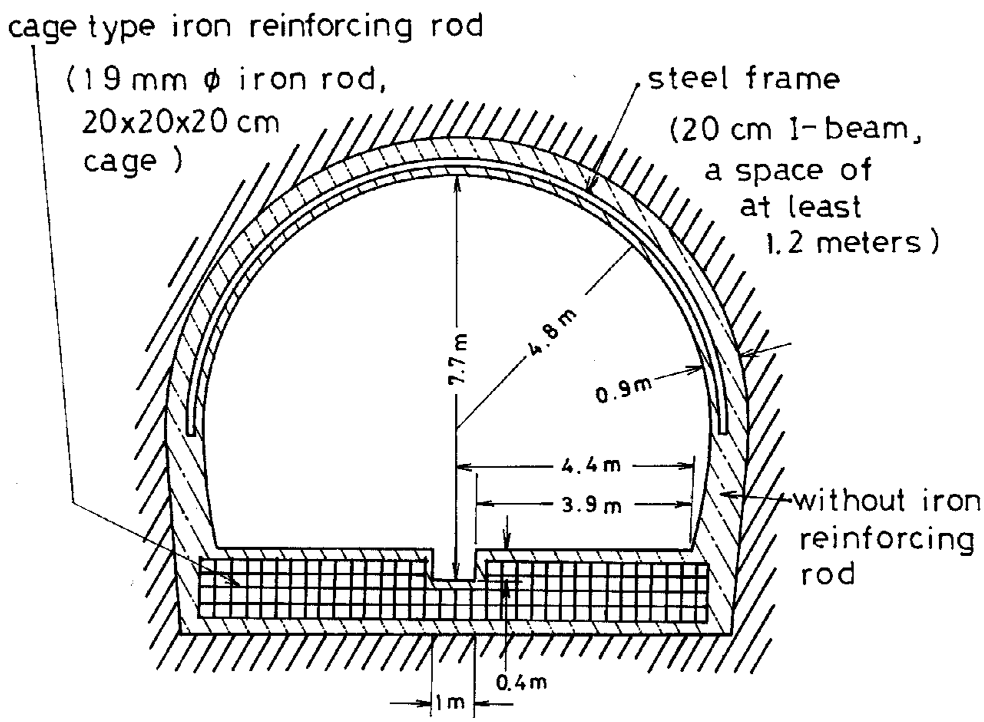
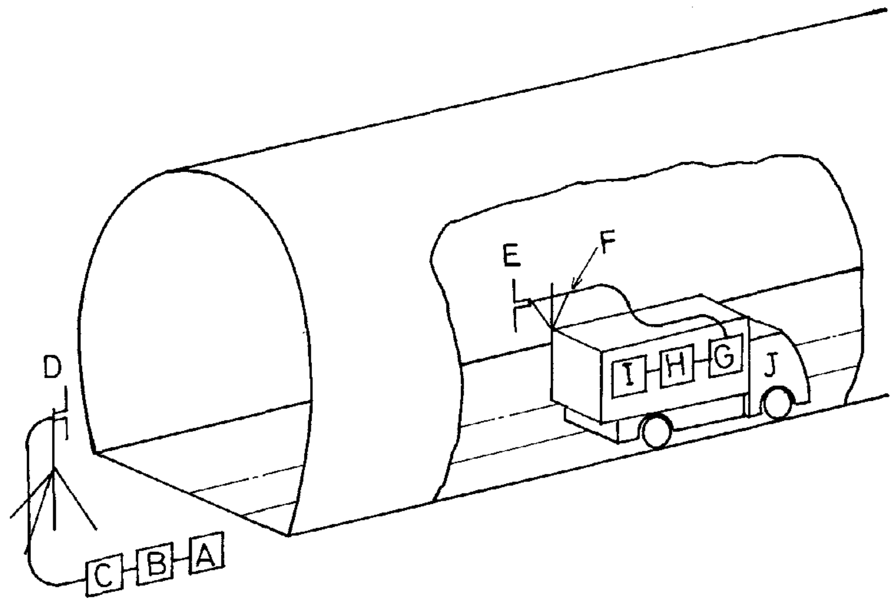


Fig. 1 . Cross-sectional view of tunnel in which measurement were mode.



A : alternating generator
B : oscillator
C : power amplifier
D : transmitting antenna
E : receiving antenna
F : sperrtop
G : field strength meter
H : pen-writer recorder
I : alternating generator
J : 3-ton truck

Fig. 2. Experimental arrangement for measuring the attenuation constant.

The transmitter and receiver, a wireless telephone commonly found in taxicab radios, was used at a frequency of 470 MHz with a transmitting strength of approximately one watt.

Experimental Method

A fixed station and antenna were set 30 meters outside the entrance of the tunnel and later, inside the tunnel, and a mobile station and an antenna were placed on a track and moved along a solid line during the experiments (Figure 2). The portable wireless telephone experiment is shown in Figure 3.

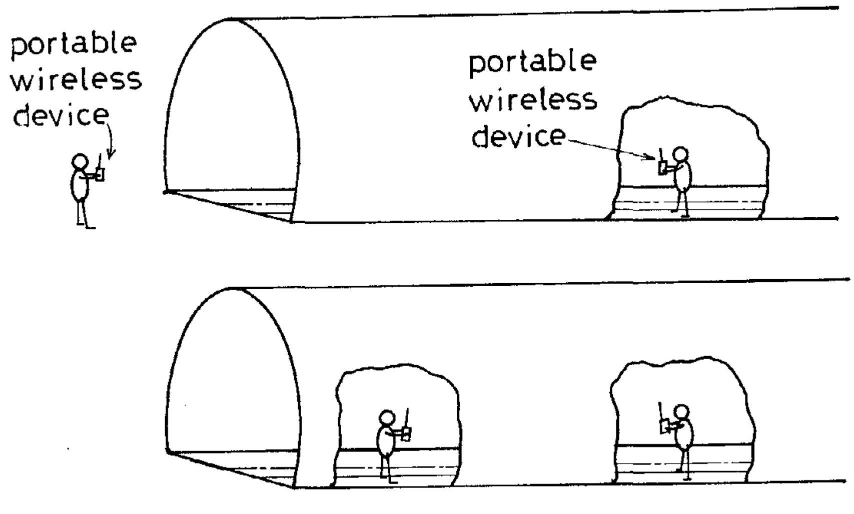


Fig. 3. Communication experiment.

Experimental Results

Experimental communication under the above-mentioned conditions well as when the antenna of the fixed station was moved into the tunnel was as intelligible as a public telephone.
Although the distance between the transmitting and receiving antenna is varied, large variations in communication could not be found.
Hence, it is conceivable that even in a longer tunnel complete communication is possible.
A communication experiment using a 27 MHz frequency approaches the cutoff frequency as a high pass transmission line as illustrated later by the circular waveguide.

3. Determination of Attenuation Constant by the Experiment

Experimental Methode

A transmitter was set outside the tunnel and the radio wave was sent into the tunnel (Figure 2).
While the reception equipment in the tunnel was moved on a truck along the tunnel, the output of field strength measurement is continuously recorded by a pen written recorder.
In this manner, the variation of field strength against the distance in the tunnel was found and

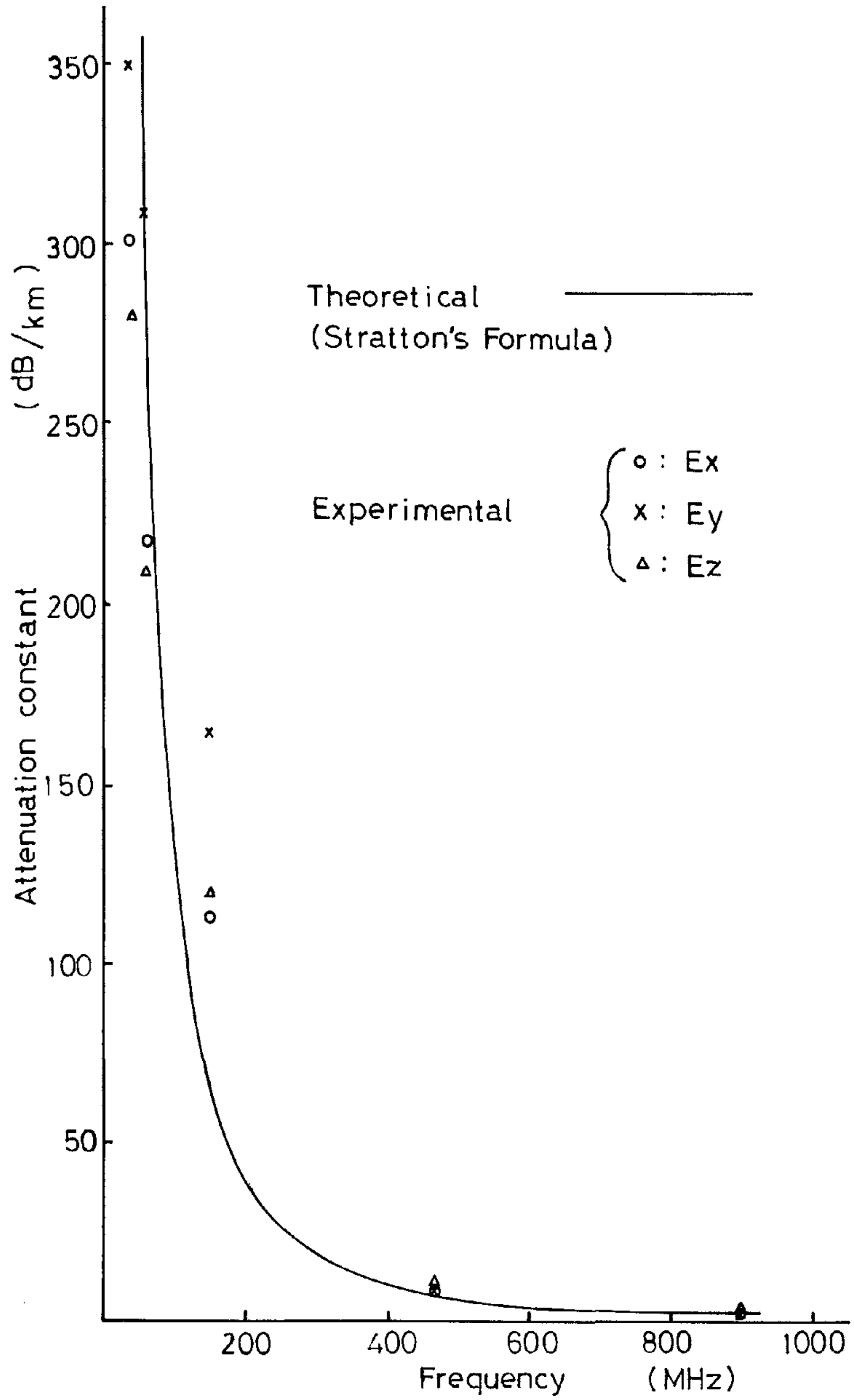


Fig. 4. The attenuation constants.

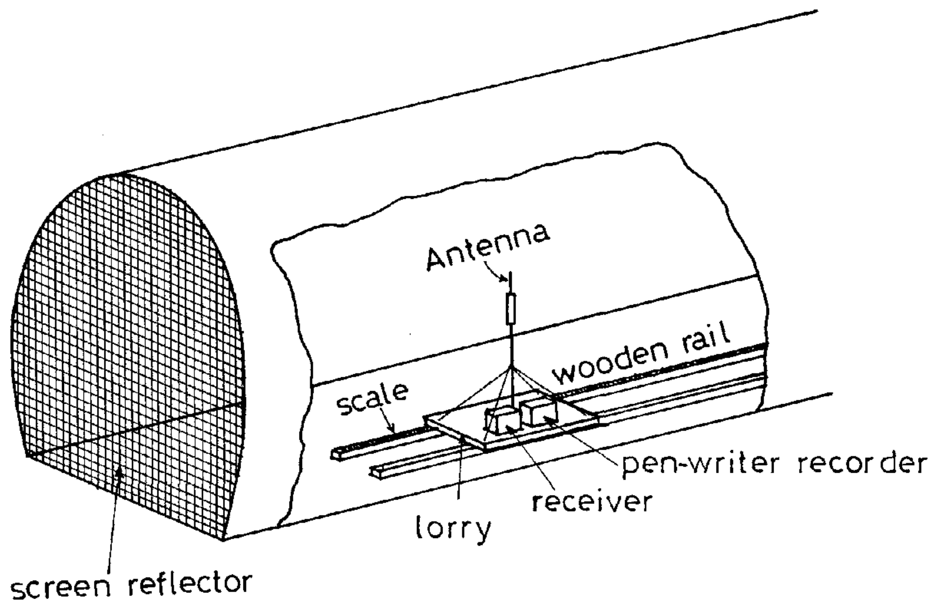


Fig. 5. Experimetal arrangement for measuring the wavelength .

immediately the attenuation constant was determined. The method of least squares was used for date treatment.

Experimental Result

The attenuation constant is shown Figure 4.
As the the frequency becomes gradually higher, e. g., 900 MHz, becomes several dB/km.
According to these results, it becomes clear that the tunnel was one kind of high pass transmission line which was similar to a circular waveguide.
According to theoretical considerations, experimental results correspond to the theoretical value of H_{01} mode (because of space consideration, the details are omitted).

4. Determination of Wavelength in the Tunnel

The experimental arrangement used is shown in Figure 5. Screen reflector acts as a short plate. A fixed station and antenna were set inside the

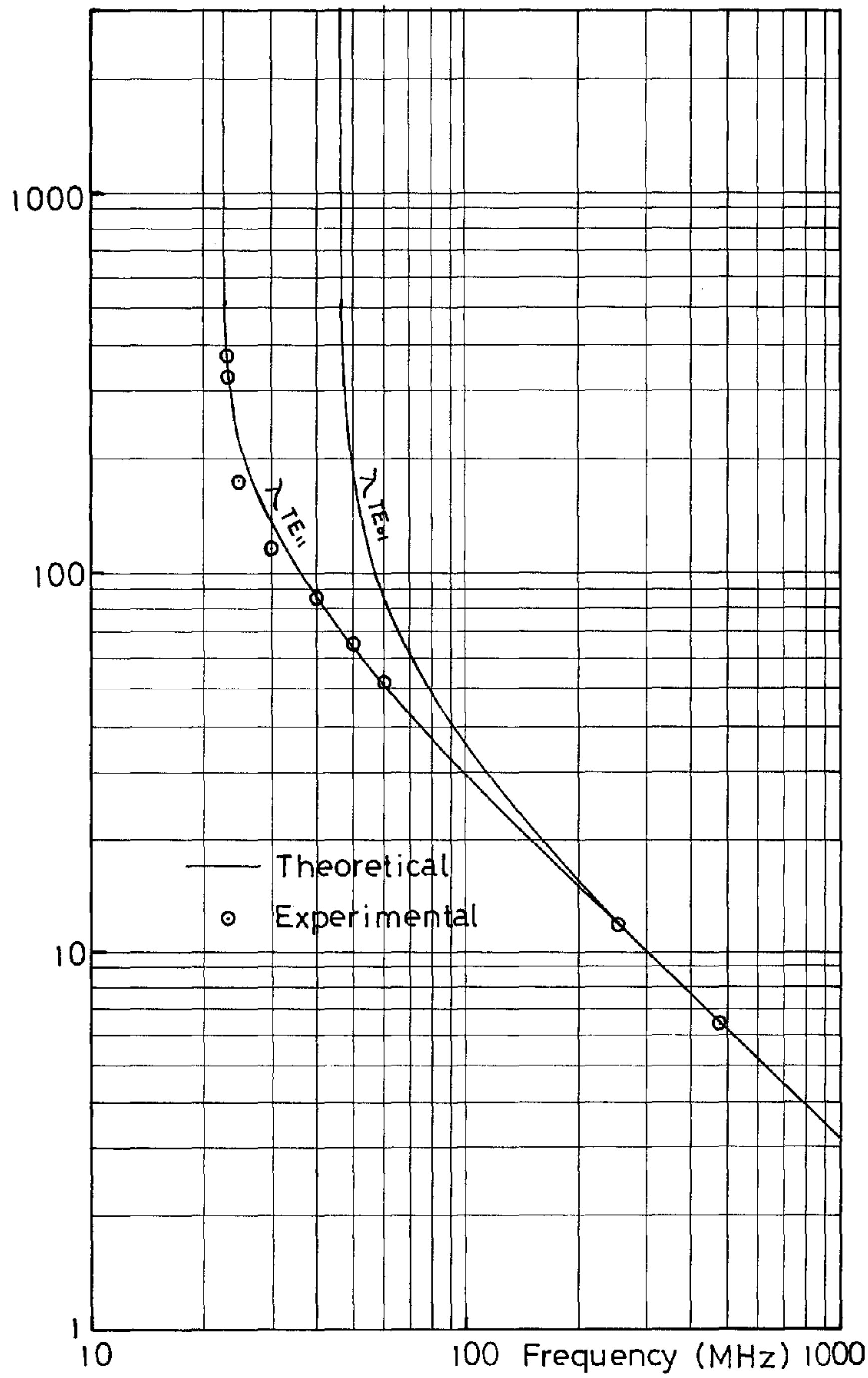


Fig. 6. Wavelength in the tunnel.

tunnel and the radio wave sent towards the screen reflector. The receiver and receiving antenna were placed on a lorry and moved along the rail during the experiment and the output of receiver is continuously recorded by a pen written recorder. In this manner, the standing wave in the tunnel was found and immediately the wavelength in the tunnel was determined (Figure 6). Wavelength increases with decreasing frequency and approaches infinity near cutoff frequency. The problem involving modified cross section can be solved by means of Rayleigh-Ritz's method [9], on the assumption of perfect conducting wall . The results for the lowest order TE_{01} and TE_{11} mode are shown in Figure 6. The experimental values agree well the theoretical results. The lowest cutoff frequency (TE_{11} mode) is 23.5 MHz.

Conclusions

By this study:

1. It was proved that radio communication using the tunnel was fully possible.
2. Attenuation constant of the tunnel could be obtaind.
3. Transmission loss and cutoff frequency, etc., can be explained by regarding the tunnel as a circular waveguide in which cross sectional areas are equal.
4. If the diameter of the tunnel is from several times to ten times longer than the transmitting wavelength if free space, the tunnel can be used as a transmission line of wave.

Acknowledgment

Authors would like to acknowledge the continuing guidance of Dr. T. Nimura, Tohoku University.

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COMMENT BY P. DELOGNE

We obtained similar results on tunnel transmission in Belgium and in France. It is clear that the attenuation decreases with increasing frequency but shadow effects due to obstacles between the transmitter and the receiver seem to increase too. At 450 MHz, we observed attenuation of 20 to 30 dB for a train located between the transmitter and the receiver.

QUESTION BY J.R. WAIT AND B.A. AUSTIN

How was the conductivity and permittivity measured and was there a frequency dependence?

REPLY BY J. CHIBA

A standing wave method was used in conjunction with a planar concrete wall of the same material as the tunnel. The three frequencies employed were 100, 900 and 1700 MHz for these tests. There was no appreciable frequency dependence.

QUESTION BY R. GABILLARD

Have you studied the mode conversion between the TE_{01} and the TE_{11}?

REPLY BY J. CHIBA

Under investigation; it is an important problem in curved regions of the tunnel.

THE TESTING OF LEAKY COAXIAL CABLES AND THEIR APPLICATION TO GUIDED RADAR

N.A.M. Mackay and J.C. Beal
Department of Electrical Engineering
Queen's University
Kingston, Ontario, Canada.

I INTRODUCTION

The use of leaky coaxial cables has become increasingly important in recent years in the provision of continuous two-way mobile communications in mine and tunnel environments. A recent application that has possibly equal importance in the long term is the detection of objects along some defined linear path. Prominent examples are the detection of obstacles on ground transportation systems such as railroads, and urban transit systems, and the perimeter surveillance of installations that must be held secure, such as prisons, airports, and certain industrial installations. This feature of restricted detection along some prescribed path has led to the use of the term "Guided Radar" [1].

Common to both these areas of application for leaky coaxial cable is the need for some reliable method of testing the important properties of various types of leaky cable. Perhaps the most significant and most difficult of these properties to test is the "coupling", i.e. the extent to which the fields being guided along the leaky cable are accessible from nearby points outside the cable.

This paper describes a novel method for the reliable testing of leaky cables in a laboratory which can provide reasonable predictions of their performance in a full-scale application. While this test has particular relevance to the use of leaky cables in Guided Radar, it also provides a general laboratory technique for the testing of the "coupling" performance, yielding information that is also useful in other applications. In doing so, it replaces the need for tests involving field strength meters and long cable lengths, with all their inherent problems of repeatability and allowance for local ground effects.

The basic concept of a Guided Radar system, as developed at Queen's University and Computing Devices Company, Ottawa, (under the name GUIDAR), is first discussed and is then followed by a description of the test procedure together with an outline of its theoretical basis.

II A GUIDED RADAR SYSTEM

Figure 1 illustrates the basic concept. Two leaky coaxial cables are laid in parallel, typically about 1 to 3 metres apart, along a prescribed linear route which may include bends and curves as necessary. For protection, the cables may be buried a few cm. in the ground. Connected to one cable is a pulsed transmitter at a V.H.F. carrier frequency, while a receiver with appropriate signal processing is connected to the second cable. Any object located along the prescribed route between or immediately adjacent to the cables intercepts that part of the signal accessible outside the leaky cable and sends back towards the source end a reflected signal, some of which couples into the second cable and is passed to the receiver. In the receiver itself a moving-target-indicator type of signal processing takes place which reduces the "clutter" and thus enhances the detection of the wanted moving target.

This process is outlined in Figure 2. In the absence of a target there is still a relatively large background or clutter which is digitized for each distance cell along the route and stored in a microprocessor, usually in a series of "windows" or regions of interest. Any moving target then introduced produces a very small time-varying change from this stored profile, which is continually updated, and by subtraction of the immediately preceding profile a greatly enhanced target return appears as the desired output. The continual updating of the profile enables the effects of slowly varying environmental changes to be removed.

A prototype system has been installed in a Canadian penitentiary, between the double chain-link fences that surround the perimeter, and many tests have been conducted on the detection of human beings.

III FIELD-SITE TESTS OF GUIDED RADAR

An almost full-scale field site is available at Queen's University and extensive tests have been conducted on systems having lengths greater than 100m. A typical test result is shown in Figure 3 which illustrates the system's sensitivity to the presence of a human being walking mid-way between the two leaky cables, for two representative spacings between the two leaky cables (1.8m and 3.5m). It is important to note the decreased sensitivity mid-way between the more widely spaced cables. Both curves also show the effect of cable attenuation along the route, as modified by the environment of the grassy ground on which they were laid.

A further set of tests is shown in Figure 4 a,b, and c. Here a human walked across the two cables at an arbitrarily chosen point along the route. These three diagrams again illustrate the effect of spacing on sensitivity; the 'M'-shaped sensitivity plot begins to appear at the wider spacing with an associated lowering of the sensitivity mid-way between the cables. This

'M'-shape is then considerably enhanced when the carrier frequency is raised from 60 MHz to 120 MHz; i.e. it appears that increasing the frequency has roughly the same effect as increasing the spacing, which is a well-understood phenomenon for co-directionally coupled open waveguides such as Goubau lines [2].

The characteristic 'M'-shape is closely similar to the transverse variation of the theoretical guided power density contours for a simple two-wire TEM transmission line, as shown in Figure 5, if we ignore the complication of the presence of the ground on which the cables are laid. As a first step, then, this suggests that the mechanism involved may be as follows:

(i) The sending leaky cable couples the signal co-directionally into the two-wire transmission line formed by the two outer conductors of the parallel leaky coaxial cables;

(ii) The signal now on this two-wire line is intercepted by the target and a reflected signal travels back along this same two-wire line;

(iii) As the reflected signal travels back towards the source end, it couples part of the reflected signal codirectionally into both coaxial cables and that part coupled into the second cable is detected at the receiver.

Many tests have been conducted over a long period to assess the performance of this Guided Radar system for a variety of human and other targets and in the presence of a wide range of environmental conditions.

IV THE TWO-CABLE CAVITY RESONATOR

The above description of the Guided Radar system suggests a technique for testing leaky coaxial cables by the introduction of a standard target, such as a large vertical metal plate connected to the outer conductors of the two coaxial cables so as to form an effective short circuit on the two-wire transmission line. For a laboratory test with only short lengths of leaky cables, the extraneous effects of laboratory walls, etc., can then be removed by adding a second vertical plate, parallel to the first, at the sending end yielding, in effect, a resonant cavity on the two-wire line.

This two-cable cavity resonator is shown in Figure 6. It has two vertical, parallel, aluminum end-plates, each 1.8m × 1.8m, spaced approximately 5m. apart. Two lengths of the leaky coaxial cable to be tested are placed spatially parallel between the end plates. At each end-plate each cable is connected to a conventional feed-through, type N, panel-mounting coaxial connector. On one end of each cable is connected, outside the end-plate via ordinary RG-8 coaxial cables, a wattmeter and a V.H.F./U.H.F. continuous-wave sweep oscillator, respectively. At the other end of each cable a conventional 50Ω coaxial load is connected to each feed-through connector, external to the end-plate.

The combined effects of the two metallic end-plates are confined to the space between them, and to that immediately around the two cables. The arrangement is now suited to use in a laboratory with negligible interference from nearby objects inside the building. Thus we have, in principle, a four-port network whose properties can be measured over a wide frequency range by use of a network analyser. These properties are intimately related to the fundamental coupling behaviour of each leaky cable and the manner of its variation with cable spacing and frequency throughout the V.H.F. and U.H.F. bands. Although the cavity length is only 5m., useful measurements can be made down to as low a frequency as 30 MHz. Future work will include the use of a controlled adjacent ground plane, introduced by means of a horizontal aluminum plate brought under the two horizontally disposed leaky cables. With the addition of material such as wet sand, it should become possible similarly to investigate the modification of the coupling behaviour of leaky cables when laid on, or buried in, imperfect lossy ground of varying wetness.

V TWO-CABLE CAVITY TEST RESULTS

For each pair of cables of a given type, the laboratory cavity test yields a characteristic cable "signature" for the particular test spacing, i.e. a plot of the power received by the wattmeter relative to the input power as a function of frequency. An example is shown in Figure 7 for two RADIAX RX4-1 cables (Cable C) spaced 0.9m. apart. The many narrow peaks indicate the cavity resonances spaced at intervals of about 29 MHz, corresponding to a relative phase velocity of 98% for the two-wire line formed by the outer conductors of the two leaky coaxial cables. The amplitudes of all these resonances vary according to an overall "envelope" (shown as an estimated dotted line) with minima at approximately every 270 MHz. The shape of this "envelope", i.e. the spacing between minima, is strongly linked to the difference in phase-velocity between that of the single, isolated, leaky cable and that of the two wire line formed by the two outer conductors. The overall level of the "envelope" and its associated peaks is similarly strongly linked to the "coupling" parameter for this type of leaky cable at this particular value of cable spacing. This graph is rather noisy because of the low level of coupling that occurs for this type of cable (RX4-1), which has only quite small holes in its outer conductor. The tendency for the "envelope" peaks to decline in value at higher frequencies is an indication of smaller coupling at these frequencies, as would be expected for an open waveguide (cf. the Goubau line).

Figure 8 shows the corresponding "signature" for RX4-3 RADIAX leaky cable, a similar type of cable but one having considerably larger holes in the outer conductor. The main difference from Figure 7 lies in the much higher relative received power, an indication of a much higher "coupling" parameter, as would be expected for this cable type with its larger holes in the outer conductor.

Figure 9 gives the "signature" for an experimental type of RX4-3 having the same hole size but a lower density foam dielectric inside the cable (to lower the attenuation). The "envelope" peaks are at the same level as for normal RX4-3 implying that the "coupling" parameter remains unchanged. The spacing between "envelope" minima, however, is greatly expanded and is associated with a higher phase velocity for this version of RX4-3 with a low-density dielectric, as might be anticipated.

Many other cable types have been tested in this cavity and their relative "coupling" parameters have been found, enabling comparisons to be made and predictions to be formed about their behaviour in full-scale tests in a Guided Radar system, all on the basis of a relatively simple and entirely repeatable laboratory test. In a communications application the nearby mobile transmitter/receiver is typically at a distance from the leaky cable comparable to the cable spacing in this cavity test. It follows that the cavity test results should also be useful in predicting the relative effectiveness of various types of leaky coaxial cable in communications systems.

VI OUTLINE OF THEORY OF CAVITY TEST

The coupling of energy among the two leaky cables and the two-wire line formed by the two outer conductors is fundamentally co-directional and can be analysed by an extension of the usual odd and even mode representation of two coupled waveguide systems to the slightly more complicated case of three coupled guiding systems [3].

We have, in principle, for the total voltages on each line in the cavity the following:

Coax line (1):

$$E_1(z) = E_{11}^{+} e^{-j\beta_1 z} + E_{11}^{-} e^{+j\beta_1 z} + E_{12}^{+} e^{-j\beta_2 z} + E_{12}^{-} e^{+j\beta_2 z} + E_{13}^{+} e^{-j\beta_3 z} + E_{13}^{-} e^{+j\beta_3 z}$$

Two-wire line (2):

$$E_2(z) = E_{21}^{+} e^{-j\beta_1 z} + E_{21}^{-} e^{+j\beta_1 z} + E_{22}^{+} e^{-j\beta_2 z} + E_{22}^{-} e^{+j\beta_2 z} + E_{23}^{+} e^{-j\beta_3 z} + E_{23}^{-} e^{+j\beta_3 z}$$

Coax line (3):

$$E_3(z) = E_{31}^{+} e^{-j\beta_1 z} + E_{31}^{-} e^{+j\beta_1 z} + E_{32}^{+} e^{-j\beta_2 z} + E_{32}^{-} e^{+j\beta_2 z}$$

$$+ E_{33}^{+} e^{-j\beta_3 z} + E_{33}^{-} e^{+j\beta_3 z}$$

where $z = 0$ to ℓ along the cables in the cavity.

where E_{rs}^{+} = s^{th} coupled mode voltage on the r^{th} line in the forward direction.

where E_{rs}^{-} = s^{th} coupled mode voltage on the r^{th} line in the reverse. direction.

where $\beta_1, \beta_2, \beta_3$ = propagation coefficients for the three coupled modes.

These three β values are each solutions of the six eigenvalue equations

$$\beta E_1^{\pm} = H_{11} E_1^{\pm} + H_{12} E_2^{\pm} + H_{13} E_3^{\pm}$$

$$\beta E_2^{\pm} = H_{21} E_1^{\pm} + H_{22} E_2^{\pm} + H_{23} E_3^{\pm}$$

$$\beta E_3^{\pm} = H_{31} E_1^{\pm} + H_{32} E_2^{\pm} + H_{33} E_3^{\pm}$$

where H_{pq} = coupling coefficient between the p^{th} and q^{th} lines.

where H_{pp} = (slightly perturbed) propagation coefficient of the p^{th} line.

where $E_n(z)$ = $E_n^{+} + E_n^{-}$, the sum of the total forward waves and the total reflected waves on line n .

The six eigenvalue equations for each of $\beta_1, \beta_2, \beta_3$ yield 18 equations of which only 12 are independent. There are 18 unknown E_{rs} quantities, the remaining 6 of which are found from the known 6 boundary conditions on each of the three lines at each end. The two leaky cables are effectively connected to 50Ω matched loads at each end, with a known source applied to one of them, while the two-wire line is connected to a good, but not perfect, short-circuit at each of the two aluminum end-plates. Thus we have an 18×18 matrix equation, in principle, for which in practice a number of simplifying approximations can be made. The eigenvalues then yield the three previously unknown β values and the theoretical value for the relative coupled power, as measured, can be calculated given assumed values for the H_{pp} and H_{pq} values. In practice, these parameters can be chosen so as to give the best fit with the observed behaviour, thus enabling the basic "coupling" parameter, H_{12}(= H_{21} = H_{23} = H_{32}, by symmetry, for co-directional coupling) to be identified.

VII CONCLUSION

A Guided Radar system has been described that enables obstacles or intruders to be detected along some prescribed path, by the use of pairs of parallel leaky coaxial cables.

A laboratory test for leaky coaxial cables has been developed which yields information on the relative "coupling" effectiveness of various types of leaky cable. This test enables one to make useful predictions of the relative performance of these cables in a guided radar system and, by extension, in a leaky cable communications system. A simple comparison of the measured guided radar sensitivity to a human target with the relative "coupling" in a laboratory cavity test is shown in Figure 10. The straight line relationship revealed for the three commercially available types of RADIAX cable, having different sizes of holes in the outer conductors, is an indication of the usefulness of this new testing procedure.

VIII ACKNOWLEDGEMENTS

The authors gratefully acknowledge the valuable work of the members of the "GRIPS" (Guided Radar Information Processing System) group at Queen's University.

This work has been supported by Computing Devices Company, Ottawa, the Canadian Institute of Guided Ground Transport at Queen's University, and the National Research Council of Canada.

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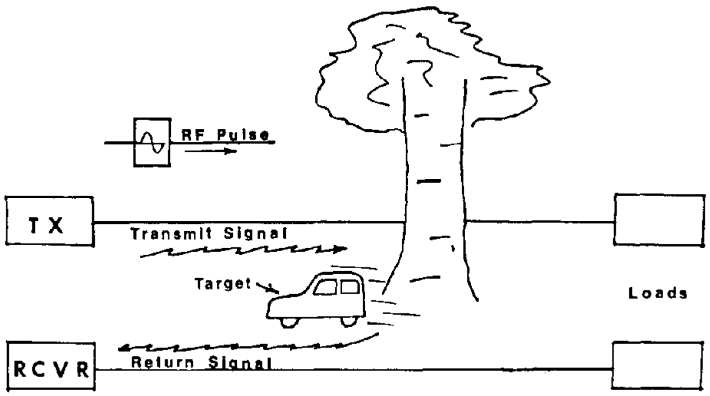

FIGURE 1: GUIDED RADAR CONCEPT.

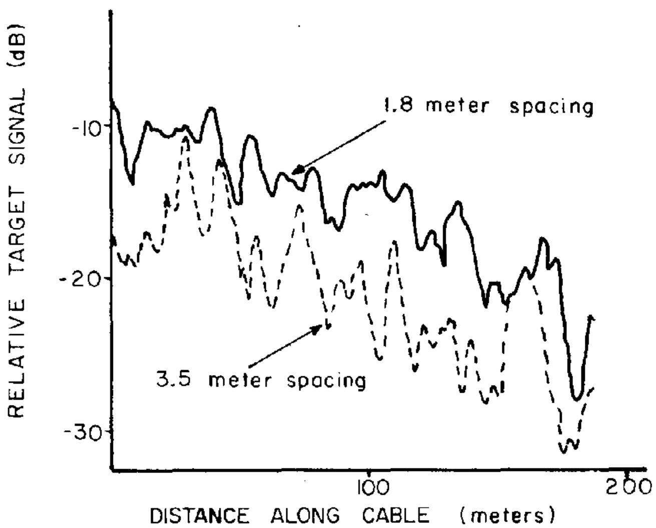

FIGURE 3: LONGITUDINAL SENSITIVITY.

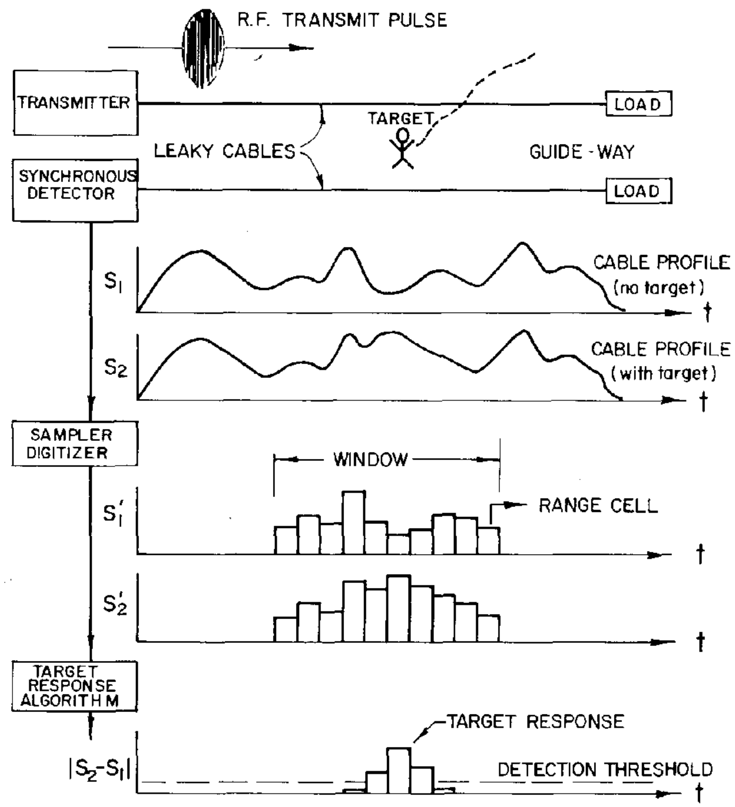

FIGURE 2: DESCRIPTION OF THE GUIDED RADAR CONCEPT.

60 MHz

1.5 meter spacing

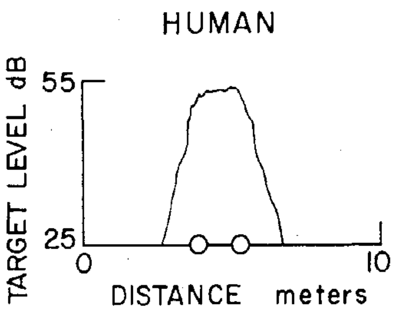

FIGURE 4a: OBSERVATION.

120 MHz

3.0 meter spacing

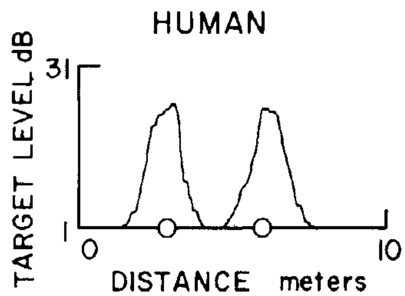

FIGURE 4c: OBSERVATION.

60 MHz

3.0 meter spacing

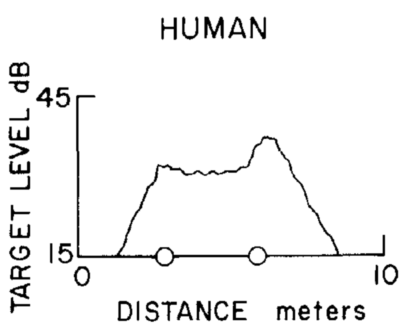

FIGURE 4b: OBSERVATION.

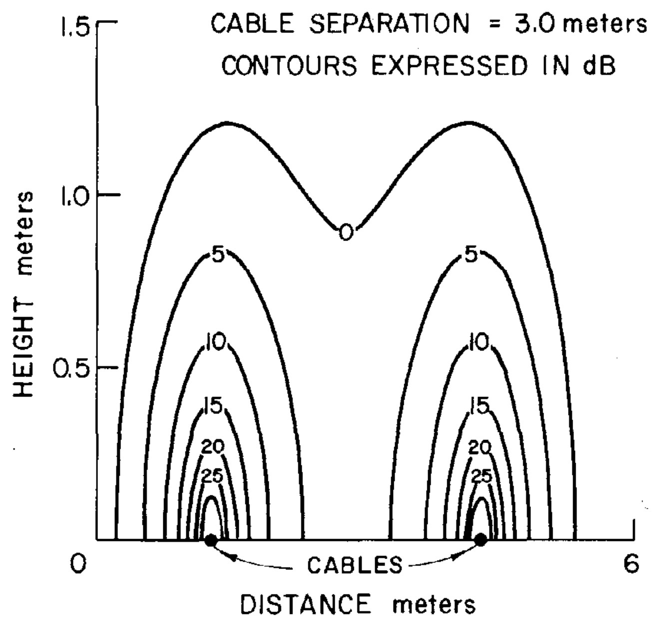

FIGURE 5: THEORETICAL POWER DENSITY DISTRIBUTION.

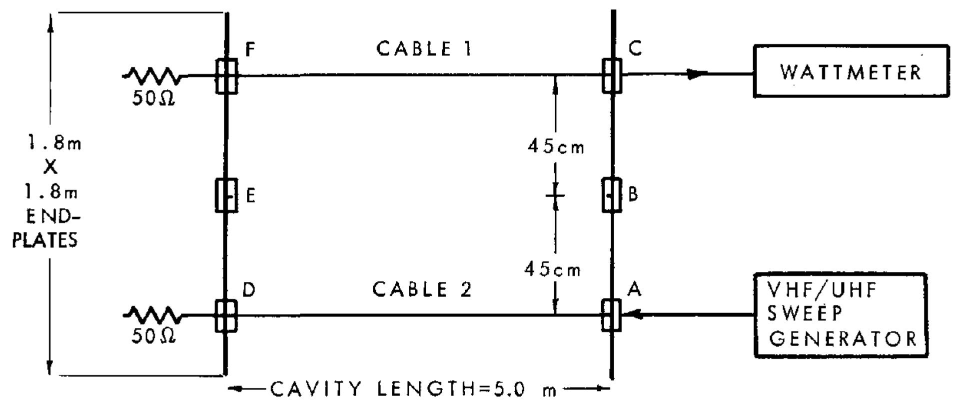

FIGURE 6: TWO-CABLE CAVITY RESONATOR.

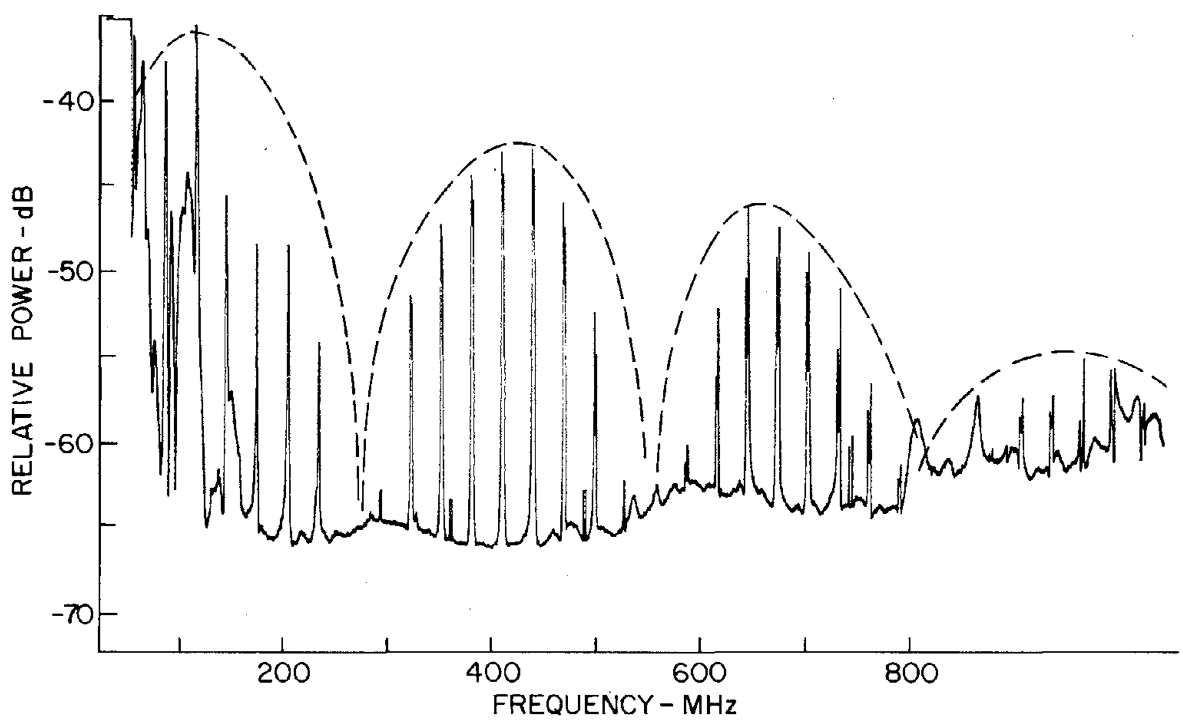

FIGURE 7: CABLE C TEST SIGNATURE CABLES SPACING: 90 cm.

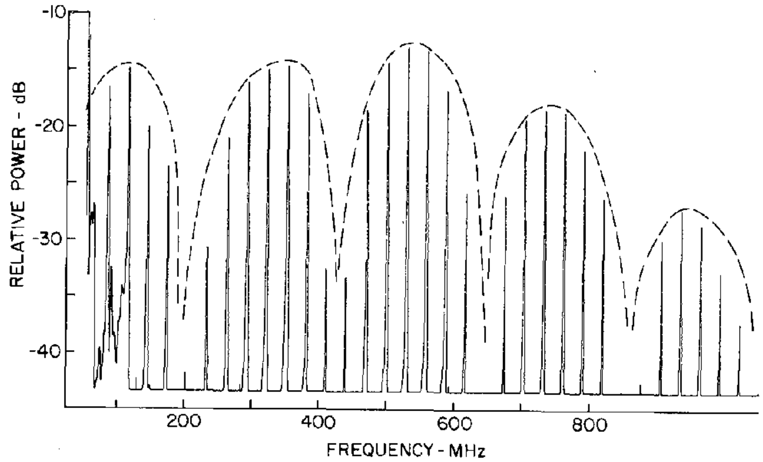

FIGURE 8: CABLE B TEST SIGNATURE CABLES SPACING: 90 cm.

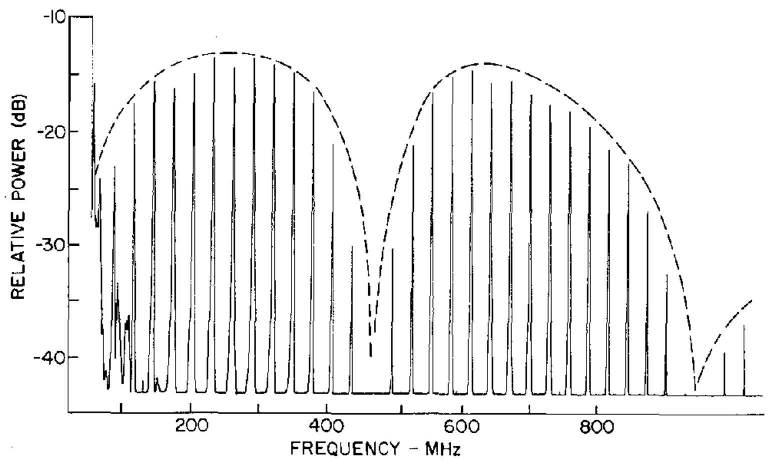

FIGURE 9: CABLE A TEST SIGNATURE CABLES SPACING: 90 cm.

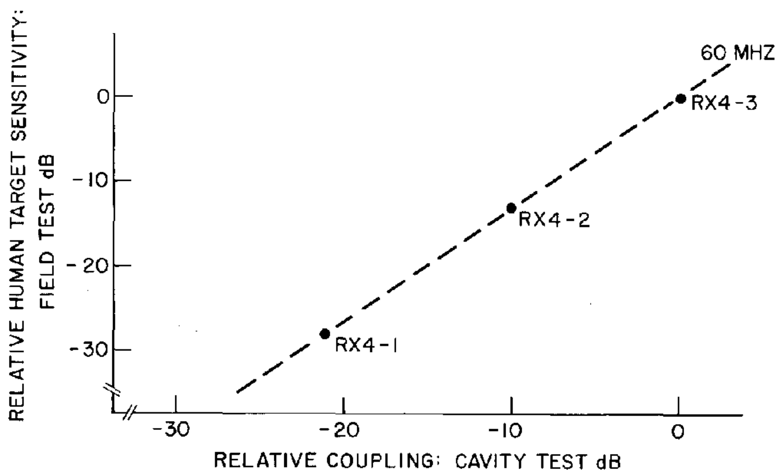

FIGURE 10: PERFORMANCE PREDICTION OF LEAKY COAXIAL CABLES.

QUESTIONS BY A.J. FARSTAD

1. What difference in sensitivity was observed between the above ground cables and the cables lying on the ground?
2. Would it be possible to bury the cables below the ground and retain the sensitivity of the device?

REPLIES BY J.C. BEAL

1. About 10 to 15 dB, which is no more than the fluctuations observed on the ground anyway due to minor imperfections.
2. Yes. This is, in fact, done in the perimeter surveillance application. Burial depth of about 6 inches in well-drained gravel is best.

QUESTIONS BY J. DURKIN

1. How is the threshold detector set?
2. What false alarm rate is expected?
3. If stationary background information is updated continually, then isn't it possible for an individual to cross over at a rate which would permit him to appear as a stationary target?
4. It was stated that it would take 20 mins. total cross over time for the individual to appear stationary. However, this might be an acceptable delay to an individual who has 20 more years to serve!
5. How does the system perform as a function of changes in weather such as rain?

REPLIES BY N. MACKAY

1. The threshold is set on a variety of different constraints and is dependent on an algorithm which is influenced by: receiver noise power, rate of change of profile, amplitude and phase of profile, and system sensitivity.
2. The false alarm rates for the early GUIDAR models were measured at about 1 per 4 days. Later models have much lower rates and false alarms are now attributed to equipment failures and not to the ability (or viability) to identify legitimate or false alarms. The customer's specifications on the early models called for a FAR of 1 per 2 days; the new models call for 1 per month.
3. Only if this rate is slow enough to be called "stationary". As might be expected, some of this information is confidential but, typically, the crossing would have to take over 20 minutes in some models to be undetected.
4. True, but the point is that all institutions of this type are under heavy observation as well. The chances of an inmate in a penitentiary have 20 minutes to escape are very low. In some Canadian prisons, the mean time to escape over or under two fences is about 30 seconds.

5. Changes in weather (e.g. rain) tend to have an effect on all range cells and is normally easily distinguished from human intruders who disturb only a portion of the sensor. In addition, the influence of the environment is fairly well known and the threshold and detection algorithms tend to take advantage of this knowledge.

QUESTION BY D. MARTIN

In regard to the laboratory test technique described:

1. Assuming a separate knowledge of the propagation velocities involved, would not this be a good way of measuring the surface transfer impedances of a cable braid rather than the standard method?

2. Increasing the coaxial mode velocities by modifying the foam dielectric would be expected to increase the coupling as well as lengthening the velocity-braid wavelength but this is apparently not shown.

3. I suggest that no laboratory test alone can compare the practical performance of leaky feeders which must depend on environmental factors such as monofilar mode attenuation which will influence feeders with different coaxial mode velocities in a variable manner.

4. Rather than reflecting the bifilar modes, would it not be simpler to use a matched termination?

REPLY BY J.C. BEAL

1. Yes. The method described in our paper is essentially a high-frequency guided-wave technique that is extended down to the low VHF band, whereas the "standard" method is really a low-frequency method extended upwards to VHF and UHF. No doubt a numerical relationship could also be developed between the "coupling coefficient" obtained from our method and the "transfer impedance" used by many others. But perhaps it might be better to consider dropping the use of the transfer impedance concept altogether at the higher VHF and UHF frequencies.

2. This is a very interesting point and underlines the unique structure of many available leaky cables. The "leakiness" appears to be a function both of the isolated, relative phase velocity and of the detailed nature of the openings in the outer conductor. Experience has shown that the cablewave or Kabelmetal type, with a continuous longitudinal slot, has to have a quite wide slot width for a given coupling, or field accessibility compared with the relatively small openings in the RADIAX cable. It appears that the manner of disturbance of the outer conductor current is a critical point. From the point of view of least loss, it seems likely that the ideal would be a hypothetical uniform partially transparent outer conductor. We suggest that the loose-braid outer conductor is probably the best practical approach to this ideal case.

3. Future work on the "Cavity Test" will include investigating its ability to predict environmental effects due, e.g. to immersion of the cables in wet sand within the cavity.

4. In principle, yes. But in practice it would be very difficult to make a good broadband matched load for the bifilar mode for the VHF and UHF bands.

COMMENTS BY A.G. EMSLIE

The laboratory transmission line results appear to be exactly analogous to the interferogram produced by an optical Fabry-Perot interferometer for the sodium D lines. The sharp resonances correspond to the case of a single frequency. The beats correspond to interference when two closely spaced spectral lines are present as for sodium light. In the microwave case, the two frequencies correspond to the two modes of the coupled cables. The method also resembles a Doppler police radar which suggests that a simple-cable guided wave might work.

REPLY TO J.C. BEAL

Yes, there is a close similarity between this technique for leaky cables and that you mention for the Fabry Perot interferometer.

In fact, extensive laboratory tests have been conducted on a single cable version of this test. Practical aspects associated with noise and "clutter" (from imperfections in the cable itself) greatly limit the obtainable sensitivity of the measurement. In its two-cable form it is, of course, directly relevant to the two-cable guided radar system, but another way of looking at the test and, indeed, that which actually occurred in its development, is to regard it as a single-cable test with the second cable acting as a precise field probe. Earlier tests with a single cable and separate small dipole or loop field probes, as are often used, were very unsatisfactory and unreliable. The two-cable test has essentially overcame this problem.

A single-cable guided radar system is possible in principle, but is limited in practice by the same noise and "clutter" problems referred to above.

GUIDED WAVES ON WIRES NEAR GROUND

D. C. Chang, E. F. Kuester and S. W. Maley

Electromagnetics Laboratory
Dept. of Electrical Engineering
University of Colorado
Boulder, CO 80309

I. INTRODUCTION

The topic of guided wave propagation along wires parallel to the earth's surface reaches back more than 50 years to the pioneering work of Carson [1] and others. Originally addressed as a transmission line problem, appropriate to frequencies of 1 MHz or less, the propagation constant and attenuation coefficient were determined from a quasi-static mutual impedance between wire and earth, under the assumption that the telegraphist's equations were valid. The past 25 years have seen the need for a more comprehensive theory of such structures valid for arbitrary frequencies, for applications to transients on power lines, guided-radar perimeter surveillance systems, or to earth effects on antennas at HF and above. The foundations of this theory can be found in [2-6]. The present work will summarize the current state of the study of this problem, and indicate some areas for which future research will be important.

II. PROPAGATING MODES ON A SINGLE WIRE ABOVE A PLANE EARTH

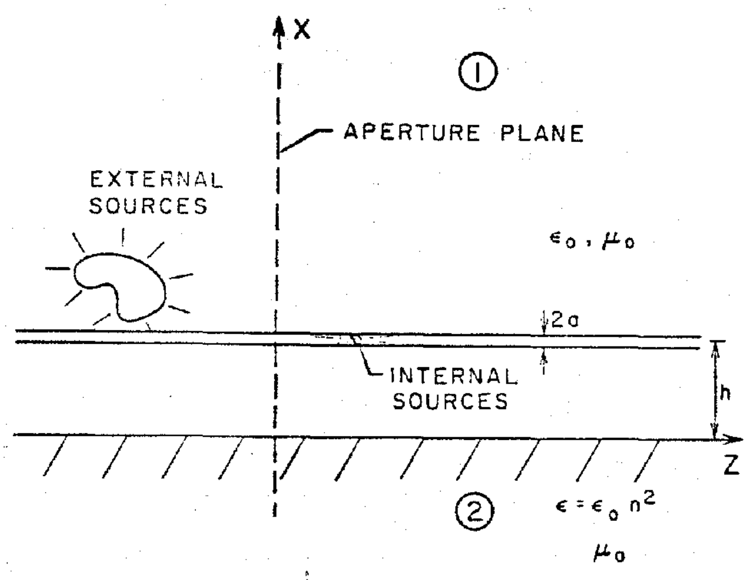

Fig. 1

Consider the structure shown in Fig. 1. A long, thin wire of radius a is located in the air at a height h above a plane, homogeneous earth, which has a complex refractive index n. Although this problem is extremely cumbersome to formulate exactly, considerable simplification can be achieved if proximity effects are neglected, that is, if current on the wire is assumed to be longitudinal and uniform around the circumference of the wire. For most practical situations, this can be shown to be an excellent approximation [7]. By computing the fields of this current radiating in the presence of the half-space as Sommerfeld integrals and setting the average longitudinal E-field equal to zero on the wire, the following characteristic equation is obtained [2-6].

$$L(\alpha) - \frac{\alpha^2}{C(\alpha)} = 0 \tag{1}$$

where α is a normalized propagation constant appropriate to an $\exp(ik_o\alpha z - i\omega t)$ propagation factor ($k_o = \omega(\mu_o\varepsilon_o)^{\frac{1}{2}}$ is the free-space wave number, and

$$L(\alpha) = H_o^{(1)}(\zeta k_o a) - J_o(\zeta k_o a)H_o^{(1)}(2\zeta k_o h) + \frac{2}{i\pi}\int_{-\infty}^{\infty} \frac{e^{-u_1(2k_o h)}}{u_1 + u_2}\, d\lambda \tag{2}$$

$$\frac{1}{C(\alpha)} = H_o^{(1)}(\zeta k_o a) - J_o(\zeta k_o a)H_o^{(1)}(2\zeta k_o h) + \frac{2}{i\pi}\int_{-\infty}^{\infty} \frac{e^{-u_1(2k_o h)}}{n^2 u_1 + u_2}\, d\lambda \tag{3}$$

where $\zeta = (1-\alpha^2)^{\frac{1}{2}}$, $\text{Im}\,\zeta \geq 0$; $u_1 = (\lambda^2 + \alpha^2 - 1)^{\frac{1}{2}}$, $u_2 = (\lambda^2 + \alpha^2 - n^2)^{\frac{1}{2}}$, $\text{Re}\, u_1, u_2 \geq 0$; and $H_o^{(1)}$ and J_o denote the Hankel and Bessel functions of the first kind, respectively. The form of (1) is suggestive of the result of transmission line theory, except that the "inductance" and "capacitance" $L(\alpha)$ and $C(\alpha)$ are spatially and temporarily dispersive (i.e., depend on α and ω), and (1) is an implicit expression which must be solved for α, rather than giving an explicit value for α^2.

If we make the assumption that $|\zeta| \ll 1$, L and C become independent of α, and (1) then yields Carson's [1] solution, which can be shown to be valid for sufficiently low frequencies. An exact solution of (1) generally shows that there is a mode located in the general vicinity predicted by Carson, and this mode, which has fields produced essentially by the wire and an (imperfect) image in the earth, is called a structure-attached mode. In addition to this mode, however, it has recently been found [6,8] that a second solution of (1) is possible, due to the fact that the integral in (3) possesses an inverse square root singularity at $\alpha = n/(n^2+1)^{\frac{1}{2}}$, so that $\alpha^2/C(\alpha)$ can be the dominant term of (1). This second mode generally has a value of α closer to $n/(n^2+1)^{\frac{1}{2}}$ than to 1, and has fields which tend to spread along the air-earth interface rather than concentrate in the earth below the wire. As a result, this mode, which we call earth-attached, has a smaller attenuation, since less of its fields actually propagate within the lossy earth. Some typical results are shown in Fig. 2. Clearly, especially at smaller wire heights, the earth-attached mode is significantly less attenuated, and this combined with the spreading of its fields is a potential advantage for use in detection or communications systems.

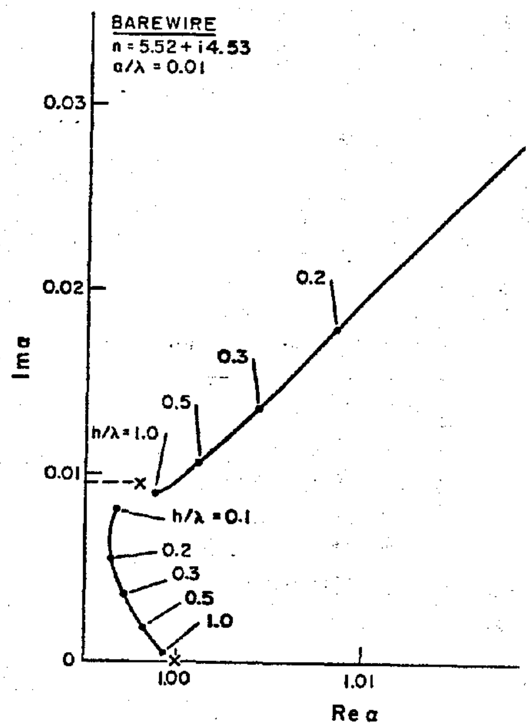

Figure 2

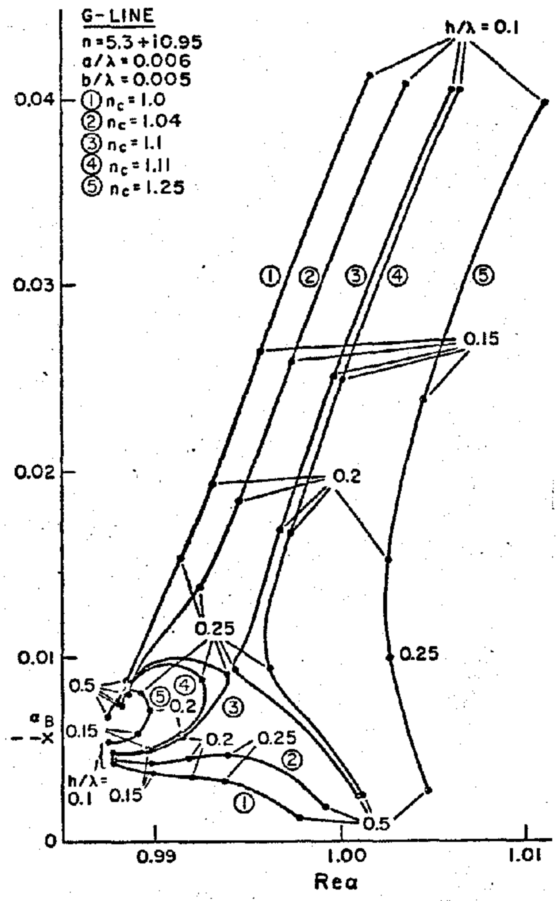

Figure 3

A minor modification of (1) allows us to examine a wire which can be characterized by a surface impedance, including a Goubau line, a wire of finite conductivity, or a leaky coaxial cable. The Goubau line, for instance, also exhibits two sets of modes, as well as the phenomenon known as <u>modal degeneration</u> wherein the propagation constants <u>and</u> fields of the two modes become identical for certain sets of parameters (Fig.3). Energy is readily coupled between two modes which are close to degeneracy, which might be either a desirable or an undesirable property for a given application.

III. DUAL-WIRE LINES ABOVE THE EARTH

The analysis for a single wire is readily generalized to multiwire lines [9-11]. An important case is that of a two-wire line whose wires again have radius a and height h, and are separated horizontally by a distance d. In this situation there are two sets of solutions: "monofilar," where the currents on the wires are identical, and "bifilar," where the currents are equal but opposite. It is typically found that both earth-attached and structure-attached monofilar modes can exist, while only a structure-attached bifilar mode can exist in general. The earth-attached bifilar mode seems only to appear at large wire spacings (Fig. 4).

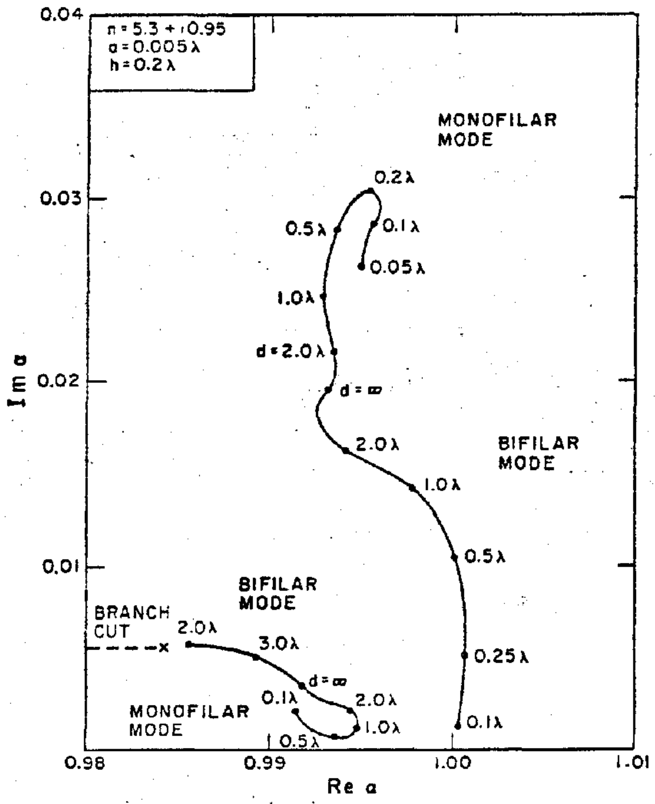

Figure 4

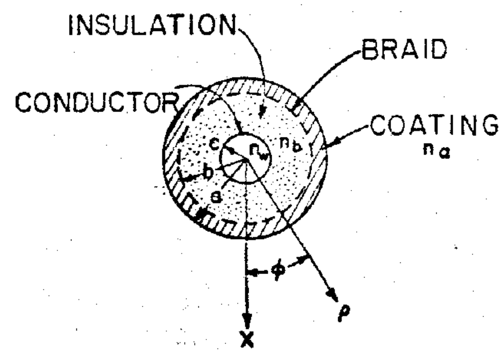

Figure 5

IV. BURIED WIRES AND CABLES

When the guiding structure is buried beneath the surface, the properties of the discrete modes again are modified. A leaky coaxial cable is illustrated in Fig. 5. For the parameters $a = 1.15$ cm, $b = 1.0$ cm, $c = 0.4$ cm, a braid transfer inductance $L_T = 40$ nH, a perfect inner conductor ($n_w \to \infty$) and a coating index $n_a = 1.449$, the behavior of the discrete modes was examined as the insulator refractive index n_b and the depth of burial h were varied. In all cases, at least two types of modes were found. A bifilar mode (currents on braid and inner coaxial conductor nearly equal but opposite) was found to exist near the point at which a TEM mode of the coax would be found if the braid were opaque. This mode has the least attenuation of any of the modes found, since virtually all its fields are concentrated inside the cable. As a result, its propagation constant is less sensitive to the depth of burial h. A second, monofilar mode is found with most of the current flowing in the braid. A similar mode is also found for a buried bare wire. Even for the rather low-loss earth considered in Fig. 6, the attenuation of this mode is quite high, ~12-15 dB/m, although its properties are much more strongly depth-dependent. The bifilar mode suffers from poor field penetration into the surrounding medium and the air, while the monofilar mode, which has the major portion of its fields outside the cable, suffers as a result of the large attenuation.

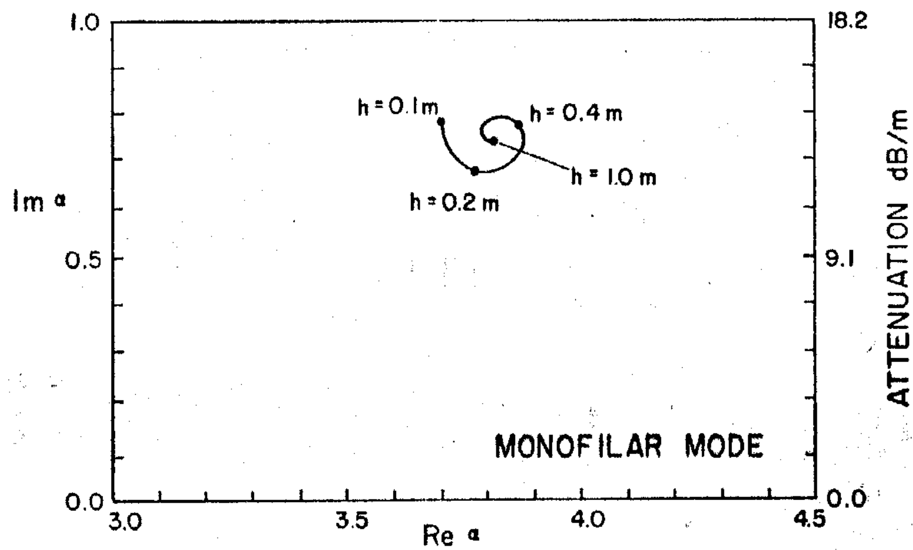

Figure 6. Complex propagation constant for the monofilar mode; $a = 0.0155$ m, $b = 0.01$ m, $c = 0.004$ m, $n_1 = 5.3 + i0.95$, $n_a = 1.449$, $n_b = 1.449$, $L_T = 40$ nH, $f = 100$ MHz.

A third surface-attached mode is sometimes found, whose fields are once again concentrated near the air-earth interface. When the cable is buried, however, the fields of this mode have a close resemblence to the Sommerfeld ground wave and are only slightly perturbed by the presence of the cable. As a result, even though this mode seems to offer the best field penetration into the air region, it would be very difficult to excite, and a disturbance of its fields produces only a weak disturbance of the current flowing on the wire.

V. EXCITATION OF THE WIRE ABOVE EARTH: MODAL DESCRIPTION

A number of special means of exciting the wire above earth have been considered in the literature, notably the delta-function voltage generator and an external Hertzian dipole source [12-14]. The excitation by an arbitrary source distribution can, in fact, be cast in a modal framework, in which the amplitudes of the discrete modes appear as pole residues, and the remaining (radiation) fields appear as branch cut integrals [5]. Orthogonality properties between fields of different modes can also be proven.

The voltage generator in the wire [12] tends to excite the structure-attached mode at low wire heights, while the earth-attached mode and the radiation fields become more important as h becomes comparable to a wavelength.

While the dipole source is perhaps not a very efficient means of excitation of the wire, it is of interest because the scattering from small obstacles can often be modeled as the fields of a dipole, and moreover, in a guided communication system, practical considerations may demand a physically small source. As an example, a VED located on the surface of the earth at a transverse distance b from the wire generates the currents shown in Fig. 7. The total current, as well as the

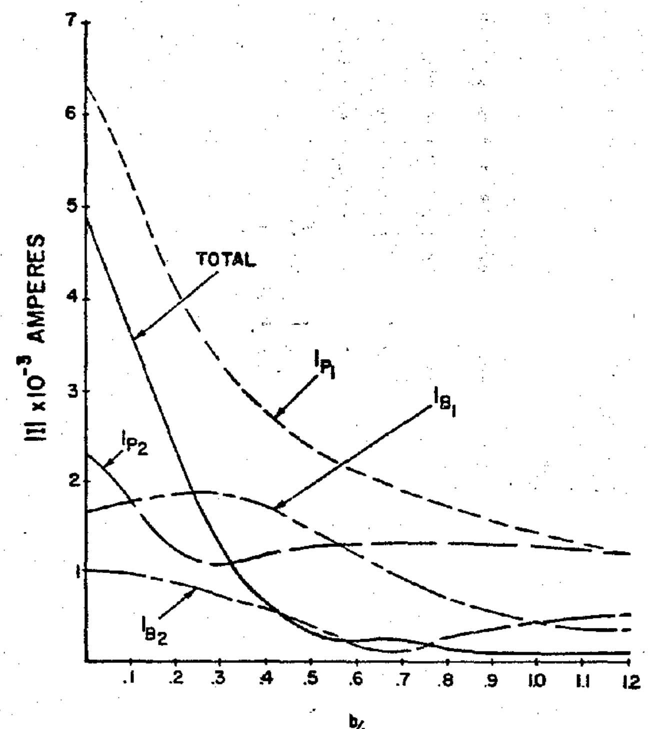

Figure 7. Current induced by a VED on the surface of the earth ($x = 0.0$) as a function of the transverse displacement b for a fixed distance $z = 0.5\lambda$ along the wire; $n = 5.3 + i0.45$, $h = 0.24\lambda$, $a = .007\lambda$.

amplitudes of the discrete mode currents, I_{p1} and I_{p2}, and the radiation currents I_{B1} and I_{B2} are shown, and while it is clear that the term I_{p1} (corresponding to the structure-attached mode) is dominant for distances up to a wavelength, a rather substantial cancellation between the various currents takes place (due to their phase differences) and the total current excited on the wire can be quite small.

Electrically small scatterers can, for some purposes, be modeled by an equivalent dipole moment, proportional to the incident field. This incident field can be taken to be that of either discrete mode, and from the results for a dipole source, the change in the wire current due to the scatterer can be estimated. Results show that changes on the order of several percent can be expected when the scatterer is within a half-wavelength of the wire. Larger percentage change at a longer distance can be achieved, however, when the incident current is primarily that of the structure-attached kind with a higher attenuation constant.

Acknowledgment

We wish to thank Professor James R. Wait of CIRES and Mr. Walter Rotman of RADC/ET for their continuous interests and stimulating discussions during the course of this investigation. This work is supported by Rome Air Development Center/Electronic Deputy under contract no.F19628-76-C-0099 and F19628-77-0093.

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COMMENT BY D.A. HILL

In our circular tunnel work, we have also found cases where the monofilar mode becomes a fast mode. We also have done some calculations for the perturbed waveguide modes which are fast modes. It would be interesting to examine these modes for larger tunnels.

REPLY BY E.F. KUESTER

It would probably be necessary to examine very large tunnels to find a mode which corresponds to the earth-attached mode of the half-space geometry.

COMMENT BY J.R. WAIT

As David Hill indicates, the monofilar mode may become a fast wave mode. However, the complete field in the large tunnel must also include the waveguide type modes. These become very numerous at the higher frequencies (i.e. greater than 100 MHz). One cannot always relate an individual tunnel waveguide mode with a corresponding mode in the planar half-space problem. Of course, in the limit of a tunnel of infinite radius (for a fixed distance of the cable to the tunnel wall), the two problems are physically and mathematically equivalent. Furthermore, the total fields must be the same. A related problem has been discussed in the literature (e.g. see J.R. Wait, "The eccentrically located wire in a cylindrical cavity in a conducting medium and the limit of a planar boundary", Radio Science, Vol. 11, No. 11, 897-899, November 1976).

REPLY BY E.F. KUESTER: I agree.

ELECTROMAGNETIC PROBING THROUGH A THIN, NONUNIFORM LAYER OF CONDUCTIVE OVERBURDEN[3]

By

Richard G. Geyer[1] and James R. Wait[2]

Three important problems in coal exploration, coal mine safety, and through-the-earth communication are the location of washouts, the detection and delineation of channel sands, and the effect of a nonuniform overburden on transmission schemes. The occurrence of the above geologic phenomena may be likened to the presence of a inhomogeneous overburden; that is, one whose thickness or conductivity may vary spatially. To a first approximation, such an overburden may be modelled by a thin sheet whose conductivity-thickness product or conductance σd is a function of coordinate position. This approximation is valid when the overburden thickness is small relative to a skin depth in the material.

Here an analytical formulation is developed for the resultant electromagnetic field of an oscillating vertical magnetic dipole over a thin conductive sheet of infinite extent. The sheet is characterized by a conductance which is a function of the horizontal coordinates. The system of integral equations arising in the general formulations is simplified greatly when azimuthal symmetry prevails. Numerical results for a Gaussian variation of σd in the radial direction are presented for the case of a symmetrically located source. These results are for the fields at the level of the source dipole over the conductive sheet. It is shown that the quadrature response of the sheet is enhanced when there is a rapid variation of the conductance. The null in the resultant wave tilt is also found to be shifted towards the direction of increasing conductance.

[1]Gulf Mineral Resources Co., 1720 S. Bellaire St., Denver, Colorado 80222

[2]Cooperative Institute for Research in Environmental Sciences, University of Colorado/ NOAA, Boulder, Colorado 80309

[3]The full paper is published in Pure and Applied Geophysics, Vol. 116, No. 1, pp. 181-197, 1978.

APPROXIMATE EXPRESSIONS FOR THE QUASI-STATIC FIELDS WITHIN AND OUTSIDE A CONDUCTING SLAB

Peter R. Bannister
Rene L. Dube*

Submarine Electromagnetic Systems Department,
Naval Underwater Systems Center, New London Laboratory,
New London, Connecticut 06320

ABSTRACT. Approximate expressions for the quasi-static fields produced by electric and magnetic dipole antennas located within a conducting slab have been derived by employing finitely conducting earth image theory techniques. Asymptotic results for the fields within and above the slab have also been obtained by applying the quasi-near approximation to the basic Sommerfeld integrals. It is shown that the resulting approximations are in very good agreement with previously derived numerical integration results. The resulting expressions are applicable to the situation where the conductivity of the slab is much greater than the conductivity of the surrounding medium. They are particularly applicable to subsurface to subsurface and subsurface to air propagation for the shallow sea case.

INTRODUCTION

During the past few years, there has been considerable interest in the determination of the quasi-static and quasi-near field components of antennas located above or buried within the earth's surface. The quasi-static range is defined as the range where the measurement distance is much less that a free-space wavelength; the quasi-near range is defined as the asymptotic part of the quasi-static range (i.e., where the measurement distance is much greater than a skin depth in the conducting medium and much greater than the depth of burial of the transmitting and receiving antennas). Quasi-static and quasi-near range results are useful for submarine radio communication and detection purposes, as well as for the buried miner problem. They are also helpful to geophysicists engaged in determining the electrical properties of the earth.

For the semi-infinite conducting medium case (i.e., air, single layered earth), some work has been done on determining the quasi-static fields produced by various subsurface sources when the measurement distance is comparable to the earth skin depth δ. (Most of the results are summarized in Kraichman's[1] book). Recently, by utilizing finitely conducting earth image theory techniques, we have derived[2] approximate expressions for the general quasi-static range electromagnetic fields (in the air and in the earth) produced by various subsurface antennas.

* Now with Electrical Engineering Department, Western New England College, Springfield, Mass.

The above mentioned investigations showed that when the source is buried not too far from the surface of the conducting half-space, the resultant field is significantly different from that which one obtains in an infinite conducting medium. It follows that if the conductor is itself two-layered, then the lower interface must likewise affect the field, especially if the upper layer (in which the source is located) is not too deep. For example, if the source is situated in a shallow sea, it is clear that the theory for a uniform conducting half-space would not accurately describe the field at the sea bottom, not at the surface of the sea if the sea depth (ℓ_1) were less than about one skin depth deep.

Weaver[3] has numerically evaluated the exact Sommerfeld integrals and has obtained numerical results for the quasi-static fields produced by a horizontal electric dipole (HED) located in the upper layer of a two-layer conducting half-space. It is the purpose of this paper to present approximate expressions for both the general quasi-static and quasi-near range electromagnetic fields produced by a HED located in the upper layer of a two-layer conducting earth. The fields produced by other dipole sources have also been derived.[4,5] Finitely conducting earth image theory techniques and the quasi-near approximation to the basic Sommerfeld integrals have been utilized in the derivation of the field component expressions.

For mathematical convenience, we will let the conductivity of the bottom layer equal to zero. Thus the problem reduces to calculating the fields in a conducting slab. The fact that the bottom layer conductivity equals zero is not that restrictive because for many practical cases, the conductivity of the upper layer (σ_1) is much greater than the conductivity of the bottom layer (σ_2)- in particular for the sea-sea bed case. This assumption limits the results to measurement distances of about $\delta_2/2$ where δ_2 is the skin depth in the bottom layer. For example at a frequency of 1 Hz, if σ_1 = 4 mhos/m (sea) and $\sigma_2 = 10^{-4}$ mhos/m (sea bed) then $\delta_1 \sim 250$m and $\delta_2 \sim 5$ km. Thus for this particular example, the results should be valid out to measurement distance of approximately 2.5 km. Furthermore, if the conductivity of the sea bed is 10^{-4} mhos/m ($\delta_2 \sim 50$km), the results should be valid out to a measurement distance of approximately 25 km.

For the purpose of this paper, the HED is oriented in the x direction and located at depth h (h positive) with respect to a cylindrical coordinate system (ρ, ϕ, z) It is assumed to be of infinitesimal length ℓ and carries a constant current I. Free space occupies the regions outside the conducting slab of thickness ℓ_1, (see Fig. 1). Displacement currents are neglected in both the slab and the air. The magnetic permeability of the conducting slab is assumed to equal μ_0, the permeability of free space. Meter-kilogram-second (MKS) units are employed and a suppressed time factor of exp $(i\omega t)$ is assumed.

MODIFIED FINITELY CONDUCTING EARTH IMAGE THEORY TECHNIQUES

For the semi-infinite conducting medium case where both the source and receiving antenna are located above the earth's surface, the quasi-static range ($\gamma_0 \sim 0$) integrals to be evaluated are of the type (which cannot be evaluated analytically throughout the quasi-static range):

$$I_1 \sim \int_0^\infty \left(\frac{u-\lambda}{u+\lambda}\right) e^{-\lambda|z+h|} J_0(\lambda\rho)\, d\lambda \qquad (1)$$

where $\gamma_0 = i\omega(\mu_0\varepsilon_0)^{1/2}$ (air),

$\gamma \sim (i\omega\mu_0\sigma)^{1/2}$ (earth),

$u = (\lambda^2 + \gamma^2)^{1/2}$, and

$J_0(\lambda\rho)$ is the Bessel function of the first kind, order zero and argument $\lambda\rho$.

Physically, the essence of the finitely **conducting** earth image theory technique is to replace the finitely conducting earth by a perfectly conducting earth located at the (complex) depth $d/2$ where $d = 2/\gamma = \delta(1-i)$. Analytically, this corresponds to replacing the algebraic reflection coefficient, $(u-\lambda)/(u+\lambda)$, in the exact integral equations by $\exp(-\lambda d)$, where λ is the variable of integration. Once this is done (1) can be readily evaluated. For antennas located at or above the earth's surface, the general image theory approximation is valid throughout the quasi-static range [6].

When both the source and receiving antenna are located below the earth's surface, the quasi-static range integrals to be evaluated are of the form

$$I_2 \sim \int_0^\infty \left(\frac{u-\lambda}{u+\lambda}\right) e^{-u|z+h|} J_0(\lambda\rho)\, d\lambda \qquad (2)$$

Integrals of this type can be evaluated analytically throughout the quasi-static range. In fact, they have been. However, the resulting expressions[1] are very complicated because they involve products of modified Bessel functions of different argument. Therefore we let[2]

$$e^{-u|z+h|} \sim e^{-\gamma a|z+h|}\, e^{-\lambda b|z+h|} \qquad (3)$$

where

a = 0 and b = 1 for $R_1/\delta \ll 1$ $(R_1^2 = \rho^2 + (z+h)^2)$
a = 0.4 and b = 0.96 for R_1/δ less than approximately 1,
a = 0.96 and b = 0.4 for R_1/δ between approximately 1 and 10, and
a = 1 and b = 0 for $\rho > 3\,|z + h|$.

Substituting (3) into (2) and substituting exp $(-\lambda d)$ for $(u-\lambda)/(u+\lambda)$ results in

$$I_2 \sim e^{-\gamma a|z+h|} \int_0^\infty e^{-\lambda[d+b|z+h|]} J_0(\lambda\rho)\,d\lambda \quad (4)$$

which can be readily evaluated throughout the quasi-static range.

CONDUCTING SLAB FIELD COMPONENT EXPRESSIONS FOR THE GENERAL QUASI-STATIC RANGE

Since we have already derived the general quasi-static range subsurface to subsurface field component expressions for the semi-infinite medium case[2] we can employ these results and the method of images to derive approximate expressions for the quasi-static fields within a conducting slab. For this case, the multiple image pattern is an infinite array that supplements the original dipole in exactly the same manner as the images in two plane mirrors of a physical object between them appear to be an apparently infinite array of that object.

The resulting expression for the HED E_ϕ component is

$$E_\phi \sim \frac{I\ell \sin\phi}{4\pi\sigma} \sum_{n=0}^{\infty} \epsilon_n \left\{ \frac{e^{-\gamma R_{0N}}}{R_{0N}^3}\left[1 + \gamma R_{0N} + \gamma^2 R_{0N}^2\right] - \frac{e^{-\gamma R_{1N}}}{R_{1N}^3}\left[1+\gamma R_{1N} + \gamma^2 R_{1N}^2\right] + \frac{2e^{-\gamma a X_2}}{A_{1N}^3}\left[1 + \frac{2A_{1N}^2}{d^2}\left(1 - \frac{A_{1N}}{A_{2N}}\right)\right]\right\} \quad (5)$$

where $X_1 = 2nd_1 + z - h$, $X_2 = 2nd_1 + z + h$, $R_{0N}^2 = \rho^2 + X_1^2$

$R_{1N}^2 = \rho^2 + X_2^2$, $A_{1N}^2 = \rho^2 + b^2X_2^2$, $A_{2N}^2 = \rho^2 + (d + bX_2)^2$

and $\epsilon_0 = 1$, $\epsilon_n = 2$, $n = 1, 2, 3, \ldots$

Similar expressions have also been derived for the other HED (and other dipole) field components.[4]

CONDUCTING SLAB FIELD COMPONENT EXPRESSIONS FOR THE QUASI-NEAR RANGE

The quasi-near range is defined as the asymptotic part of the quasi-static range (i.e., where the measurement distance is much greater than a skin depth in the conducting medium and much greater than the depth of burial of the transmitting and receiving antennas). Generally, ρ must be greater than 3δ and greater than $3(z + h)$. For the conducting slab, ρ should also be $> 3\ell_1$. However, as we shall later see, the requirement that $\rho \geq 2\delta$ and $\rho \geq 2\ell_1$, may be sufficient.

The quasi-near range approximation is setting the function $u = (\lambda^2+\gamma^2)^{\frac{1}{2}}$ in the exact integral expressions equal to γ, which is the propagation constant in the conduction medium.

If the slab is not too thin, the quasi-near range HED field component expressions valid within the conducting slab are[4]:

$$E_\rho \sim \frac{I\ell Q^2 \cos\phi}{2\pi\sigma\rho^3} A(h,z) \tag{6}$$

$$E_\phi \sim \frac{I\ell Q^2 \sin\phi}{2\pi\sigma\rho^3} 2A(h,z) \tag{7}$$

$$E_z \sim 0 \tag{8}$$

$$H_\rho \sim \frac{I\ell Q \sin\phi}{\pi\gamma\rho^3} B(h,z) \ , \tag{9}$$

$$H_\phi \sim -\frac{I\ell Q \cos\phi}{2\pi\gamma\rho^3} B(h,z) \ , \tag{10}$$

and

$$H_z \sim \frac{3I\ell Q^2 \sin\phi}{2\pi\gamma^2\rho^4} A(h,z) \tag{11}$$

Note that these equations are each broken up into three parts. The first part is the quasi-near range field component expression valid at the surface of a semi-infinite conducting half-space[2]. The second part is the familiar plane wave correction factor employed to account for the presence of stratification in the earth[7]. For $\sigma_1 >> \sigma_2$,

$$Q \sim \coth \gamma\ell_1 \ . \tag{12}$$

The third part accounts for the depth of burial of the transmitting and receiving antennas. For the conductivity slab case when

$$A(z,h) \sim \frac{2\cosh\gamma h \cosh\gamma(\ell_1 - z)}{\cosh\gamma\ell_1} - \frac{\sinh\gamma(\ell_1 + h - z)}{Q^2 \sinh\gamma\ell_1} \quad (13)$$

and

$$B(z,h) \sim \frac{\cosh\gamma(\ell_1 + h - z)}{\cosh\gamma\ell_1} - \frac{2\cosh\gamma h \sinh\gamma(\ell_1 - z)}{\sinh\gamma\ell_1} \quad (14)$$

THE QUASI-NEAR FIELDS ABOVE A CONDUCTING SLAB

If the slab is not too thin, the quasi-near range field component expressions outside the conducting slab produced by a HED located within the slab are[5]:

$$E_\rho \sim \frac{I\ell Q^2 \cos\phi}{2\pi\sigma R^3} \left\{ A(h,\ell_1) - \frac{\gamma z}{Q} B(h,\ell_1) \right\}, \quad (15)$$

$$E_\phi \sim \frac{I\ell Q^2 \sin\phi}{2\pi\sigma R^3} \left\{ \left(2 - 3\frac{z^2}{R^2}\right) A(h,\ell_1) + \frac{\gamma z}{Q} B(h,\ell_1) \right\}, \quad (16)$$

$$E_z \sim \frac{I\ell \cos\phi}{2\pi\sigma R^3} (\gamma\rho) Q B(h,\ell_1), \quad (17)$$

$$H_\rho \sim \frac{I\ell Q \sin\phi}{2\pi\gamma R^3} \left(2 - 3\frac{z^2}{R^2}\right) B(h,\ell_1), \quad (18)$$

$$H_\phi \sim -\frac{I\ell Q \cos\phi}{2\pi\gamma R^3} B(h,\ell_1), \quad (19)$$

and

$$H_z \sim \frac{3I\ell\rho Q^2 \sin\phi}{2\pi\gamma^2 R^5} \left\{ \left(1 - 5\frac{z^2}{R^2}\right) A(h,\ell_1) + \frac{\gamma z}{Q} B(h,\ell_1) \right\}. \quad (20)$$

where

$$A(h,\ell_1) \sim \frac{2\cosh\gamma(\ell_1-h)}{\cosh\gamma\ell_1} - \frac{\sinh\gamma(\ell_1-h)}{Q^2\sinh\gamma\ell_1}, \tag{21}$$

and

$$B(h_1\ell_1) \sim \frac{\cosh\gamma(\ell_1-h)}{\cosh\gamma\ell_1}. \tag{22}$$

If $\ell_1 > \delta$, $Q \sim 1$. Furthermore, if ℓ_1 is greater than 2h

$$A(h,\ell_1) \sim B(h,\ell_1) \sim e^{-\gamma h}, \tag{23}$$

and (15) through (20) reduce to the single layered earth quasi-near field subsurface to air propagation equations[1,8].

DISCUSSION

It would be of interest to compare the results derived in this paper with some known results. Plotted versus ρ/δ in Figure 2 is a comparison of modified image theory, quasi-near asymptotic theory, and Weaver's numerical integration results[3] for the normalized E_ϕ component at the surface of a one skin depth thick ($\ell_1 = \delta$) conducting slab produced by HED located in the middle of the slab ($h = \delta/2$). The normalized amplitude of the E_ϕ component is given by

$$E'_\phi = \frac{4\pi\sigma\delta^3 E_\phi}{I\ell\sin\phi} \tag{24}$$

Two values of a and b are considered for the modified image theory plots. The a = 0.4 and b = 0.96 results should be valid close to the source and the a = 0.96 and b = 0.4 results should be valid at further distances. For each of these cases, only five terms of the infinite summation were needed for 1% accuracy.

As Weaver[3] has already pointed out, this particular model possesses symmetry about the plane z = h. When z = h, all components that vary as B(z,h) are equal to zero. Furthermore, if z = 0 or ℓ_1, all components that vary as A(z,h) are equal, while all components that vary as B(z,h) are equal and opposite.

From Fig. 2, we see that for the E'_ϕ component, the modified image theory a = 0.4 and b = 0.96 curve is in good agreement with the numerical integration results for $0 \leq \rho/\delta < 0.75$, while the a = 0.96 and b = 0.4 curve is in better agreement for $0.75 < \rho/\delta \leq 3.0$.

Beyond approximately two skin depths, the asymptotic theory provides the better fit to the numerical integration data.

CONCLUSION

Approximate expressions for both the general quasi-static and quasi-near fields produced by a horizontal electric dipole located within a conducting slab have been derived by employing finitely conducting earth image theory techniques and by applying the quasi-near approximation to the basic Sommerfeld integrals.

We have shown that the resulting approximations are in very good agreement with previously derived numerical integration results. In particular, it appears that the (simple form) asymptotic theory will provide results of sufficient accuracy when the measurement distance is greater than two skin depths and greater than twice the slab depth.

Although displacement currents in the conducting slab have been ignored in the analysis, they can be included simply by replacing σ by $\sigma + i\omega\epsilon$ in the field strength equations (as long as $|\gamma^2| >> |\gamma_0^2|$).

The resulting expressions are particularly applicable to short range subsurface to an propagation for the shallow sea case.

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EDITORIAL NOTES BY J.R. WAIT:

P.R. Bannister gave a later impromtu presentation on the development of the complex image theory concept. He called attention to the references already cited, the two additional references listed above, and an earlier unpublished work (circa 1965) by L. Ball, E.L. Maxwell, and A.D. Watt of the Westinghouse Electric Corporation.

The interested reader may wish to consult the early work of Van der Pol* who obtained some physically meaningful yet rigorous representations of dipoles over a homogeneous half-space. He interpreted the secondary field as a volume integration over image sources that, in some cases, reduce to an elongated ellipsoid.

*B. Van der Pol, "Theory of the reflection of light from a point source by a finitely conducting flat mirror, with an application to radio telegraphy", *Physica,* Vol. 2, pp. 843-853, 1935.

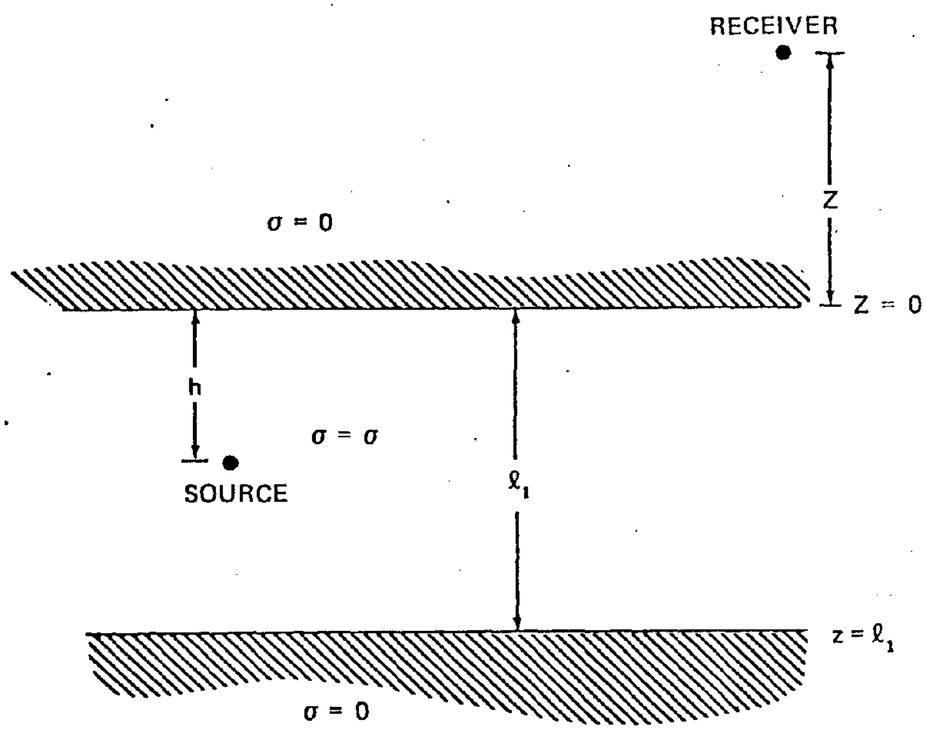

Figure 1. Conducting Slab Geometry

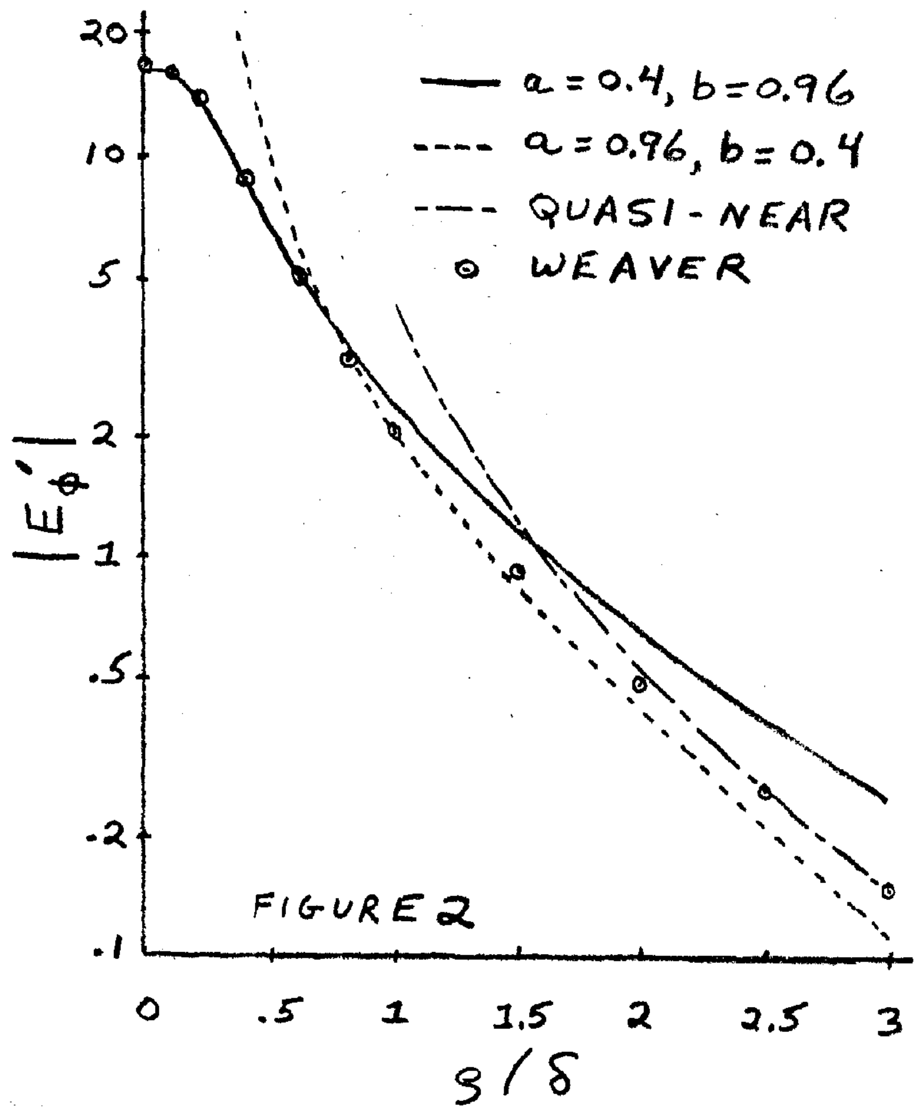

FIGURE 2

Theory of Transmission of Electromagnetic Waves Along a Metal Rod in a Conducting Medium

(Being a preliminary theory of cable-less telemetry for drilling operations)

JAMES R. WAIT and DAVID A. HILL

Institute for Telecommunication Sciences
Office of Telecommunications
U.S. Department of Commerce
Boulder, Colorado 80302

Abstract-We present an analysis of the electromagnetic fields excited by a toroidal coil that is coaxial with a circular metal rod located in a conducting medium. The integral representation for the induced axial current on the rod is cast in a form that is suitable for numerical evaluation. Some results for the electric current distribution on the infinitely long configuration are given for a fixed magnetic current source. Good agreement is obtained with an asymptotic estimate for this quantity. The effects, of truncating the bottom end of the rod and terminating the top in an ideal ground plane, are treated by an approximate method. Numerical results for this configuration are also given.

INTRODUCTION

The possibility exists that electromagnetic signals can be transmitted along a drill stem or metal rod in a conducting host rock. The idea is that the lower portion of the vertical rod is to be excited inductively in such a fashion that axial currents will propagate to the surface where they may be detected. Because the rod is in contact with the surrounding conducting medium, we expect that the attenuation rate will be fairly high even in the best of circumstances. Nevertheless, it is important to establish the theoretical limits of this type of transmission as a mechanism for transmitting information. This is the purpose of the present study.

An initial examination of the problem suggested that the metal rod be excited by a coaxial solenoid that could be fed with alternating current. This configuration is sometimes employed in non-destructive testing of cables. However, some consideration indicated that this scheme would not excite axial currents on the rod of any substantial magnitude. A more promising configuration that suggests itself is a toroidal coil that encircles the rod. Electromagnetically, such a coil, when energized by an A.C. source, is equivalent to a circumferential magnetic current. If the inner radius of this toroid were allowed to coincide with the outer surface of the rod, we would, in effect, be exciting the rod by a voltage generator. Of course, this would be most effective in driving currents along the rod. It is not necessary, however, to require that the toroid be tightly wound about the rod. Thus, in our analysis, we will consider the toroid to have a mean radius somewhat greater than the radius of the rod.

Another important practical consideration is that the rod will be of finite length. At the bottom end, the axially induced current will vanish so this will be a boundary condition that must be imposed. At the surface, the rod could be terminated in various ways. However, we feel that the optimum situation would be a simulated metal ground plane. Thus, for this study, such is assumed.

FORMULATION FOR INFINITELY LONG ROD

We begin first with the model of a circular metal rod of radius a located in an unbounded medium of conductivity σ, permittivity ε and permeability μ. A cylindrical coordinate system (ρ,ϕ,z) is then chosen such that the surface of the rod is $\rho = a$. The rod is to be excited by a uniform ring magnetic current K located at $z = z_o$, that has a radius b. To allow for subsequent generality, the finite conductivity of the metal in the rod can be characterized by an effective series impedance Z_s in ohms per meter.

In view of the symmetry of the problem, the excited magnetic field has only an $H_\phi(\rho,z)$ component that is independent of ϕ. The quantity of prime interest is the total axial current $I(z)$ on the rod that, by definition, is $2\pi a H_\phi(a,z)$.

A formal integral representation for $I(z)$ follows from an earlier analysis for the excitation of surface waves on a cylindrical structure in free space [1]. For a time factor $\exp(i\omega t)$, the corresponding result for the present configuration is

$$I(z) = -(\sigma + i\varepsilon\omega) bK \int_{-\infty}^{+\infty} \frac{H_1^{(2)}(vb)\exp[-i\lambda(z-z_o)]d\lambda}{vH_o^{(2)}(va) - (\sigma+i\varepsilon\omega)2\pi a Z_s H_1^{(2)}(va)} \tag{1}$$

where $v = (k^2 - \lambda^2)^{\frac{1}{2}} = -i(\lambda^2 - k^2)^{\frac{1}{2}}$ and $ik = [i\mu\omega(\sigma + i\varepsilon\omega)]^{\frac{1}{2}}$. $H_o^{(2)}$ and $H_1^{(2)}$ are Hankel functions of the second kind, of order zero and one, respectively. The integration here is along the entire real axis of the complex λ plane. Branch points occur in the second and fourth quadrants at $\lambda = \pm k$. We should also note that the integral has poles at $\lambda = \pm\lambda_s$ where λ_s is the solution of

$$vH_o^{(2)}(va) - (\sigma+i\varepsilon\omega)2\pi a Z_s H_1^{(2)}(va) = 0 \tag{2}$$

in the fourth quadrant of the λ plane.

It is worth pointing out here that the magnetic current ring could be a band with finite vertical extent as actually depicted in Fig. 1. The resultant electric current $I(z)$ for this case is then a superposition of elementary rings of thickness dz_o over the range from say $z_o = -s$ to $z_o = +s$. The up shot is that to account for the finite vertical extent of the source, we would insert a factor $(\sin\lambda s)/(\lambda s)$ in the integrand of (1) when K is still regarded as the total ring magnetic current. In what follows, we will set $s = 0$ but we let $z = z_o$ denote the vertical location of the toroid or ring source. As a matter of interest, the finite radial extent of the toroidal source coil could be handled in the same fashion.

In order to evaluate the integral for $z - z_o > 0$, it is convenient to deform the contour about the singularities in the lower half of the λ plane. This immediately suggests that we write (1) in the form

$$I(z) = -(\sigma+i\varepsilon\omega)bK \int_k^{k-i\infty} \left\{ \frac{H_1^{(2)}(vb)}{vH_o^{(2)}(va) - (\sigma+i\varepsilon\omega)2\pi a Z_s H_1^{(2)}(va)} - \frac{H_1^{(2)}(vbe^{-i\pi})}{ve^{-i\pi}H_o^{(2)}(vae^{-i\pi}) - (\sigma+i\varepsilon\omega)2\pi a Z_s H_1^{(2)}(vae^{-i\pi})} \right\} e^{-i\lambda(z-z_o)} d\lambda \quad (3)$$

$$+ I_s(z)$$

where the integral from k to $k-i\infty$ is the contribution from the two sides of the branch line, $I_s(z)$ in the residue at the pole $\lambda = \lambda_s$ that could be computed from

$$I_s(z) = 2\pi i(\sigma+i\varepsilon\omega)bK \frac{H_1^{(2)}(v_s b)\exp(-i\lambda_s|z-z_o|)}{\left\{\frac{\partial}{\partial\lambda}\left[vH_o^{(2)}(va) - (\sigma+i\varepsilon\omega)2\pi a Z_s H_1^{(2)}(va)\right]\right\}_{\lambda=\lambda_s}} \quad (4)$$

where $v_s = (k^2-\lambda_s^2)^{\frac{1}{2}} = -ik(\lambda_s^2-k^2)^{\frac{1}{2}}$.

The rotation formulae $H_o^{(2)}(xe^{-i\pi}) = -H_o^{(1)}(x)$ and $H_1^{(2)}(xe^{-i\pi}) = H_1^{(1)}(x)$ are useful to simplify the branch line integral in (3). Also, it is useful to change the variable to the dimensionless y via $\lambda = k(1-iy)$. Then, without difficulty, (3) becomes

$$I(z) = i(\sigma+i\varepsilon\omega)bK\exp(-ik|z-z_o|)\int_o^{\infty e^{-i \arg k}} \left\{ \frac{H_1^{(2)}(\delta kb)}{H_o^{(2)}(\delta ka) - gH_1^{(2)}(\delta ka)} - \frac{H_1^{(1)}(\delta kb)}{H_o^{(1)}(\delta ka) - gH_1^{(1)}(\delta ka)} \right\} \frac{1}{\delta} \exp(-ky|z-z_o|)dy + I_s(z) \tag{5}$$

where $\delta = -[y(y+2i)]^{\frac{1}{2}}$ and $g = (\sigma+i\varepsilon\omega)2\pi aZ_s$.

While (5) appears rather complicated, it is in a suitable form for numerical computation due to the rapidly convergent exponential factor in the integrand. For the present application, the drill stem or rod is effectively a perfect conductor so g can be set equal to zero. Also, in this case, the surface wave is not excited. Thus, for this case, we write

$$I(z) = ib(\sigma+i\varepsilon\omega)Ke^{-ikz}\, S \tag{6}$$

where

$$S = \int_o^{\infty e^{-i \arg k}} \frac{1}{\delta} \left[\frac{H_1^{(2)}(\delta kb)}{H_o^{(2)}(k\delta a)} - \frac{H_1^{(1)}(\delta kb)}{H_o^{(1)}(k\delta a)} \right] e^{-iyz}\, dy \tag{7}$$

where we have set $z_o = 0$. Since the integrand of S is singular at $y = 0$, it is convenient to break the integration range into two parts; in the first the range is from 0 to d and, in the second, it is from d to $\infty \exp(-i \arg k)$. Here d is chosen to be sufficiently small that the small argument approximations are valid for the Hankel functions. Then we find that

$$S \simeq S_d + \int_d^{\infty e^{-i \arg k}} \frac{1}{\delta} [\ldots\ldots] e^{-iyz}\, dy \tag{8}$$

where

$$S_d = \frac{2\pi}{kb} \int_0^d \frac{dy}{y\ell n^2(yk^2a^2)} = -\frac{2\pi}{kb\ \ell n(dk^2a^2)} \tag{9}$$

It is found that for small d, S evaluated in this manner is independent of d.

Integrals of the type given by (7) were evaluated asymptotically by Kunz [2]. Using his method, applied to the present configuration, we obtain, for $z \to +\infty$,

$$I(z) \simeq -\frac{K\ 2\pi}{\eta} \frac{e^{-ikz}}{\ell n\left[-\frac{2iz}{\Gamma^2 ka^2}\right]} \tag{10}$$

where $\eta = [i\mu\omega/(\sigma+i\varepsilon\omega)]^{\frac{1}{2}}$ and $\Gamma = 1.781.....$ For lossless media this reduces to Kunz's result.

RESULTS FOR THE INFINITELY LONG ROD

We now present some numerical results for the infinitely long rod shown in Fig. 1. Since $I(z) = I(-z)$ we only need consider the case $z > 0$. The integration program developed for dealing with (8) uses a variable increment size to sample the variation of the integrand over y. Some results for the magnitude of $I(z)$ are shown in Fig. 2 for a frequency of 5 kHz where the distance z varies from 10 to 1000 m. The radius a of the rod is 2 cm and the magnetic current K is fixed at a/b volts. As indicated, the ratio b/a is either 1 or 2. For this example, the rock conductivity σ is 10^{-2} mhos/m and the permittivity, relative to free space, $\varepsilon/\varepsilon_o$ is 10. But for comparison purposes, the corresponding results for a free space environment are also shown. Both the numerical integration data and the corresponding asymptotic solution for these two cases are indicated in Fig. 2.

It is apparent that the excited current or the rod near the ring source of fixed K is much greater for the conductive environment than that for free space. However, beyond about $z = 500$ m the rate of decay of the current on the rod is much greater in the case of the conductive environment. We also note that, in Fig. 2, the magnitude of the excited current for $b/a = 2$ is approximately half that for $b/a = 1$. In fact, we can say, in general, that the excited current is inversely proportional to b provided that $z \gg b$. This dependence is explicitly shown in (10).*

In order to show a meaningful frequency dependence, the current magnitude of the magnetic current K itself is allowed to be proportional to frequency. This result follows from the fact that K of a solenoid is proportional to the product of the electric current in the winding and the frequency. Thus, in Fig. 3, we plot the magnitude of $I(z)$ as a function of z for $\sigma = 10^{-2}$ mho/m for frequencies of 1, 2, 5, and 10 kHz and for $K = a/b$ volt at 5 kHz but otherwise proportional to frequency (i.e. $K = (a/b) \times f(\text{kHz}) \div 5$). For comparison, the single curve for $\sigma = 10^{-3}$ mho/m and $f = 5$kHz is also shown in Fig. 3. The frequency dependence for this case would be very similar.

Not surprisingly, the results in Fig. 3 show that the currents excited near the source toroid are greater for the higher frequencies but the corresponding attenuation is much greater. This produces a "cross-over" effect wherein, at a distance z beyond about 100 m, the current magnitude is always much reduced with increases of frequency.

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* *The method of normalizing the source calls for some explanation. We adjust K to be a/b volts so that the electric current in the solenoid, the winding radius and the number of turns remain constant.*

EXTENSION TO TRUNCATED AND TERMINATED ROD

We now consider an approximate extension of the foregoing analysis to allow for end effects. Specifically, we consider the configuration shown in Fig. 4. Here we represent the termination of the metal rod at the surface as a perfectly conducting ground plane. On the other hand, the rod is truncated at a distance ℓ below the source toroid. An approximate solution for the total current $I_t(z)$, due to end effects, is then obtained by imposing the conditions of perfect reflection at the ground plane at $z = d$ and the vanishing of the current at $z = -\ell$. This leads to a superposition of "current waves" as follows:

$$I_t(z) = I(z) - I(z+2\ell) + I(2d-z) - I(2(d+\ell)-z) + \ldots\ldots$$

where $d < z < -\ell$. In particular, the current at the top of the rod can be approximated by

$$I_t(d) \simeq 2[I(d) - I(d+2\ell)]$$

where $I(d)$ and $I(d+2\ell)$ can be obtained from (8) using appropriate numerical integration.

Some results for the magnitude of the total current I_t at $z = d$ are shown in Figs. 5 and 6 for $\sigma = 10^{-2}$ and 10^{-3} mho/m, respectively. In both cases the frequency is fixed at 5 kHz, $a = 2$ cm, $b/a = 2$, $\varepsilon/\varepsilon_o = 10$, and $K = 1/2$ volt. As indicated in Figs. 5 and 6, the magnitude of the resultant current at the top of the rod generally decreases as ℓ is reduced except there is slight enhancement for $\ell = 100$ m. This corresponds approximately to the condition that ℓ is of the order of a quarter of an effective wavelength within the medium. The effect could be more noticeable for higher frequencies and when the rock conductivity is lower.

CONCLUDING REMARKS

Admittedly, the model we have employed is highly idealized. However, the assumption in the calculations that the rod is perfectly conducting can be easily relaxed by using an appropriate value for the series impedance Z_s. However, it appears that, to within graphical accuracy, the results would not be changed. A more important limitation would be the imperfect contact between sections of the rod if it were not continuous throughout its length. In such cases, the axial current flow could be inhibited with an effective increase of the total signal loss.

For the present model, the resultant attenuation will be at least as great as plane waves in the conducting medium that surrounds the rod. The only way this intrinsic loss could be circumvented would be to provide a layer of insulation on the rod. The formulation we have given could actually apply to this situation if, again, the effective (spatially dispersive) value [3] of Z_s be used. In such cases, the pole contribution to the integral representation would play an important role.

As indicated in the formulation, we can easily allow for the finite vertical and radial extent of the source toroid. However, for the numerical examples here, this would not have any noticeable effect on the current distributions. Also, the toroid itself could be ferrite loaded in order to improve the flux linkage for a given electrical current in the winding wires.

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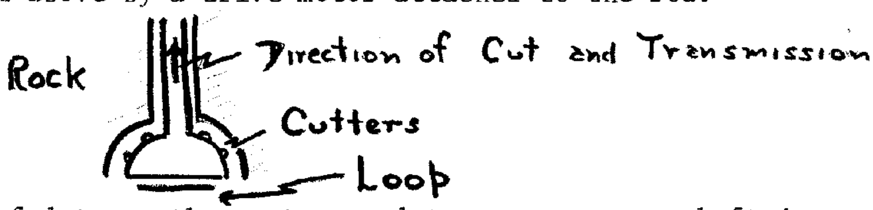
COMMENT BY B.A. AUSTIN

These results are particularly significant in that they shed real light on a problem we have in South African gold mines. In raise boring, a large dia. (2-3m) cutter is attached to a long (300m at times) drill rod and is driven from above by a drive motor attached to the rod.



Telemetry of data on the cutters and torque, etc. on shaft is required. We have mounted a loop below the cutting head (in, apparently the worst configuration) and have transmitted data at 1 MHz over a rod length of $\sim$ 300m. The problem has been the lack of reliability from site to site, where sometimes signal levels are unusable over only 100m lengths of rod. The practical problem of attaching a toroidal loop around the rod is severe in that it cannot easily be protected during the cutting operation.

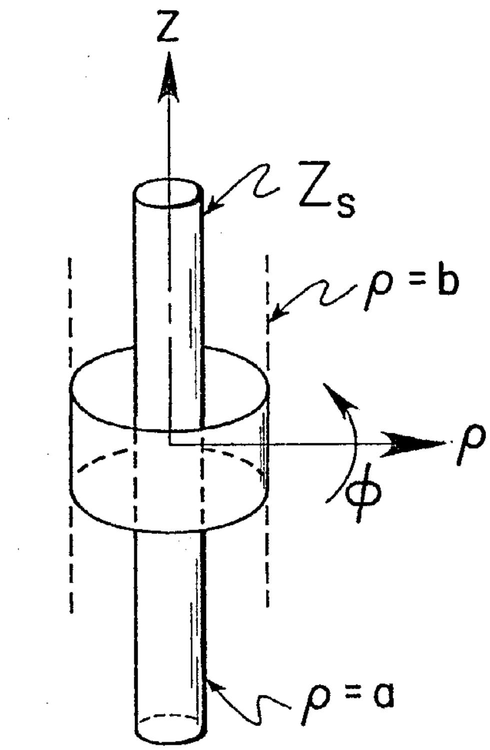


Fig. 1. The basic geometry of an infinitely long metal rod that is excited by a ring or band of magnetic current.

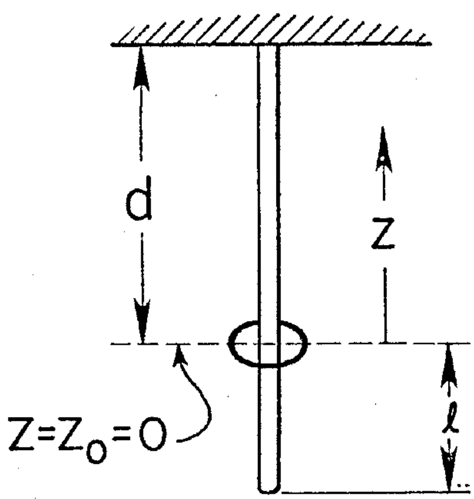


Fig. 4. The model employed to consider the finite depth extent of the rod and the termination of the rod at the surface.

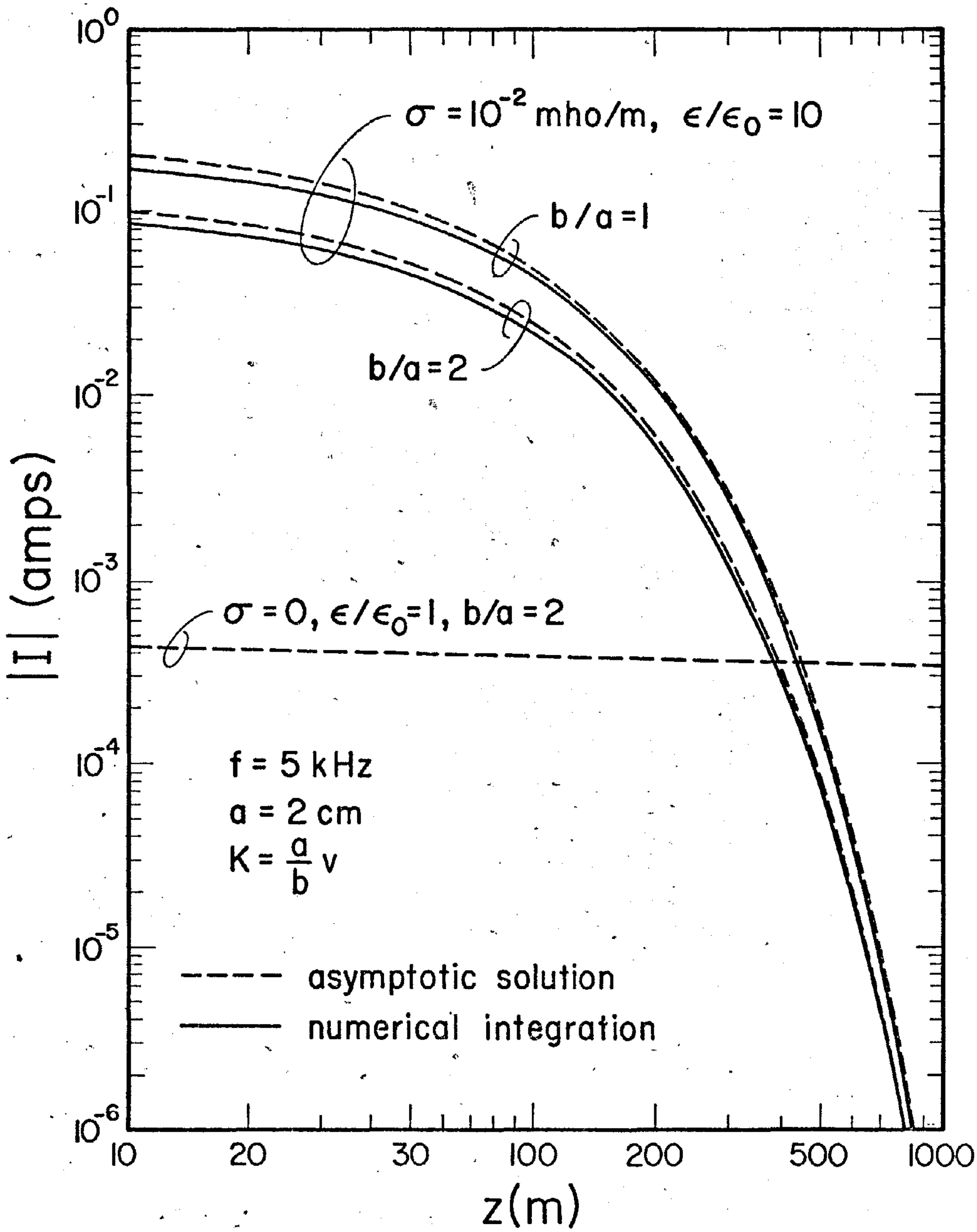


Fig. 2. The excited current on the infinitely long rod excited by the magnetic ring source for a frequency of 5 kHz.

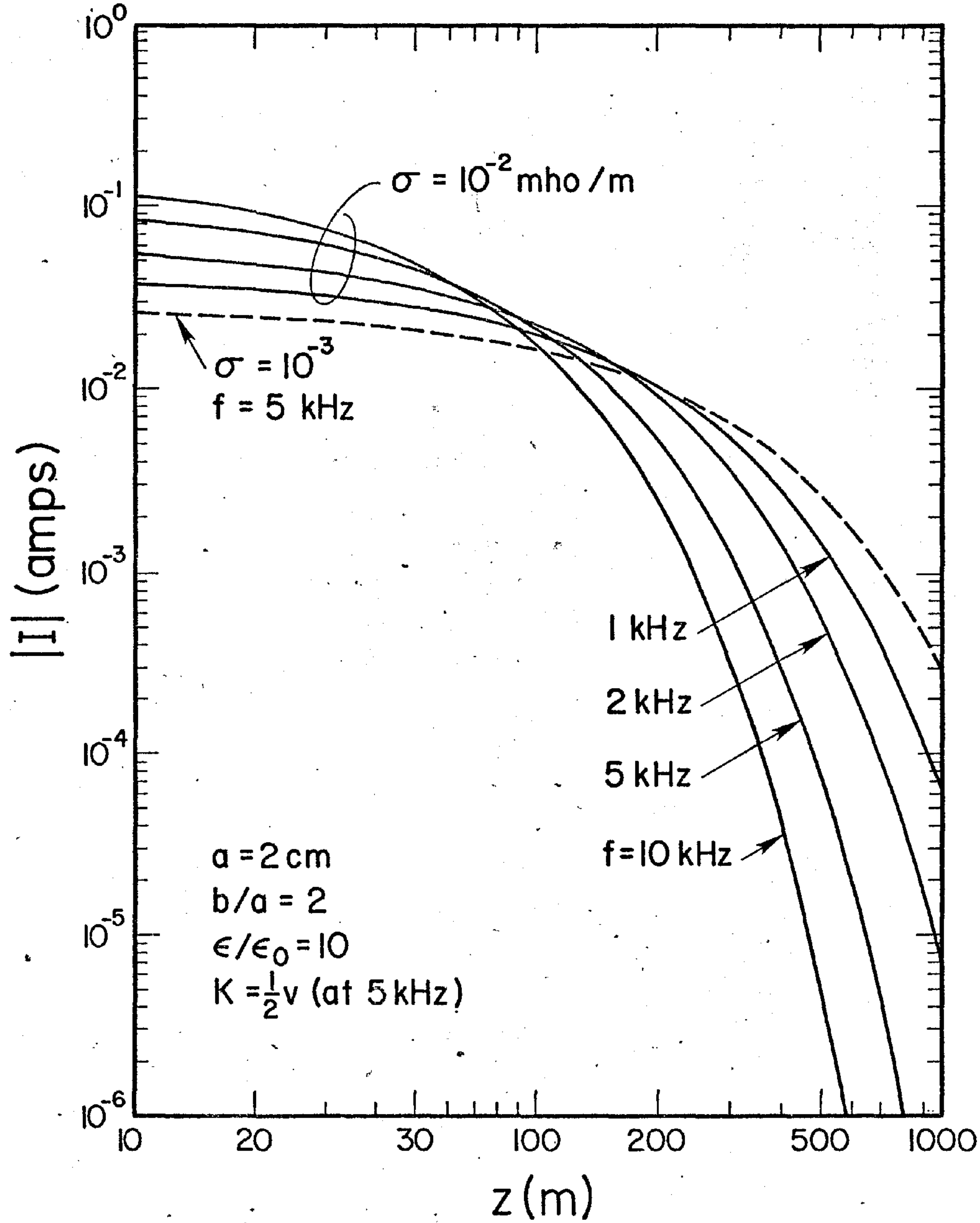


Fig. 3. The excited current on the infinitely long rod excited by the magnetic ring source for various frequencies normalized for a fixed electric current in the winding wires of the source toroid.

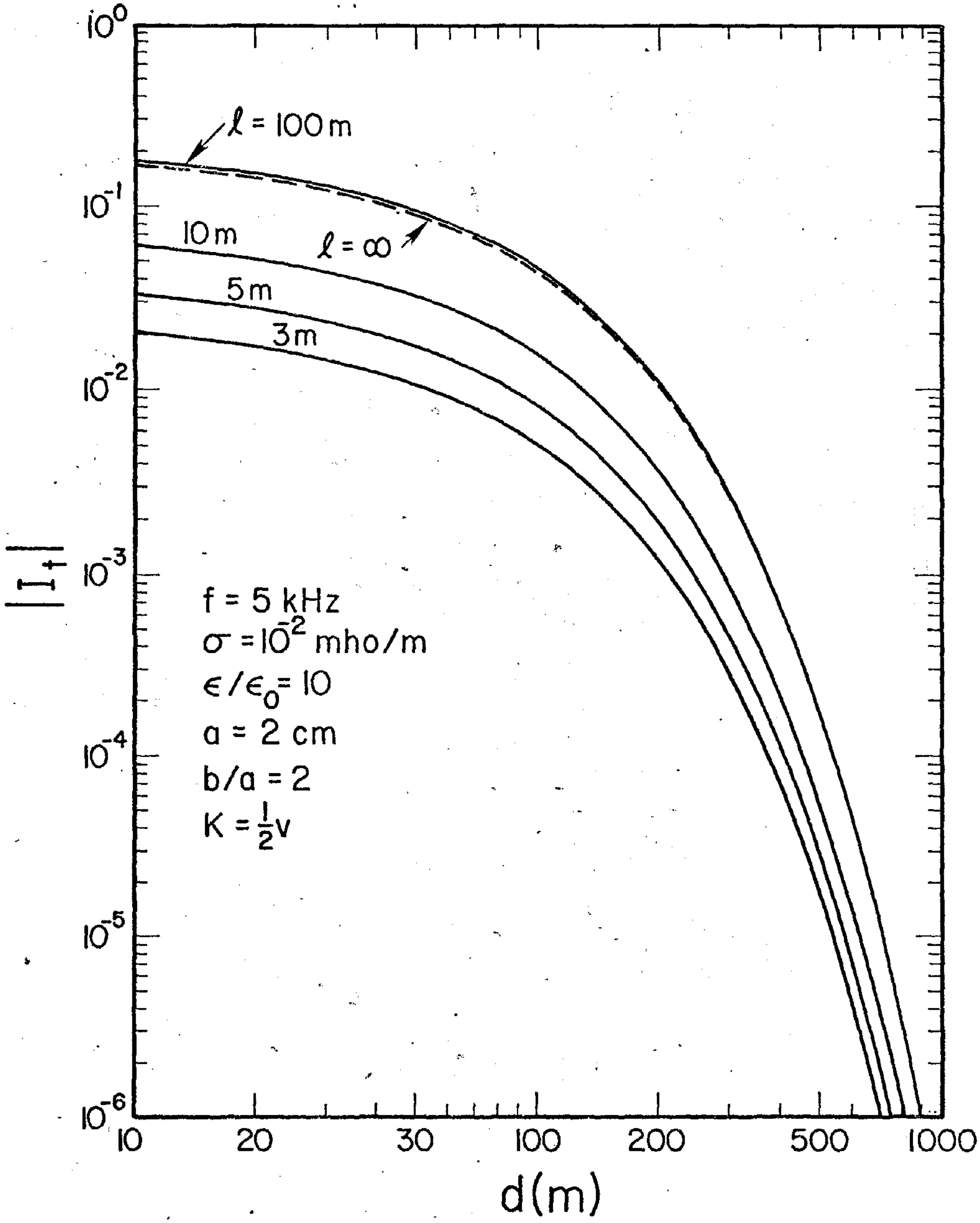


Fig. 5. The current at the top end of the rod as a function of the depth d of the source toroid for various values of the extension ℓ of the rod below the toroid.

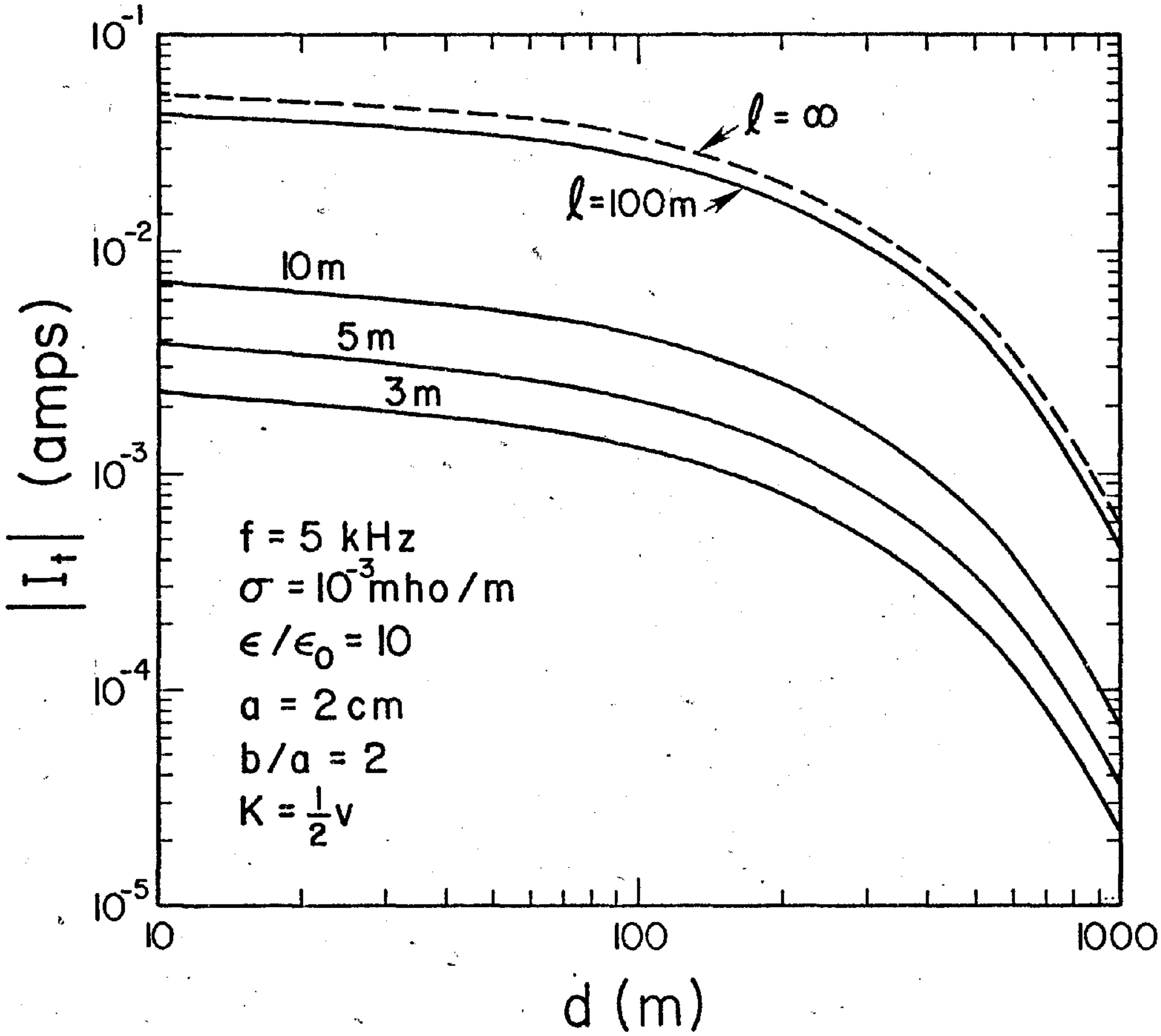


Fig. 6. The same as Fig. 5 but for a lower rock conductivity.

Guide to the Use of Laboratory Electrical Measurements for Interpretation

Gary R. Olhoeft
U. S. Geological Survey
Denver, Colorado

The electrical properties of rocks and minerals range over more than 24 orders of magnitude in resistivity and more than 17 orders of magnitude in frequency. The range of resistivity (or of dielectric permittivity) that is possible at any given frequency decreases with increasing frequency. This is primarily because the higher frequencies have correspondingly shorter wavelengths and shorter periods for the charged particles to interact with matter through various mechanisms. As mechanism effects are additive, fewer possible mechanisms for interaction mean less range of possible electrical properties. Over the entire frequency range noted above, the most important single interaction in the electrical properties of earth materials is due to the content and state of the water in the material. Depending upon the specific frequency discussed, there are up to 10 possible mechanisms related to the presence of water in a material. As a corollary of the importance of water is the importance of places within the material for the water to reside either physically or chemically. Of secondary importance is the chemical composition of the material. Third in importance is the temperature. Fourth is a wide variety of other parameters, including pressure, amplitude of stimulus, time, and others. Due to the extreme variability of the electrical properties of earth materials (as an example, basalt may change electrical resistivity by more than 12 orders of magnitude with changes in water content or temperature or frequency), the importance of parallel laboratory investigations with field studies cannot be over emphasized. Laboratory studies can measure electrical properties under simulated environmental conditions while perturbing those conditions to identify the mechanisms and find the parameter that is most important in determining the electrical properties in a given situation. The resulting laboratory based model may then be used to constrain the possible range of electrical properties for interpretation of field data to a more tractable set of limits.

Electrical Properties of Eastern Bituminous Coal as a Function of Frequency, Polarization and Direction of the Electromagnetic Wave, and Temperature of the Sample

Constantine A. Balanis, John L. Jeffrey, and Yang K. Yoon

Department of Electrical Engineering
West Virginia University
Morgantown, West Virginia 26506

ABSTRACT

The development of electromagnetic systems to detect and monitor an in-situ underground coal gasification process requires a knowledge of the electrical properties of coal. Using a two-path interferometer at microwave frequencies (≃ 9 GHz), the samples of solid eastern bituminous coal were tested as a function of temperature, electric field polarization, and direction of the coal samples. There were slight decreases in the electrical properties of coal as a function of temperature from ambient up to 700°F. However, there were significant variations in the conductivity as a function of polarization, with the vertical polarization possessing lower conductivity. Only slight changes in conductivity were detected as a function of direction, with the face cleats exhibiting somewhat higher losses. The dielectric constant remained essentially constant as a function of polarization and direction of the electromagnetic wave.

Measurements of electrical properties of coal were also made in the 0.5-100 MHz region using capacitance techniques. In general, there was a decrease in the values of dielectric constant and an increase in conductivity as the frequency increased.

I. INTRODUCTION

The development of electromagnetic systems to detect, monitor, and communicate through an underground coal environment necessitates a knowledge of the electrical properties of coal (dielectric constant ε_r and conductivity σ). Associated with coal, as with all kinds of rock, are natural fractures which usually occur systematically in conjugate pairs almost perpendicular to each other. These fractures are called cleats in coal beds and joints in rocks. In contrast to the joints, the cleats are usually more numerous, more well developed, and almost parallel to each other.

Associated with a Pittsburgh coal seam, there are three distinct and almost mutually perpendicular physical planes, as shown in Figure 1. The most well developed and smoother vertical plane is called the face cleats while the other is the butt cleats. The horizontal plane is referred to as the bedding plane.

In this paper, two laboratory methods are discussed which have been used to determine the dielectric constant and the conductivity of solid coal samples as a function of temperature, and frequency, polarization and direction of the electromagnetic wave propagation. The first method used to determine the electrical properties of coal is accomplished with the use of a two-path interferometer system. This microwave system permits the measurements of phase and amplitude variations of the signal through a coal sample. The system utilizes a variable phase shifter and a variable attenuator to compensate for the phase and amplitude changes in the system due to the insertion of the sample. The second method uses a capacitance technique in a bridge arrangement to determine the electrical properties as a function of frequency in the 0.5-100 MHz region.

II. TWO-PATH INTERFEROMETER

The two-path interferometer, shown in a block diagram form in Figure 2 consists of a reference and a working path. The reference path is composed of a variable phase shifter, a variable attenuator and a directional coupler, whereas the working path is composed of a variable attenuator, a slotted line, two horn antennas and a directional coupler. By proper adjustment of the variable attenuator and phase shifter in the working and reference paths, respectively, a null indication is produced at the phase detector. The change of amplitude and phase, which contains the information needed to compute the constitutive parameters, are obtained from the initial and final measurements of the system. The initial measurements are defined without the presence of the dielectric medium in the system with the reference and working paths having the same amplitude and phase which will produce a null at the detector. The final measurements are defined with the dielectric medium inserted in the system. With proper adjustments of the variable attenuator and phase shifter in the proper paths, a null indication at the detector is reproduced. The dielectric constant of the sample is computed from the changes in phase whereas the conductivity is computed from the changes in the amplitude.

Several samples from each direction of the coal, shown in Figure 3, and ranging in thickness from 0.25 to 1.5 inches for the two-path interferometer and from 0.13 to 0.20 inches for the capacitance technique were measured. The samples were bored out of solid blocks of coal into cores of about 5.5 inches in diameter for the two-path interferometer measurements ($\simeq$ 9 GHz) and about 2 inches in diameter for the capacitance bridge measurements. The large diameter cores were reinforced with grout as shown in Figure 3.

III. MEASUREMENTS, COMPUTATIONS, AND RESULTS

The coal samples were tested at microwave frequencies ($\simeq$ 9 GHz) using the two-path interferometer as a function of the temperature of the sample, and direction and polarization of the electromagnetic wave. At the VHF range (0.5-100 MHz) the properties of the samples were measured as a function of frequency.

A. Two-Path Interferometer

In order to determine the constitutive parameters as a function of direction and polarization, each of the samples were placed within the horns and were rotated for each measurement to obtain the electric field orientation indicated in Figure 1, and will be referred to as vertical polarization (VP), 45° polarization (45°P), and horizontal polarization (HP). To obtain the constitutive parameters as a function of temperature, the samples were heated in an oven over a temperature range of 65° to 700°Fahrenheit (F) and the measurements were taken at every 100°F intervals.

The computed, from measurements, data [dielectric constant ε_r (dimensionless) and conductivity σ (mhos/m)] are shown listed in Tables 1 and 2. At the time the temperature measurements were made face and butt cleats samples were only available, and the results are shown in Table 1. Due to the decomposition (swelling) of the eastern bituminous coal samples at temperatures above about 650-700°F, the measurements were limited to an upper temperature of 700°F. Listed in Table 2 are the data for all three planes for wet and dry samples.

It is evident that there are slight decreases in the values of dielectric constant and conductivity as the temperature is increased from ambient to about 700°F and as the samples become dry. Within measuring accuracies, there are no noticeable variations in the dielectric constant as a function of polarization and direction of travel of the electric field. There were slight differences in conductivity as a function of direction of electromagnetic wave travel with the face cleats possessing on the average slightly larger values.

It has been established, however, that there are marked polarization effects in the conductivity of a eastern bituminous solid coal sample. Conductivities lower by a factor of 3 to 6, depending upon the temperature and moisture content of the sample, for each of the following, by referring to Figure 1, have been measured:

a. the electromagnetic wave travels perpendicular to the face cleats and its electric field is parallel to the face and butt cleats but perpendicular to the bedding plane
b. the electromagnetic wave travels perpendicular to the butt cleats and its electric field is parallel to the butt and face cleats but perpendicular to the bedding plane
c. the electromagnetic wave travels perpendicular to the bedding plane and its electric field is parallel to the bedding plane and face cleats but perpendicular to the butt cleats.

B. Capacitance Bridge

In a related investigation capacitance techniques in a bridge arrangement were used to compute the constitutive parameters as a function of frequency in the 0.5-100 MHz region. Samples of eastern bituminous coal and lignite 2 inches in diameter with maximum thickness of 0.2 inches were placed within the electrodes of a sample holder, as shown in Figure 4, and comprising one arm of a bridge arrangement. The electric field orientation within the sample is also shown in Figure 4.

The computed data, from measurements, is shown listed in Table 3. It is evident, as expected, that the dielectric constant decreases and the effective conductivity increases as the frequency increases.

For coal, the face cleats possess the lower values of dielectric constant and conductivity. Also it is clear that lignite exhibits higher values of dielectric constant and conductivity than eastern bituminous coal.

A comparison of the data from Tables 2 and 3 indicates that there are only small decreases in values of the dielectric constant of coal from 100 MHz to 9 GHz (approximately from 6 to 4) while there is a significant increase in the conductivity at a rate slightly higher than the increase in the frequency.

IV. CONCLUSIONS

Laboratory techniques have been used to determine the electrical properties [dielectric constant (dimensionless) and conductivity (mhos/m)] of eastern bituminous solid coal samples as a function of temperature and frequency, polarization, and direction of travel of the electromagnetic wave. From the results, it is concluded that

a. the dielectric constant is essentially independent of the polarization and direction of travel of the electromagnetic wave and decreases slightly from ambient to 700°F.
b. the conductivity decreases slightly over the tested temperature range.
c. the conductivity for a vertically-polarized electromagnetic wave is markedly smaller (by a factor of 3 to 6 depending upon the temperature and moisture content) than the corresponding one for horizontally-polarized waves.
d. the conductivity increases and the dielectric constant decreases as the frequency increases.

These findings conclude that, because of the polarization phenomenon of coal, the efficiency of an underground electromagnetic system can be improved significantly by orienting the transmitting and receiving elements to radiate and receive vertically-polarized waves for which coal possesses more permeable electrical characteristics.

V. ACKNOWLEDGEMENTS

This work was sponsored by the U.S. Department of Energy [formerly the Energy Research and Development Administration] under contract E-(40-1)-5081. The authors would like to thank J.W. Martin and Dr. L.D. Strickland of the Morgantown Energy Research Center for their cooperation and interest on the project.

TABLE 1

Electrical Properties of Eastern Bituminous Coal as a Function of Temperature and Polarization

Frequency ≈ 9 GHz [ε_r(dimensionless), σ(mhos/m)]

Temperature (°F)			65	200	300	400	500	600	700
Face Cleats	Vertical Polarization	ε_r	3.89	3.70	3.79	3.68	3.65	3.78	3.53
		σ	0.26	0.17	0.17	0.17	0.14	0.12	0.10
	45° Polarization	ε_r	3.84	3.72	3.93	3.60	3.79	3.72	3.58
		σ	0.43	0.39	0.40	0.36	0.33	0.24	0.29
	Horizontal Polarization	ε_r	3.74	3.68	3.92	3.58	3.67	3.89	3.71
		σ	0.73	0.69	0.77	0.74	0.73	0.66	0.61
Butt Cleats	Vertical Polarization	ε_r	3.88	3.38	3.39	3.41	3.43	3.48	3.48
		σ	0.18	0.12	0.11	0.10	0.10	0.09	0.11
	45° Polarization	ε_r	3.85	3.40	3.36	3.38	3.44	3.50	3.54
		σ	0.32	0.28	0.29	0.28	0.27	0.27	0.23
	Horizontal Polarization	ε_r	3.93	3.41	3.40	3.36	3.36	3.49	3.56
		σ	0.51	0.44	0.48	0.47	0.49	0.53	0.54

TABLE 2

Electrical Properties of Eastern Bituminous Coal as a Function of Direction and Polarization [ε_r(dimensionless), σ(mhos/m)]

f ≃ 9 GHz Ambient Temperature		FACE CLEATS			BEDDING PLANES			BUTT CLEATS		
		V P	45° P	H P	V P	45° P	H P	V P	45° P	H P
WET	ε_r	3.89	3.84	3.74	3.97	3.93	3.87	3.88	3.85	3.93
	σ	0.26	0.43	0.73	0.28	0.39	0.45	0.18	0.32	0.51
DRY	ε_r	3.71	3.72	3.68	3.81	3.79	3.75	3.38	3.40	3.41
	σ	0.17	0.39	0.69	0.13	0.21	0.36	0.12	0.28	0.44

TABLE 3

Electrical Properties of Eastern Bituminous Coal and Lignite as a Function of Frequency [ε_r(dimensionless), σ(mhos/m)]

Frequency (MHz)			0.5	1	5	10	25	50	100
Coal	Face Cleats	ε_r	16.3	10.7	5.1	4.8	4.3	4.2	4.1
		$\sigma x10^3$	0.18	0.25	0.39	0.67	0.76	0.92	1.26
	Bedding Plane	ε_r	35.2	28.2	20.2	14.2	8.3	7.0	5.9
		$\sigma x10^3$	0.32	0.69	2.75	3.57	3.71	5.20	6.90
	Butt Cleats	ε_r	41.1	34.3	15.3	10.1	8.0	6.2	5.9
		$\sigma x10^3$	0.45	0.94	2.23	2.92	2.95	3.15	3.47
Lignite	Face Cleats	ε_r	52.3	39.4	28.7	26.3	21.1	15.8	14.3
		$\sigma x10^3$	0.44	0.84	3.86	5.42	6.98	9.21	20.94
	Butt Cleats	ε_r	59.1	44.2	25.4	22.3	16.9	14.2	12.6
		$\sigma x10^3$	0.42	0.89	3.60	5.50	7.51	8.22	17.54

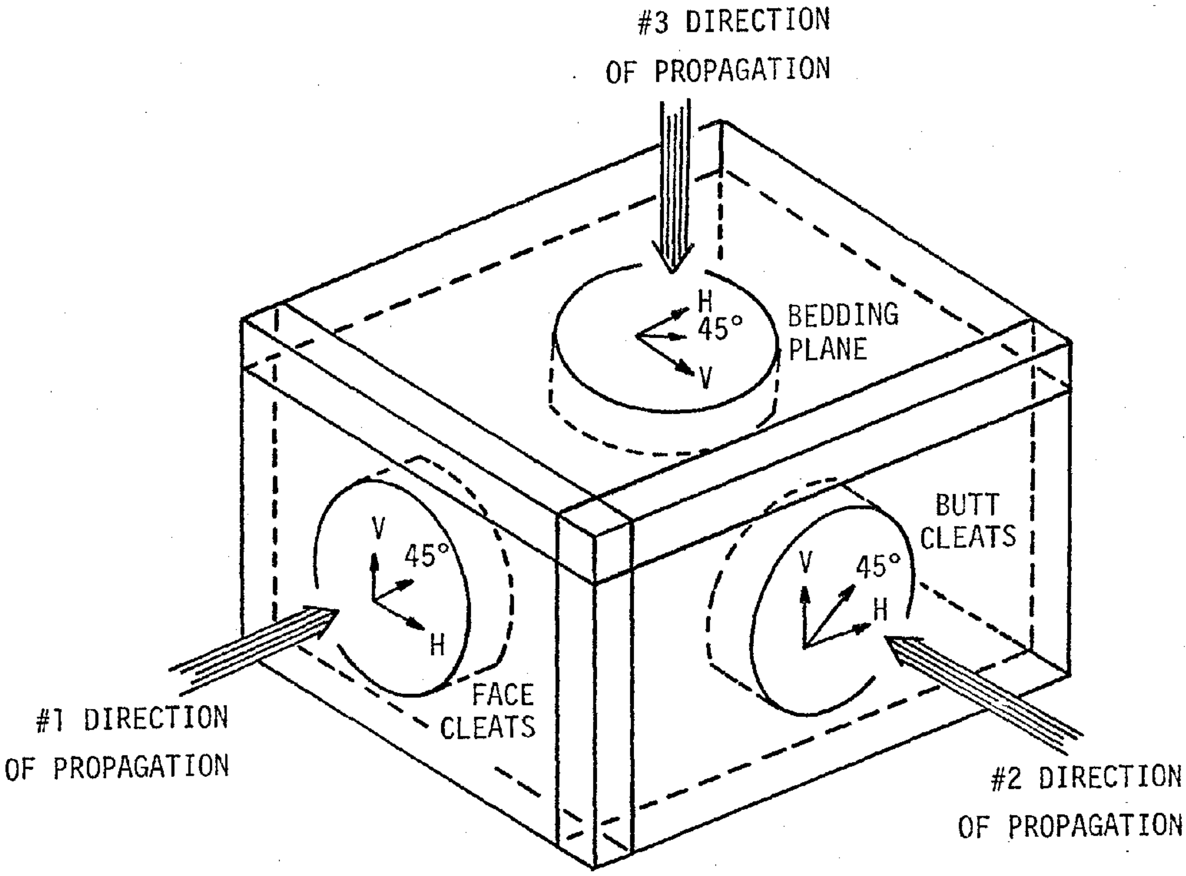


Fig. 1. Coal medium with coal sample orientations.

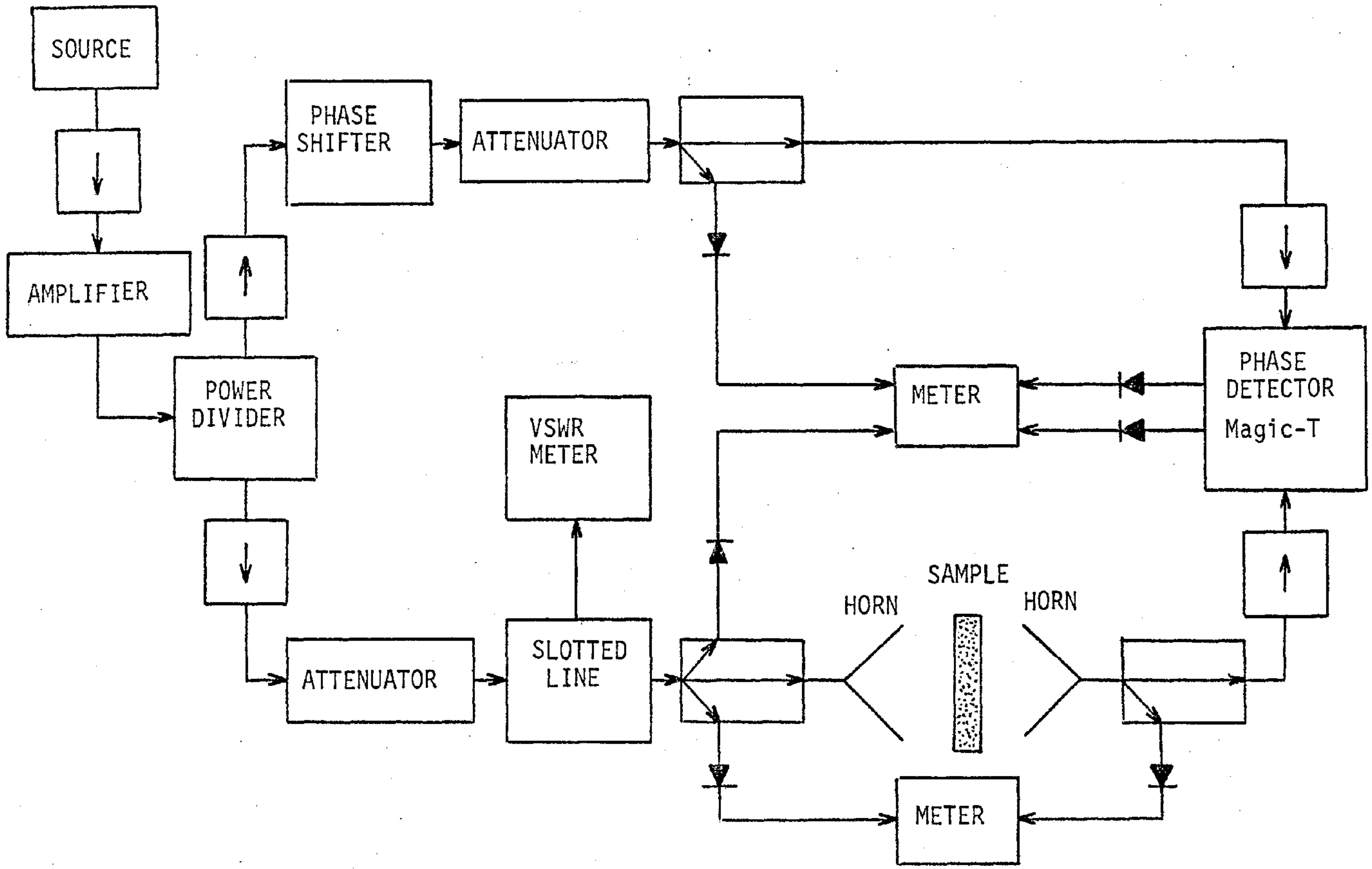


Fig. 2. Block diagram of a microwave two-path interferometer.

Fig. 3. Photograph of coal sample showing cleatings.

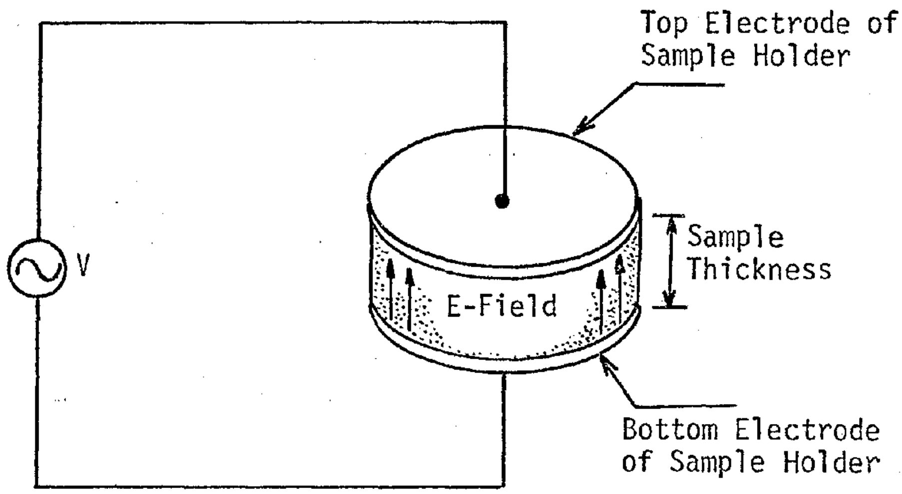


Fig. 4. Coal sample orientation and electric field polarization within sample holder.

LOCATING A BURIED MAGNETIC DIPOLE*

Thurlow W. H. Caffey
TM Development Division 9421
Sandia Laboratories, Albuquerque, New Mexico, 87115

Consider a buried magnetic dipole oriented as shown in Figure 1. It is desired to find the location of the dipole in the surface coordinate frame as well as the angles (α, β) to facilitate recovery of the instrument package or person who deployed the dipole. The frequency of the dipole current is chosen so that the dipole field near the surface of the earth is virtually the same as a static dipole in free space.

Theoretical Outline

The most convenient coordinates to choose with respect to the buried dipole is a cylindrical $(\rho, \phi, \mathscr{z})$ system with the $\mathscr{z}$-axis along the dipole axis (Figure 2). The unit vector in the $\mathscr{z}$ direction can be written immediately:

$$\hat{a}_{\mathscr{z}} = \hat{a}_x \sin\alpha \cos\beta + \hat{a}_y \sin\alpha \sin\beta + \hat{a}_z \cos\alpha \quad . \qquad (1)$$

The total H-vector measured at surface position P_i is

$$\vec{H}_i = \hat{a}_x H_{xi} + \hat{a}_y H_{yi} + \hat{a}_z H_{zi} \quad . \qquad (2)$$

and the unit vector in the ϕ-direction is given by

$$\hat{a}_{\phi i} = (\hat{a}_{\mathscr{z}} \times \vec{H}_i) / |\hat{a}_{\mathscr{z}} \times \vec{H}_i| \quad . \qquad (3)$$

The unit vector $\hat{a}_{\rho i} = \hat{a}_{\mathscr{z}} \times \hat{a}_{\phi i}$ can now be found and the observer's position, P_i, is located with respect to the dipole if α, β, ρ_i, and $\mathscr{z}_i$ are known.

Suppose measurements are made at several surface points and one point, P_j, is chosen as a reference. The horizontal

*This work was sponsored by the U.S. Department of Energy.

radius vector drawn from P_j to another point P_i is given by

$$\vec{r}_i = \hat{a}_x (x_i - x_j) + \hat{a}_y (y_i - y_j) + \hat{a}_z (z_i - z_j) \tag{4}$$

and $\vec{r}_j \equiv 0$. To get from P_j to the dipole via P_i requires traversing any of "m" paths given by the dipole location vector

$$\vec{D}_i = \vec{r}_i - (\hat{a}_\rho \rho)_i - \hat{a}_z z_i \qquad i = 1,2, \ldots m. \tag{5}$$

By equating $\vec{D}_j$ to each of the other location vectors a set of (m - 1) equations of this form can be written:

$$(\hat{a}_\rho \rho)_j + \hat{a}_z z_j = - \vec{r}_i + (\hat{a}_\rho \rho)_i + \hat{a}_z z_i \qquad i = i \ldots m. \quad i \neq j \tag{6}$$

When each of these equations is dot multiplied by $\hat{a}_{\varphi i}$, a set of (m - 1) equations of this form result:

$$K \left\{ A_i + B_i \cos\beta \tan\alpha + C_i \sin\beta \tan\alpha \right\} = D_i + E_i \cos\beta \tan\alpha + F_i \sin\beta \tan\alpha . \tag{7}$$

The coefficients$(A \ldots F)_i$ are functions only of surface position and measured H-components; the coefficient, $K = |\vec{H}_j| / \rho_j$, is eliminated in the solution process.

Computational Outline

The vertical magnetic dipole has been proposed for the rescue of trapped miners and will be treated first as an introductory case. Measurements are made at two surface points, denoted by subscripts 1 and 2. The cylindrical radius vector which partially locates P_1 with respect to the dipole is given by:

$$\rho_1 = \frac{H_{x2}\left(y_2 - y_1\right) - H_{y2}\left(x_2 - x_1\right)}{H_{x1} H_{y2} - H_{y1} H_{x2}} \left(H_{x1}^2 + H_{y1}^2 \right)^{1/2} \tag{8}$$

because $\alpha = 0$ and we are free to choose $\beta = 0$. The ratio of the H_1-fields along $\hat{a}_{\mathfrak{z}}$ and $\hat{a}_{\rho 1}$ is denoted by γ_1 which can be found from measured values as

$$\gamma_1 = H_{z1} \left(H_{x1}^2 + H_{y1}^2 \right)^{-1/2} . \tag{9}$$

In both Eqs (8) and (9), it is necessary to avoid locating P1 directly over the VMD which would make $H_{x1} = H_{y1} = 0$. The depth may be obtained from the formula for the axial field [1],

$$\mathfrak{z} = \frac{3\gamma_1 \rho_1}{4} \left[1 \pm \left(1 + 8/9\, \gamma_1^2 \right)^{1/2} \right] . \tag{10}$$

One of the two solutions to Eq (10) may be eliminated by physical reasoning, and the computations can be checked by repeating with P_2 as a reference.

In the general case, Eq (7) is solved by using a modified Gauss elimination method and sets of five surface points. Five points appear to be the minimum required to avoid extracting β-values as roots of inverse trigonometric quantities.

The uncertainty in each of the measured field quantities may be as great as 3%, which makes the solution depend upon the order in which the data are presented. Some solutions can be eliminated by dividing each of Eq (7) by its right-hand side and setting each result equal to the next result in turn. This eliminates K, and multiplication followed by collection of terms gives three equations of the form:

$$(GA + GB + GC + GD + GE + GF)_i = 0 \tag{11}$$

where the $(GA, \ldots GF)_i$ are combinations of the coefficients $(A \ldots F)$ and a pair of (β, α) solutions. When the measurement accuracy is unlimited,

$$\sum_{i=1}^{3} (GA + GB + \ldots GF)_i \equiv 0 . \tag{12}$$

so that the amount by which Eq (12) is nonzero may be used to eliminate (β, α) solutions.

The radial distance ρ_i is found by dot-multiplying each side of Eq (6) by $\hat{a}_{\phi j}$ to obtain

$$\rho_i = \frac{-\hat{a}_{\phi j} \cdot \vec{r}_j}{\hat{a}_{\phi j} \cdot \hat{a}_{\rho i}} \qquad (13)$$

because both $\hat{a}_{\phi j}$ and $\hat{a}_{\rho i}$ can be computed for each (β, α) pair. Finally, $\mathfrak{z}$ is given by Eq (10). Depending upon the accuracy of the input data, a number of location vectors may be generated with the use of Eq (5). The best location is selected by computing the fields for each location vector and comparing those fields with the measured field values.

Conduct of a Field Test

A tiltable coil was mounted on the rear of a small electronics package and lowered into a vertical borehole within a three-layered earth near Edgewood, NM. The conductivities of the successive layers had been established as 0.024, 0.165, and 0.046 S/m with interfaces at depths of 9 and 49 meters. The operating frequency of 390 Hz was chosen by comparing the static free-space Hz-field with the exact values computed from Wait [2], assuming a uniform conductivity of 0.2 S/m and a vertical dipole. At a horizontal radius of 10 m and a depth of 20 m, the free-space approximation was in error by about 2 percent. A moment of 0.45 amp-m^2 was used which required 150 mW. The coil was tilted at 65° and lowered to 20 meters. The rotation angle was not controlled but was observed after the test. The surface fields were measured with a pair of tripod-mounted coils, and the origin of coordinates was established by finding a point where good SNR's could be obtained for all three components. The second coil was positioned about the first coil at 5 stepped-off intervals on the perimeter of a circle with a 10-meter radius (Figure 3). Six surface locations were used instead of the minimum of five and the data were processed as six sets of five points. The measured and computed values are compared below.

	x-Offset (meters)	y-Offset (meters)	Depth (meters)	Tilt (degrees)	Rotation (degrees)
Measured	11.4	0.10	-20.0	65	158
Computed	11.4	0.16	-20.6	67.4	157
Difference	0.0	0.06	- 0.6	2.4	1

The results are illustrative of 26 field tests conducted with 9 different combinations of location and orientation [3] which confirm the method outlined here.

References

1. R. Plonsey and R. E. Collin, Principles and Applications of Electromagnetic Fields, McGraw-Hill, New York, NY, 1961, p. 209.

2. J. R. Wait, "Electromagnetic Induction Technique for Locating a Buried Source," IEEE Trans. Geo Sci. Electr., April 1971, pp. 95-98.

3. T.W.H. Caffey, "Locating a Buried Earth Penetrator," Sandia Laboratories Report SAND77-0646, Nov. 1977. Available NTIS, Springfield, Va. 22151.

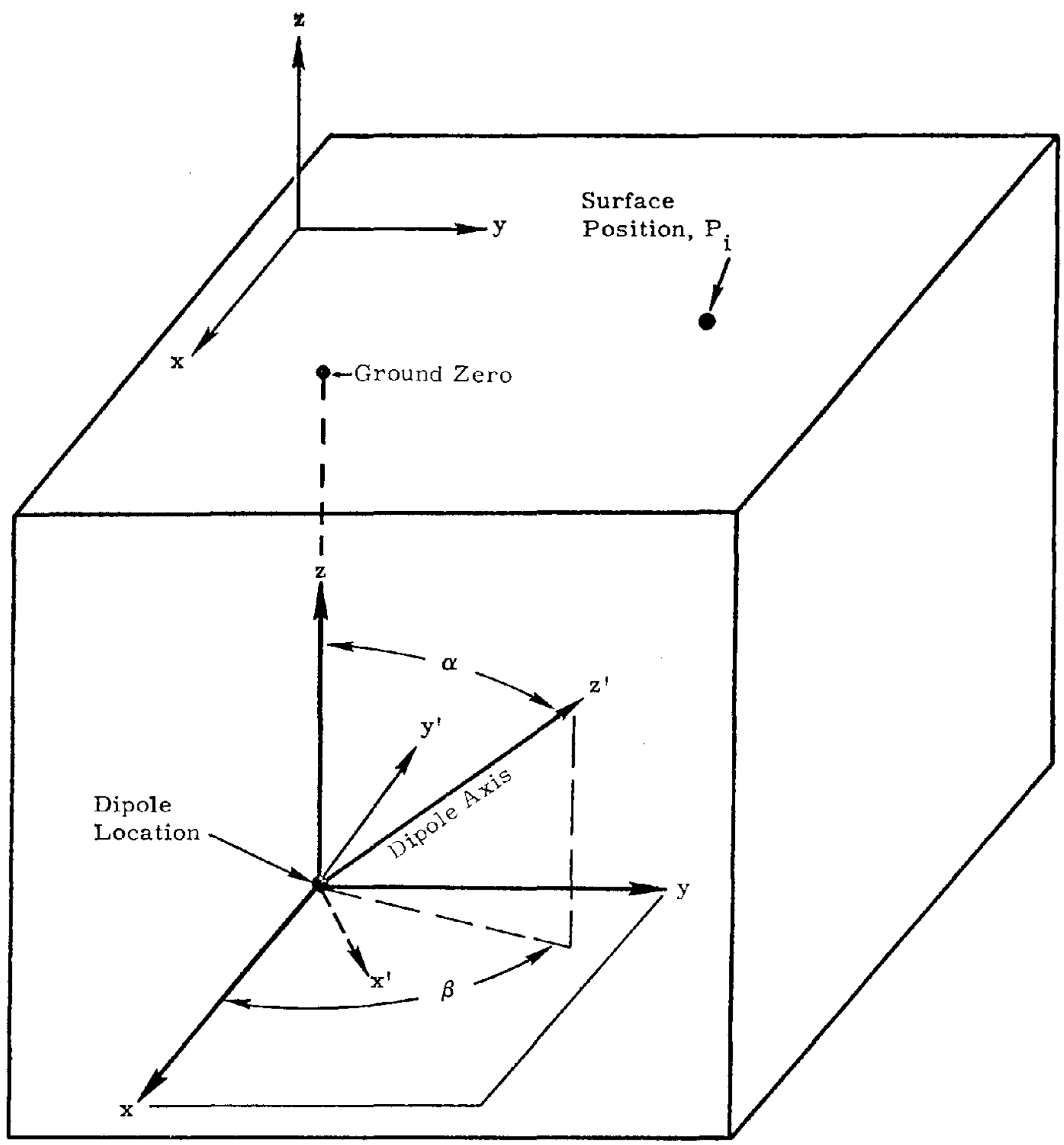


Figure 1. Cartesian Coordinate Frames

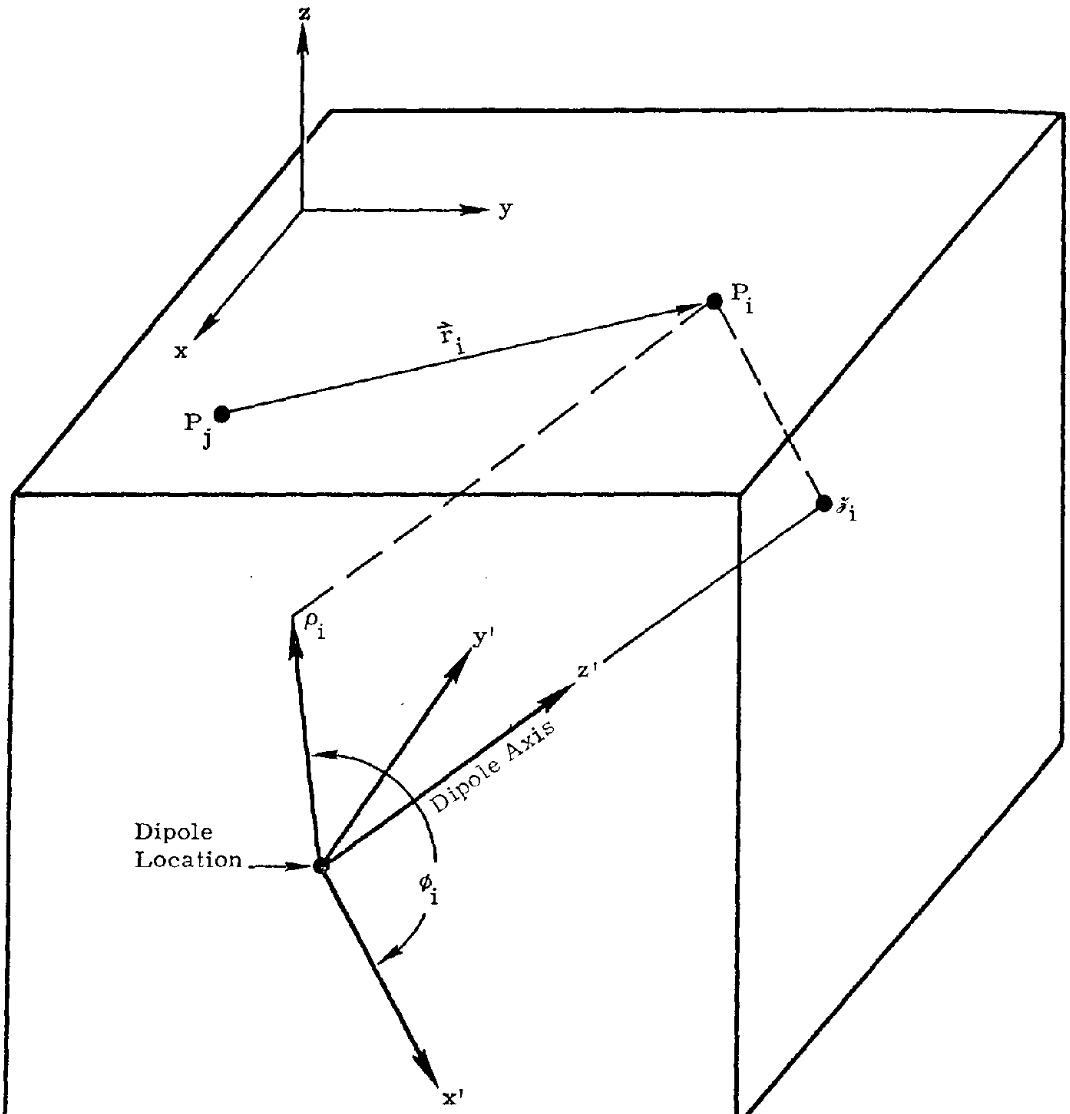


Figure 2. Dipole Location via Cylindrical Frame

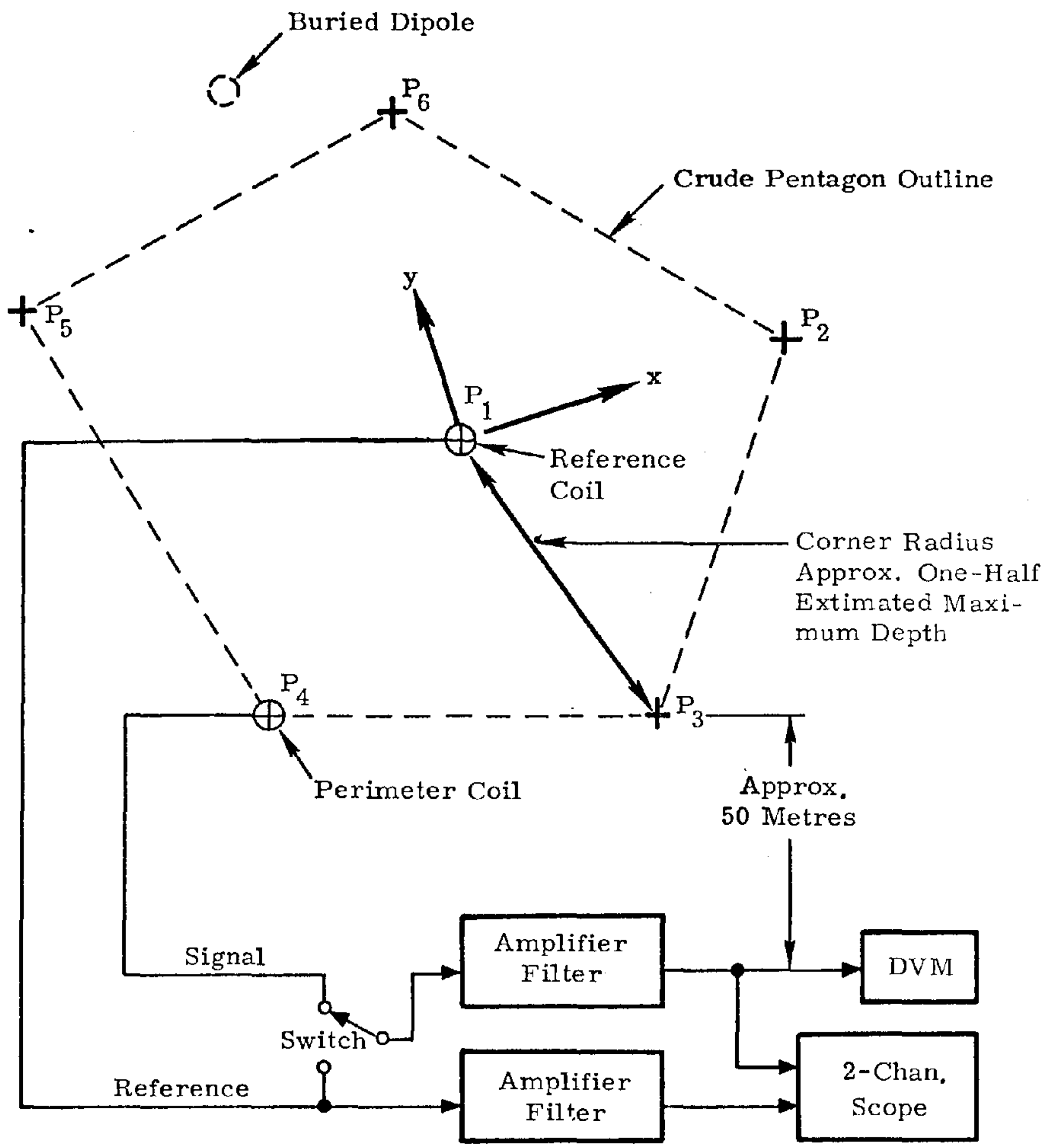


Figure 3. Surface Layout and Block Diagram

QUESTION BY R. GABILLARD

What may be the effect of eddy currents on the accuracy of location, especially in non-horizontally layered earth?

REPLY BY T.H. CAFFEY

Beyond using Wait's formula for the VMD in a conducting half-space as a criteria for choice of operating frequency, the effect of eddy currents has not been considered.

QUESTION BY D.A. HILL

What is your computer search method?

REPLY BY T. CAFFEY

Simply a brute force method which considers a large number of dipole orientations and positions.

QUESTIONS BY J. DURKIN

1. How does radial or depth error vary as function of SNR?
2. How would system perform for SNR = 3 → 6 dB?

REPLY BY T. CAFFEY

Assume that the measure of uncertainty in a field component is given by

$$\delta = 100 \cdot 10^{-\mathrm{SNR(dB)}/20} \text{ percent.}$$

As the simplest case, assume a VMD and that all six field components have the same SNR and same uncertainty. From eqs. (8) through (9), the uncertainty in the results is the same as the uncertainty in the field components. The uncertainties would be 71% and 50% at SNR's of 3 dB and 6 dB, respectively.

One means of reducing the uncertainty in the signal, which I have not tried, is as follows. Let a third measuring coil be stationed well removed from the measurement array, and oriented to sample the spatial component of the noise corresponding to the spatial component being sampled in the array. Let SPN = the instantaneous sum of signal plus noise sampled in the array, and let N = the sample noise value. Then

$$(SPN)^2 - N^2 = Q^2$$

and

$$(S+N)^2 - N^2 = S^2 + 2NS = Q^2 \quad \text{also.}$$

Hence,

$$S = -N + \sqrt{N^2 + Q^2}$$

which is useful only if $Q^2 \geq 0$ and the principal wire source is the background EM spectrum. Alternatively, a signal averager such as a par model 4202 might be useful.

INVERSION OF GEOPHYSICAL DATA

by

Charles H. Stoyer

Colorado School of Mines
Golden, Colorado

ABSTRACT

Inversion allows us to determine an earth model which best fits a set of geophysical data. The model is initially defined by a geometry and a first-guess set of model parameters. Linearization of the forward problem allows us to reverse this process; this reversal is ill-posed. Critical damping of the system removes the singularities and near-singularities so that a best-fit set of model parameters can be obtained in minimum number of iterations. This critical damping is achieved through the use of the principle of feedback.

Results for direct-current geophysical surveys show that a best-fit model can be obtained in two to five iterations, even from a relatively poor initial guess. Similar results can be obtained for transient electromagnetic data, and these forward calculations are obtained from catalog look-up and interpolation. this is several orders of magnitude cheaper than the normal forward calculations, and seems to be quite adequate.

INTRODUCTION

One of the most important problems in geophysics is the inverse problem. Geophysical measurements are taken because we wish to learn more about the subsurface structure than can be obtained from geological or drill-hole data. However, there are many models of the subsurface which will be consistent with a given set of measurements; this is almost without exception. This fact makes inversion difficult, and at the same time, suggests that we need an indication of the ambiguities present in any particular inverse solution. The inversion process described here yields both the model and the extent to which it is ambiguous.

The inversion problem cannot be solved unless the forward solution is already known. That is, we must be able to simulate the geophysical observations which would be obtained from the models of interest. Therefore, if the observed quantities are arranged in a vector, $\underline{o}$, we must be able to obtain the vector of calculated data, $\underline{c}$, as a function of the model parameters, $\underline{p}$, which define a unique model from the chosen class of models. The inversion is then done by choosing an initial set of model parameters, $\underline{p}$, linearizing the forward solution, and iteratively improving the model until the squared error is minimized or falls below a certain specified value. Note that if the data are normally displayed in logarithmic form, the vectors $\underline{o}$, $\underline{c}$, and $\underline{p}$ will contain the logarithms of the observations, calculations and the model parameters, respectively. A by-product of the use of logs is the restriction of the parameters to positive values.

LINEARIZATION

Linearization of the forward problem is achieved by expanding the forward solution in a Taylor series about the current model, $\underline{p}$, and truncating the series to the first two terms:

$$\underline{c}(\underline{p}+\underline{dp}) = \underline{c}(\underline{p}) + \sum_{j=1}^{n} \frac{\partial \underline{c}}{\partial p_j} dp_j . \qquad (1)$$

In matrix form, this becomes

$$\underline{c}(\underline{p}+\underline{dp}) = \underline{c}(\underline{p}) + \underline{A}\ \underline{dp}, \qquad (2)$$

where $\underline{A}$ is the partial derivative matrix and $\underline{c}(\underline{p})$ is calculated using the known forward solution. The partial derivatives can be calculated using finite forward differences.

Now we want $\underline{p} + \underline{dp}$ to fit the observed data, so we write

$$\underline{o} = \underline{c}(\underline{p}) + \underline{A}\ \underline{dp}. \qquad (3)$$

Subtracting $\underline{c}(\underline{p})$ from both sides gives us

$$\underline{o} - \underline{c} = \underline{d} = \underline{A}\ \underline{dp}, \qquad (4)$$

which is the equation we need to solve to find the model improvement, $\underline{dp}$. Note that the squared error is $\underline{d}^{\dagger}\underline{d}$.

CALCULATION OF THE MODEL IMPROVEMENTS

There are two major difficulties associated with the solution of equation (4). It is rectangular because there are generally more data than model parameters, and it is singular because not all of the model parameters can be determined uniquely. The rectangularness can be circumvented by using the generalized inverse (Lanczos, 1960):

$$\underline{A}^{-1} = \underline{V}\,\underline{\Lambda}^{-1}\underline{U}, \tag{5}$$

where the columns of $\underline{V}$ are the eigenvectors of $\underline{A}^{\dagger}\underline{A}$, $\underline{\Lambda}$ is the diagonal matrix of the (positive) square-root of the associated eigenvalues, and $\underline{U}$ is obtained by

$$\underline{U} = \underline{A}\ \underline{V}\,\underline{\Lambda}^{-1} \tag{6}$$

(since $\underline{A}^{\dagger}\underline{A}$ is positive definite and symmetric, its eigenvalues are real and its eigenvectors are orthogonal).

Now, the singularities are removed by applying a filter to the diagonal matrix $\underline{\Lambda}^{-1}$, such that the inverses of the small eigenvalues are decreased considerably, while the inverses of the larger eigenvalues are generally left untouched. The filter used is $\lambda_i^4/(\lambda_i^4 + k^4)$, where λ_i are the eigenvalues in $\underline{\Lambda}$ and k is a constant to be determined. this filter is multiplied times the inverse eigenvalue, λ_i, and damps the system so that smaller and more realistic model corrections are obtained. Critical damping is achieved by finding a near-optimum value for k. This is done by trying several values of k, calculating the vector $\underline{c}(\underline{p}+\underline{dp})$, finding the squared error, and repeating this procedure until a near-optimum value is found or until a certain number of trials have been made. The initial value for k is the better of the last value used, or the value required to produce a certain damping of the model corrections.

RESULTS FOR DIRECT-CURRENT METHODS

The direct current method of geophysical prospecting involves the injection of a current into and out of the ground at two current electrodes, and the measurement of the potential difference between two potential electrodes. These measurements are put in the form of an apparent resistivity; if the ground were homogeneous, this would be its resistivity. The electrode spacings are expanded logarithmically and one observation of apparent resistivity is recorded for each spacing. the logarithms of these observations are put into the vector $\underline{o}$.

The class of models considered here includes only stratified models. For ℓ layers, there are ℓ resistivities and ℓ-1 thicknesses. the logarithms of these are put into the vector $\underline{p}$.

The calculated data, $\underline{c}$, correspond to the observations, $\underline{o}$. These are the logarithms of the apparent resistivities which are calculated by an inverse Hankel transform of order zero. The kernel function in this transform integral is a function of $\underline{p}$ only.

Due to space limitations, only one example will be shown here. The presented paper will contain several additional examples, including inversion of real data and of transient electromagnetic simulated curves.

Figure 1 shows log apparent resistivity versus log electrode spacing for several models. Curve D is the original data, $\underline{o}$, simulated from the model shown in table 1. Curve Ø, which is even of the wrong type to fit the data, is the original guess. The curves generated by subsequent iterations are labeled 1, 2, and 3. As shown in Table 1, convergence was reached in 3 iterations, with a final error of 1.8 percent. We note that the final inverted model is not identical to the model from which the simulated data came.

Of course, we know that the presence of a thin layer such as the "true" model posesses will make the inverse solution nonunique. This nonuniqueness is pointed out by $\underline{A}^{-1}\underline{A}$ as shown in Table 2. This matrix shows the linear combinations of the (logs of the) model parameters which are unique. If the entire model is unique, this matrix, called the parameter resolution matrix, would be the identity matrix. Inspecton of Table 2 shows that the first and third layer resistivities are unique, but that the rest of the parameters are not. In other words, there are five parameters, but only about four unique quantities associated with the model.

CONCLUDING REMARKS

The algorithm described here has been shown to be useful for the inversion of direct current electrical data. Results of research by graduate students at the Colorado School of Mines have shown that it is also useful for inverting transient electromagnetic data (both near-zone and far-zone) and gravity data. Future and ongoing programs include seismic surface-wave and magnetic data inversion. Provided the forward solutions are available (these can be by catalog look-up and interpolation), these algorithms appear to be generally applicable to the inversion of any data whose solution is nonunique.

REFERENCES

Lanczos, C., 1961, Linear differential operators: London, D. Van Nostrand Co., 564 p.

Table 1. "True" model and iterated models during the inversion shown in Figure 1. Resistivities are in ohm-m and thicknesses are in m.

Model	Error	ρ_1	ρ_2	ρ_3	h_1	h_2
"True"	none	100	10	100	5	2
0	48%	60.	90.	60.	10.	10.
1	16%	87.	35.	102.	14.	8.
2	9%	88.	37.	98.	3.7	5.9
3	1.8%	100.	31.	100.	4.7	7.3

Table 2. Parameter resolution matrix. It is symmetric, so only half of it is shown.

	ρ_1	ρ_2	ρ_3	h_1	h_2
ρ_1	1.00				
ρ_2	-0.01	0.71			
ρ_3	0.00	-0.00	1.00		
h_1	0.01	0.18	0.00	0.88	
h_2	-0.01	-0.42	-0.00	0.24	0.40

Note that ρ_1 and ρ_3 are uniquely resolved, while the other parameters are nonunique and uncertain to various degrees.

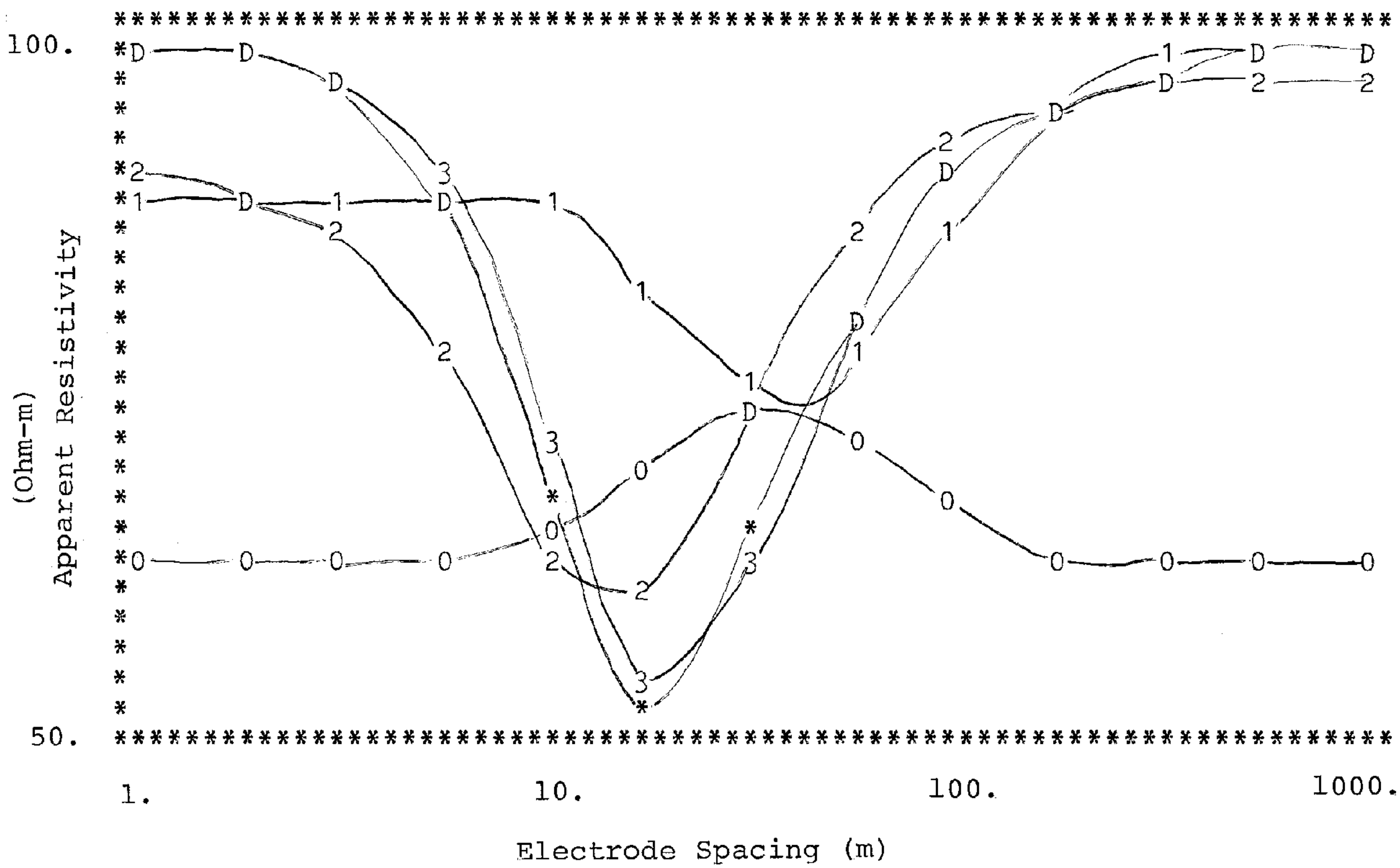


Figure 1. Log-log computer plot of apparent resistivity versus electrode spacing for the models shown in Table 1. Letters 'D' represent duplicate points. The 'data' are plotted with an '*'.

FINDING HIGH CONTRAST ANOMALIES USING CROSS-BOREHOLE HIGH-FREQUENCY CW PROBING

D. T. Davis, R. J. Lytle, E. F. Laine, D. L. Lager
University of California, Lawrence Livermore Lab.
P.O. Box 808, L-156
Livermore, California 94550

Recent tests conducted at a tungsten mine, near Inyokern, California, afforded us the opportunity to observe tunneling through undisturbed ground. That is, we obtained cross-borehole electromagnetic transmission data "before" and "after" a tunnel existed. The tunnel is in granite and is shaped roughly round (7 x 8 ft.), with a flat floor. This shape was used in subsequent theoretical modeling.
An array of four linear boreholes was drilled about 300 ft. from the new adit. These boreholes straddled the tunnel. Using these boreholes, we performed cross-borehole high-frequency (50 MHz) probing using various two-borehole and three-borehole combinations. By lowering transmitter and receiver in a "horizontal" view (having both at the same depth), transmission data were obtained "before" and "after" the tunnel existed. Deep minima were present in the "after" data, but not in the "before" data. From theoretical modeling we learned that these minima are indicative of the tunnel roof and tunnel floor. For example, the vertical distance between these minima is what is theoretically predicted for a tunnel of the same size and shape. By using various combinations of transmitter and receiver depth, we obtained various "views" of the tunnel. An inclined view (e.g., with the transmitter 10 ft. higher than the receiver) gave a stronger indication of the tunnel than a horizontal "view". By superimposing the various views, the tunnel was accurately located both vertically and horizontally. Using the experimental signature, the approximate shape of the tunnel was accurately inferred. An interesting sidelight was that we were able to "see" the rubble removal cart passing through the tunnel.

QUESTION BY J. DURKIN

How effective would this method be in detecting a void in front of a mining face?

REPLY BY E.F. LAINE

We feel our method would be very effective to detect voids in front of a mining face. We are presently preparing such a proposal to USBM. This proposal covers all aspects of underground exploration and includes the problem you pose. The distance and size of the anomaly is governed by the high frequency conductivity of the media. In granite, we would expect to get ranges of several hundred feet in coal, it may be only 50 feet. The cavity size detected is also governed by the propagating frequency; higher frequencies detect smaller cavities but, again, higher frequencies are attenuated more by the media. Our field experience has shown we have been able to detect cavities (tunnels) in a very distinct manner and we were able to locate their position accurately. We were also able to detect quartzite layering which tended to mimic a tunnel.

QUESTION BY J.R. WAIT

What is the validity of the straight-line approximation for the ray paths, e.g. what % error can be expected?

REPLY (COMMUNICATED) BY R.J. LYTLE

In reference to your question on the validity of the straight line ray path method: we are not saying the interaction of the waves with the high contrast anomaly is governed by straight line ray optics. It is a complex diffraction problem not governed strictly by simple ray optics. However, the complex interaction results have an interesting feature. In our published article* (these results are the solution of the complete diffraction problem using only those assumptions characteristic of an integral equation approach), one can note the "signal minima" or shadow region on the "transmission side" of the tunnel. We have used these deep signal minima as the "diagnostic". These signal minima are observable up to 100 feet beyond the tunnel and their shape is well approximated as parallel straight lines passing above and below the tunnel by a distance about 10% of the tunnel height. Thus, this interpretation diagnoses a high contrast anomaly as being larger (by about 20%) than it really is.

*See page 13 of R.J. Lytle "Mapping underground structure with radio waves", <u>Energy and Technology Review</u>, Published by Lawrence Livermore Lab., January 1977.

LIST OF ATTENDEES

J.R. Wait - Senior Scientist, ERL/NOAA and Consultant to ITS/OT, U.S. Dept. of Commerce, Boulder, CO 80303.

H.K. Sacks - Supervisory Electrical Engrg., Pittsburgh Mining & Safety Research Center, U.S. Bureau of Mines, 4800 Forbes Ave., Pittsburgh, PA 15213.

R. Gabillard - Universite des Sciences de Lille, Dept. Electronique (IEEA), B.P. 36-59650 Villeneuve D'Ascq, Lille, France.

P. Delogne - Faculte des Sciences Appliquees, Laboratoire de Telecommunications et D'Hyperfrequences, Universite Catholique de Louvain, Batiment Maxwell, Place du Levant 3, B-1348 Louvain-la-Neuve, Belgium.

D.B. Seidel - Cooperative Institute for Research in Environmental Sciences (CIRES), University of Colorado, Boulder, CO 80309.

A.G. Emslie - Research and Development Div., Arthur D. Little, Inc., Acorn Park, Cambridge, MA 02140.

R.L. Lagace - Research and Development Div., Arthur D. Little, Inc., Acorn Park, Cambridge, MA 02140.

R.H. Spencer - Research and Development Div., Arthur D. Little, Inc., Acorn Park, Cambridge, MA 02140.

D.A. Hill - Institute for Telecommunication Sciences, Office of Telecommunications, U.S. Dept. of Commerce, Boulder, CO 80303.

D.J.R. Martin - Head of Communications Group, National Coal Board, Stanhope Bretby, Ashby Rd. [illegible]on-on-Trent, Staffs DE15 OQD, England.

B.A. Austin - Chamber of Mines of South Africa Research Org., Corner Kew Rd. and Landau Terrace, P.O. Box 809, Johannesburg 2000, Richmond, Johannesburg 2092, South Africa.

A.J. Farstad - Westinghouse Electric Corp., Geophysical Instrumentation Systems, 3655 Frontier Ave., Boulder, CO 80302.

D. Kalvels - Westinghouse Electric Corp., Geophysical Instrumentation Systems, 3655 Frontier Ave., Boulder, CO 80302.

C.A. Balanis - Dept. of Electrical Engrg., West Virginia University, Morgantown, WV 26505.

T.S. Cory - Spectra Assoc., Inc., P.O. Box 2044, Cedar Rapids, IA 52406.

L.D. Whittington - Harrison R. Cooper Assoc., Inc., AMF Box 22014, Salt Lake City, UT 84122.

G.R. Olhoeft - USGS Stop 964, Box 25046, Denver Federal Center, Denver, CO 80225.

R.L. Chufo - Pittsburgh Mining & Safety Research Center, U.S. Bureau of Mines, 4800 Forbes Ave., Pittsburgh, PA 15213.

J.N. Murphy - Pittsburgh Mining & Safety Research Center, U.S. Bureau of Mines, 4800 Forbes Ave., Pittsburgh, PA 15213.

J. Chiba - Dept. of Electrical Engrg., Faculty of Engrg., Tohoku University, Aoba Aramaki, Sendai, Japan.

N.A.M. Mackay - Dept. of Electrical Engrg., Queen's University, Kingston, Canada K7L 3N6.

J.C. Beal - Dept. of Electrical Engrg., Queen's University, Kingston, Canada K7L 3N6.

D.C. Chang - Dept. of Electrical Engrg., University of Colorado, Boulder, CO 80309.

E.F. Kuester - Dept. of Electrical Engrg., University of Colorado, Boulder, CO 80309.

S.W. Maley - Dept. of Electrical Engrg., University of Colorado, Boulder, CO 80309.

R.G. Geyer - Gulf Mineral Resources Co., 1720 S. Bellaire St., Denver, CO 80222.

P.R. Bannister - USN Underwater Systems Center, New London Lab., Fort Trumbull, New London, CT 06320.

T.W.H. Caffey - TM Development Div. 9421, Sandia Labs., Sandia Corp., Albuquerque, NM 87115.

C.H. Stoyer - Institute of Electromagnetic Research, Colorado School of Mines, Golden, CO 80401.

E.F. Laine - L-156, Lawrence Livermore Lab., University of California, P.O. Box 808, Livermore, CA 94550.

R. Kessler - 110 Dartford Rd., Rochester, NY 14618.

L. Burkett - Motorola Corp., 1060 Concord Dr., Medina, OH 44256.

W. Seefeldt - Motorola Corp., 8000 West Sunrise, Ft. Lauderdale, FL 33322.

J. Durkin - Pittsburgh Mining & Safety Research Center, U.S. Bureau of Mines, 4800 Forbes Ave., Pittsburgh, PA 15213.

L.L. Baggerly - TRW-DSSG, 2218 Grand Ave., Claremont, CA 91711.

G. Collins - Cambridge Research Corp., RR1 Dundas, Ontario, Canada.

H. Collins - Cambridge Research Corp., RR1 Dundas, Ontario, Canada.

J. Mason - Dept. of Electrical Engrg., Queens University, Kingston, Ontario, Canada.

R.H. Thompson - The Franklin Institute, Philadelphia, PA.

R.J. Pogorzelski - M.S. M5/1027 TRW-DSSG, One Space Park, Redondo Beach, CA 90278.

H.E. Bussey - National Bureau of Standards, U.S. Dept. of Commerce, Boulder, CO 80303.

J.R. Juroshek - Institute for Telecommunication Sciences, Office of Tele-Communications, U.S. Dept. of Commerce, Boulder, CO 80303.

J.A. Hull - Institute for Telecommunication Sciences, Office of Tele-Communications, U.S. Dept. of Commerce, Boulder, CO 80303.

EXECUTIVE SUMMARY

ELECTROMAGNETIC GUIDED WAVES IN MINE ENVIRONMENTS

This document is the edited Proceedings of a workshop held from 28-30 March 1978 at the U.S. Department of Commerce facilities in Boulder, Colorado. This informal conference was sponsored by a contract from the Pittsburgh Mining and Safety Research Center of the U.S. Bureau of Mines with the Institute for Telecommunication Sciences. ITS is now part of the National Telecommunications and Information Administration.

The purpose of the workshop was to review current research in electromagnetic guided waves that is relevant to mine communications. An excellent background to this subject is found in a significant review paper recently published by Murphy and Parkinson*. As they indicate, underground mines are extensive labyrinths that employ many people working over many square kilometers. Communications are necessary to achieve coordinated tasks. The most desirable method to communicate with key personnel within the mine would be wireless radio either two-way or personal paging. Unfortunately, the range is very limited; for example, at a frequency of 27 MHz, the signal strength is not useable in a mine entry for distances greater than about 30 meters. Three options that have been considered are: 1) to use frequencies sufficiently high that the mine entries act as waveguides, 2) to use frequencies low enough that direct transmission through the seams is possible, and 3) to employ an auxiliary axial conductor or cable that can permit transmission to much greater distances than either 1) or 2). A related telecommunications problem deals with the need to locate miners following a disaster. In this case, the signal transmission may be directly through the overburden--either down link or up link. This subject was the major focus of an earlier workshop sponsored by the U.S. Bureau of Mines†.

In the present workshop, the emphasis was on item 3) above, namely the communication mechanism that utilized axial conductors whether they be especially designed such as the leaky feeder cables or via trolley wires, metal rails, pipes and other in-place conductors. An effort was made to assemble currently active researchers and provide a forum for the exchange of information on this and related topics. About 30 participants from the U.S.A., Canada, Belgium, England, France, South Africa, and Japan attended the full three days. In addition, there was limited participation from organizations in the Boulder area.

The workshop participants were encouraged to take part in the discussions in addition to giving formal presentations. Manuscripts were solicited several months before the workshop and, in most cases, the prompt submission permitted the duplication and distribution of the material to attendees. Reprints of previously published papers on related subjects were also made available.

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*J.N. Murphy and H.E. Parkinson, *Underground Mine Communications*, Proc. IEEE, Vol. 66, No. 1, 26-50, January 1978.

†R.G. Geyer (editor), Thru-the-earth Electromagnetic Workshop, 1973, available from the National Technical Information Service, Springfield, VA 22151, under Accession No. PB231 154.

The present Proceedings consists of, for the most part, the photo duplication of the submitted manuscripts that were available before or at the workshop. However, some editing was performed in order to render a more uniform appearance of the whole document and to clarify terminology and unfamiliar conventions. The lively and uninhibited discussions that took place after each presentation were very worthwhile. Edited versions of these questions, replies and comments are included in the Proceedings after each paper. For the most part, these are based on hand written sheets that were solicited continuously throughout the workshop. In a few cases, supplementary items are included such as relevant excerpts from earlier letters and written communications from the participants. The undersigned assumes full responsibility for any inaccuracies that might have been created in the editing process. Written comments or corrections received from the participants shortly after the workshop have been included in this revised version of the Proceedings.

A surprising unity of viewpoint emerged from the workshop. The concept that transmission is via eigen (or characteristic) modes was a common theme in many of the presentations. Also, for the first time the confusion on the meaning of *monofilar* and *bifilar* to describe these dominant modes was cleared up. However, there was still a healthy divergence of opinion on what constitutes an optimum system. Before a clear cut design procedure can be prescribed, it was generally agreed that further analytical and experimental work is definitely needed. In particular, it was stressed that independent and in-situ measurements of the constitutive properties of the electromagnetic environment are needed. For example, there was considerable concern that laboratory measurements of rock samples could lead to misleading results for the effective conductivities. Also, there was a general recognition that over simplified propagation theory should not be used to interpret field data. Finally, in this same vein, it was appreciated that conventional antenna design techniques need to be carefully scrutinized and revised for in-mine applications

From a theoretical and practical point of view, it was realized that mode couplings play a vital role in the transmission channel. To generalize this concept, we must not only understand which and how many propagation modes are excited by the transmitting antenna, but we also need to account for the conversion of energy from one mode type to another. This conversion can either be done intentionally (by fixed mode convertors) or non intentionally (by random wall roughness). In fact, mode conversion can be quite complicated when various propagation paths exist such as when insulated cables, leaky feeders, grounded rails, coal seams, etc. all convey portions of the electromagnetic energy. However, this state of affairs should be regarded as an opportunity for extended communication coverage in mines rather than a limitation of a design procedure based on preconceived notions of the dominant propagation mechanism.

The opinions and conclusions indicated above are those of the undersigned. In no way are they to be considered the official view of the sponsoring or host agency for this workshop.

15 May 1978 James R. Wait

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SUMMARY OF CURRENT STATE OF THE ART AND SUGGESTIONS FOR FURTHER RESEARCH ON GUIDED WAVE COMMUNICATIONS

Based on Presentations and Discussions
at the WORKSHOP on

ELECTROMAGNETIC GUIDED WAVES IN MINE ENVIRONMENTS

Held at the
Institute for Telecommunication Sciences
Boulder, Colorado
March 28-30, 1978

SUMMARY REPORT
for
USBM Contract No. J0387214

Robert L. Lagace - Project Leader
Alfred G. Emslie and Richard H. Spencer

May 2, 1978

C-81536

Arthur D. Little, Inc.
Cambridge, Massachusetts 02140

SUMMARY OF CURRENT STATE OF THE ART AND SUGGESTIONS FOR FURTHER RESEARCH ON GUIDED WAVE COMMUNICATIONS

I. OBJECTIVE

The EM Guided Wave Workshop was held in Boulder, Colorado, during the last week in March and was attended by an international group of participants who have been working in this field of interest over several years. Since the last such meeting, the Through-the-Earth EM Workshop held in Golden, Colorado, in 1973, much progress has been made and a significant increase in understanding and unification of the results of this progress now appears possible. The objective of this memorandum is not to give a detailed account of all that transpired during this most recent workshop. On the contrary, the objective is to identify major areas or topics of significance to the U. S. Bureau of Mines, to give a broad brush outlook of the state of the art in these areas, and to offer recommendations on where further R & D would provide beneficial results to the furtherance of the Bureau's objectives of improving the health and safety of miners.

II. LEAKY FEEDERS AND DEDICATED WIRES IN MINE TUNNELS

A significant amount of work and progress has been made by the European community in the area of leaky feeder cables for providing mobile communications to vehicles and personnel in haulageway tunnels. Although each of the major proponents, France, Belgium, and the U.K., have taken somewhat different routes in the choice and utilization of cables, there appears to be emerging a unifying understanding and appreciation of the propagation mechanisms that occur in mine tunnels in which these cables are present.

In each case the objective is to provide communications between a roving miner in the tunnel and either a fixed station dispatcher and/or another roving miner located a significant distance away down the tunnel. This is accomplished through the use of a leaky feeder coaxial cable, which allows guided radio energy to "fill" the tunnel cross section, and to propagate down its length with a sufficiently low attenuation rate to allow communications to occur along many kilometers of tunnel length. Agreement seems to have emerged that the methods of excitation and propagation of the so-called guided wave modes in tunnels by different types of leaky feeders really have a common basis of reference for understanding the observed behavior. Until now, this agreement has been somewhat clouded by slight, almost semantic, differences in the definitions, descriptions, or specifications of the so-called "modes" of propagation, particularly with regard to the mode descriptors, monofilar and bifilar, which in many cases have served to, if not confuse the issue, somewhat distract attention from the more fundamental viewpoint of the "natural modes" of multiconductor systems in tunnels in which the tunnel walls serve as one of the conductors. This is discussed in more detail in the following section on Modes.

It also became apparent that there is not necessarily one universally preferable cable or system implementation that is suitable to all mine situations and countries. The important factors of: economics; available equipment, manufacturers, and frequency spectrum; particular application requirements or constraints; and also personal preferences may sometimes play an equal, if not greater, role than the actual propagation characteristics of the cable in determining the final selection to satisfy or meet the overall objectives of a particular mine radio communication system. Thus, it is becoming clearer that there is no so-called "right" cable or cable design that everyone should use, but that each of the cables and techniques developed to date are good ones for solving a wide range of tunnel communication problems. Thus, it is the responsibility of the system designer to select that approach that will "best" suit his specific needs.

Examples of the different European leaky feeder approaches were reported by Gabillard, Delogne, and Martin. The French, as reported by Gabillard, have evolved a specially manufactured, loosely braided, continously leaky cable of highly rugged construction regarding tensile and crushing strength while maintaining a good practical degree of mechanical flexibility. The Belgians, as reported by Delogne, use by and large more conventional, tightly braided cable together with discrete or distributed "mode coupling" sections placed at either periodic or strategically located positions along the cable, which allow larger amounts of cable energy to be coupled directly to the tunnel region external to the cable. The U.K., as reported by Martin, has taken an even different tack and relied mainly on an overall systems approach, whereby optimization of the cable leakage parameters is subordinated to overall system economics and performance requirements. This is accomplished through the use of a slightly specialized cable somewhat more loosely braided than standard coaxial cable, or a specially manufactured BDR twin coaxial cable, together with the use of low cost line amplifiers and/or special interconnecting and bandpass filtering techniques and some rather ingenious overall systems engineering. The operating frequencies range from 7 MHz in France and Belgium to 70 to 90 MHz and 40 to 50 MHz in the U.K.

The U. S. contributions have been by and large on a more theoretical basis, as reported by Wait, et al, aimed at achieving a better understanding of the coupling and propagating mechanisms that occur when such leaky feeder cables are placed in tunnels, as opposed to designing specific systems for implementation in mines. Perhaps the most outstanding exception to this theoretical approach in the U.S. has been the system design and installation of Radiax type leaky feeder cable-based systems operating at 150 and 450 MHz in metal and nonmetal mines by the Bureau of Mines and Motorola. Leaky feeders have also come under study in Canada, and their convenient and favorable use in guided radar surveillance and obstacle detection types of applications was presented by Beal.

By and large it was also shown (Seidel) that the importance of the shape of the tunnel in which the cable is placed is negligible in influencing the rate of signal attenuation down the length of the tunnel, compared with that of the distance of the cable from the nearest wall of the tunnel. However, even this distance is not of critical importance for those cable structures where the energy traveling internal to the cable is only loosely coupled to that traveling in the external tunnel region. It was furthermore noted by Beal that present leaky cables could be placed into four classes, ones having a solid outer conductor with holes, a solid outer conductor with a longitudinal gap in the shield running the length of the cable, a solid but very thin outer conductor with no openings, and a loosely braided outer conductor. Although the geometrical form factor of the fields in the tunnel external to the cable would be identical for these different cables, the longitudinal attenuation for each could be significantly different. This difference can be caused by the nature of the current interruptions caused by the openings in the outer conductor of the cable in some cases, while in others, such as the cable with the longitudinal gap, by the strong interaction the internal fields with the tunnel wall if that type of cable is placed close to the wall. This subject may be worthy of further investigation to better clarify the impact that the shield openings or coupling has on the longitudinal attenuation rate, cable economics, and practical mine installations.

Further insight into the unity of behavior for the different cable systems was also obtained by observations of the similarity of the experimental signal versus longitudinal distance results for the different types of leaky feeder cables, between experiment and theoretical results, and between the results obtained for leaky feeders and those for a dedicated wire (Spencer) placed in the vicinity of a trolley wire/rail transmission line. The observation of these similarities led to considerable discussion of modes and so-called mode conversion and description, etc., which is summarized in the following section on Modes. Also discussed in the following section is the proposition that the propagation of radio waves along leaky feeders in tunnels is generally not a radiation process, but a multi-mode guided wave process.

In summary, the time now appears to be ripe for a consolidation and integration of the work that has been done independently by investigators from several nations. This should take the form of a unifying basic description and explanation of the coupling and propagation processes, together with the engineering and economic factors that should be considered in arriving at suitable alternative cable choices for specific types of mine applications. From such an investigation, practical guidelines should emerge for the design and implementation of practical systems by mine communications suppliers and mine operators for a wide variety of mine communication applications.

III. MODES OF PROPAGATION IN TUNNELS

The title of the conference, "EM Guided Waves Workshop," expresses the idea of radio communication by means of various possible modes of propagation of electromagnetic waves. Thus, the subject of modes on wire and cable based transmission systems came up, and there was considerable discussion of modes and mode conversion. Although considerable progress has been made in the description and understanding of these terms with regard to the leaky feeder application, there is still room for improvement in clarifying the differences between characteristic natural modes that are generated and propagate on a complex arrangement of conductors, and the idea of mode "conversions" created by discontinuities, whether periodic or random, on such propagating structures. Furthermore, when such discontinuities become continuous, questions arise as to the most reasonable and rigorous ways to describe and represent the resultant propagation phenomena, i.e., whether in terms of mode conversion or the generation of a unique set of natural modes. Thus, it seems important that commonly accepted and unifying definitions of appropriate parameters and descriptors for representing modes and mode conversions be developed.

A beginning of this process is attempted here. As mentioned above, the key word "mode" was used repeatedly throughout the meetings and was the subject of some discussion with regard to definition. There seemed to be general agreement that a pure electromagnetic mode should be defined as a form of EM propagation in which a configuration or distribution of currents and voltages, or electric and magnetic fields, propagates with a unique propagation constant which causes it to die off with distance with a single exponential decay factor. Several such pure modes with different decay factors will in general be excited in a tunnel containing multiple conductors if the radiation or excitation source is not matched to the spatial shape and boundary conditions for a single pure mode. The term "mode conversion" was also used frequently during the meetings and was the subject of some confusion. Pure exponentially-decaying modes are orthogonal, and thus do not interact. Coupling, transfer, or conversion of energy from one mode to another can only occur then when some special form of change or discontinuity is introduced at some point along the direction of propagation.

One form of confusion on this issue is the one where so-called "conversion" between modes is said to "occur" when some form of continuous coupling is introduced which alters the original physical system, and the modes of the original system are still retained to describe the behavior of the altered system, rather than using the pure modes of the new or altered system. Under these conditions each so-called "mode" does not decay with a single exponential decay factor, and the term "mode conversion" can be a confusing one. An example is the case of a coaxial cable, with a loosely-braided outer conductor, installed in a coal mine tunnel. The resultant propagation behavior is commonly described in terms of the coupling between two physically decoupled modes, namely the coaxial mode inside the cable itself for the case of an impervious outer conductor, and the coaxial mode within the tunnel, when the impervious cable braid is regarded as the inner conductor and the conducting medium surrounding of the tunnel as the outer conductor. These two physically decoupled "modes," typically called bifilar and monofilar, respectively, are not the pure natural modes of the conductor configuration for a real, leaky coaxial cable having a nonimpervious braid installed in a tunnel. However, though the terminology may be confusing, this type of analysis does yield valid results that have been confirmed experimentally.

An alternative way of analyzing any such multi-conductor transmission-line system is to calculate the pure modes of the actual total system, including the coupling. Then the problem reduces to determining the mode amplitudes caused by any given type of excitation or discontinuity of the system. This method has been used by ADL to describe the propagation mechanisms and to calculate the beneficial effects of placing a dedicated wire in a tunnel containing a heavily shunt-loaded trolley wire/rail transmission line. These effects have also been observed experimentally during an in-mine dedicated wire experiment. As noted previously, the theoretical propagation curves for the dedicated wire installation exhibit the same longitudinal attenuation behavior observed experimentally for continuous loosely braided coaxial cables in tunnels. Dedicated wire calculations have also been made by Wait and Hill for the case of a single lumped shunt load instead of a continuously distributed shunt loading as in the ADL calculations. In this case the pure modes for three unshunted conductors in a tunnel are derived, and the effect of the lumped shunt on the distribution of the power amongst the pure modes is calculated. These results support the dedicated wire behavior predicted by the ADL continuous shunt load approach.

The same concept of two pure, or eigen, modes is also helpful in many other instances, such as the various forms of leaky cable communication system that have been investigated by the workshop participants. The simplest arrangement is where the leakage is uniform along the length of the cable, as in the case of a uniform loose braid. As described by Wait, Seidel, Hill, and others, the two eigen modes consist of a "bifilar" mode, in which the field is mostly concentrated in the coaxial cable, and a "monofilar" mode, in which the field is mostly concentrated in the space between the coaxial cable and the tunnel wall.

Other cable designs incorporating (a) holes, or slots instead of loose braid, (b) equally-spaced discrete couplers in a braided leaky cable, (c) braided cable with sections of more loosely braided, i.e., more leaky, cable, and (d) continuous, thin solid shield cable were also discussed. However, it was not apparent that such designs had any significant advantage over an optimized continuously leaky braided cable, except possibly an economic one and/or one related to a particular practical engineering or system requirement. Thus, as recommended in the previous section, an overview assessment which addresses these factors would be beneficial to perform.

The theoretical work presented by Kuester suggests that there may be a clear and unified way of understanding guided waves on cables and other conductors located in the vicinity of conducting surfaces. If this is possible, the subjects of coupled transmission lines, leaky feeders, so-called monofilar and bifilar modes, evanescent and radiating modes, etc., in tunnels could probably be handled under one generalized theory, from which each specialized case could be developed, classified and understood using the techniques of Kuester. Thus, it appears that some effort aimed at assessing the practical application of the Kuester work is warranted.

IV. APPLICATION OF IMAGE THEORY

At frequencies below HF, particularly in the MF band, another type of guided mode quite different from the leaky feeder guided mode described above is seen to propagate in the coal seam waveguide. This is a two dimensional TEM transmission line type of mode guided by the conducting rock above and below the much less conducting coal seam. This mode can be excited by several types of antennas or mode exciters placed within a tunnel in the coal seam. The principal means of excitation used to date has been the vertically oriented loop antenna; however, horizontal grounded long wires and vertical grounded wires also appear to have favorable performance potential for fixed station installations. The simple three layer theoretical model used by ADL to describe the coupling and propagation behavior of this mode in coal seams is in good agreement with the experimental data taken by Cory within the frequency band of interest for MF mine wireless communications. The results also prove that the transmission loss experienced by the waves depends critically on the conductivities of both the rock and the coal, both of which are seen to vary considerably from seam to seam within the U. S. coal fields.

In all seams, having either high or low loss, the presence of conductors, such as power cables, in the same or neighboring tunnels greatly enhances mobile to mobile communication range. A simple theory has been formulated by ADL to model the coupling of the coal seam propagation mode to such conductors in the seam. Although these theoretical results also appear to be in reasonable agreement with experimental observation, the type of data taken to date, particularly in terms of orientation of the loop antennas, is not sufficient to confirm or deny the validity of this simplified theoretical approach. Consequently, at least one additional set of in-mine measurements should be conducted in the vicinity of such cables in mines, and in such a way as to test for the behavior predicted by the ADL theoretical approach.

In addition to the simplification allowed by the use of a three layer model to represent the coal seam and the surrounding rock above and below the seam, considerable simplification in the derivation of the coupling factors quantifying the coupling mechanism between a loop and the coal seam waveguide mode and between a loop and the current carrying cable within the coal seam waveguide were derived. This was achieved by using an image theory approach similar to that espoused by Bannister of U. S. Naval Underwater Systems Center, New London Laboratory, in many of his applications related to ELF to VLF radio communications with submerged submarines. Questions regarding the validity of his approach for the dimensions, frequencies, and conductivities of the coal seam waveguide, were raised, and Bannister supported the use of this approach for the application in question. In addition, Bannister addressed several situations and conditions for which complicated theoretical analyses could be avoided by using image theory techniques and solutions to provide good approximate practical performance estimates. Since this theory of images can provide good approximate answers for what might be quite complicated propagation environments, some additional communications with Bannister should be undertaken. The objective should be to more clearly understand and define those situations where image techniques can be readily applicable to coal seam mode propagation problems, and to subsequently document this in the form of a memorandum report for distribution to investigators working in the area of coal seam mine communications.

Another semi-empirical technique for describing and quantifying the effects of conductor aided wave propagation in coal seams, namely, the concept of scatter gain, was presented by Cory. This concept is used in several free space type of propagation applications where scattering from structures and objects is important. Thus, further examination of its potential utility for describing and quantifying conductor aided propagation in mines appears to be warranted.

V. UHF RADIO WAVE PROPAGATION

Yet another type of guided wave mode that propagates well in coal mine tunnels is the waveguide mode at UHF. This waveguide mode is confined essentially within the air duct defined by the four walls of the tunnel and propagates down this tunnel without the aid of conductors.

Although many modes are excited by a transmitter in the tunnel and regenerated by discontinuities and changes in the tunnel cross section, only the dominant lowest order mode usually prevails over long distances (between 1000 - 2000 feet in U. S. coal mines) because of the significantly higher loss associated with the higher modes. In a rectangular tunnel the dominant polarization is furthermore the one which is parallel to the long dimension of the rectangular cross section. The theory describing this mode was first reported by ADL at the Through-the-Earth Electromagnetics Workshop given in Golden, Colorado, in 1973, and further work has been pursued and reported by both ADL and Wait.

Since the publishing of these reports, several in-mine experiments and tests have been performed by representatives of Motorla in both coal mines and metal and nonmetal mines. It was reported at this workshop that operational type tests using hand held walkie talkies at both 450 and 850 MHz confirmed the results predicted in the ADL reports. Furthermore, it was stated that there is an increasing trend toward the use of 450 MHz, and in the near future 850 MHz portable radio equipment. Consequently, hardware suppliers would be in a better position to serve the needs of mine operators if they had access to more quantitative experimental data than there is available to date, to support or modify the theoretical predictions. It is the scarcity of this quantitative data that makes it difficult to specify systems and costs with the confidence required to implement practical systems without undue risk.

In support of this need for experimental data, Chufo of the Bureau of Mines (PMSRC) recently completed a rather comprehensive series of UHF measurements in a limestone room and pillar mine at the frequencies of 466 and 812 MHz. Furthermore, he demonstrated experimentally that the installation of passive reflectors near the roof of the tunnel at strategic intersections can greatly extend the range of UHF signals around corners, which normally introduce large and prohibitive corner losses otherwise. This rather unique contribution should be given wider dissemination by publication in the professional journals. The measurements also confirmed the predicted decrease in attenuation rate with frequency in a tunnel without crosscuts. However, the attenuation behavior in the room and pillar area tunnels was found to be about the same at both frequencies. It is presently believed that the significantly greater ratio of crosscut area to pillar area is responsible for this frequency independent behavior. A brief theoretical extension of the previous ADL theoretical results should be conducted to confirm the plausibility of this explanation.

It was noted that the smooth-walled railroad tunnel proagation results at UHF reported by Chiba of Japan also support the smooth wall model results obtained by ADL and others, which predicts a monotonic decrease in attenuation rate with increasing frequency. Since Chiba's application will include blockage due to the presence of trains in the tunnel, he was interested in and should be sent the ADL theoretical results on the effects of obstacles in tunnels. The Bureau of Mines should also request Chiba to pass on results of further work to U. S. investigators.

VI. INDEPENDENT MEASUREMENT OF ELECTRICAL PROPERTIES OF COAL AND ROCK

Two approaches to this problem were treated during the workshop, one being the laboratory measurement of the electrical properties of coal and rock and the second being independent in situ, direct or indirect, measurements of these properties in mines. Laboratory measurement results were presented by two authors, Balanis and Olhoeft. The electrical properties were found to be a function of water content, pressure, temperature, and direction of electric field excitation. Because of the anisotropy of the sedimentary rock and coal media, the electrical characteristics are different along three orthogonal axes. The conductivity in particular was higher in the direction parallel to the bedding planes of the coal. Temperature assumed a high importance only at the highly elevated values found in coal gasification applications. Nonlinear effects can also be expected for certain ranges of frequency and source strength. It was suggested that measurements made in coal should be tried at several levels of excitation to determine if nonlinear effects occur over the source strength level ranges of interest.

Of particular importance is the dependence of the conductivity on the moisture content associated with the coal in its natural in situ state. Furthermore, an extremely strong conviction was expressed by Olhoeft concerning the extreme difficulty of inferring in situ properties of coal and rock materials from laboratory measurements on these materials. The disturbing effects created by the act of acquiring the sample are critical in that they introduce cracks and other changes in the sample that could critically affect the electrical properties. These effects are particularly problematical for coal if attempts were made to acquire small samples. Large scale bulk samples apparently are necessary if one wishes to obtain representative in situ behavior.

A wide variety of information was presented in detail by Olhoeft of the U.S.G.S. on the precautions and methods required to simulate or reproduce the in situ material environment during the course of laboratory measurements. Of principal importance was the ability to subject the material to the in situ water content environment, i.e., under high pressure. The experience of Austin of South Africa was very much in agreement with the observation that laboratory measurements of material properties in no way give reliable estimates of the in situ properties of these materials. The essence of the whole discussion suggests that there is a great deal yet to be learned about the in situ properties of coal and rock, and that a lot of valuable information and techniques on this matter rests with Olhoeft of the U.S.G.S. This information should get wider publicity and dissemination, and possibly utilization.

The subject of making in situ measurements of the conductivity of coal and rock received considerable attention during the workshop, unfortunately without a clear, simple, unambiguous and practical method emerging. At present the material properties must be inferred from amplitude versus distance and frequency measurements in the mine. Although this is an indirect method, there was a feeling among many of the participants that the averaging process over a long distance that is inherent in this particular method may be sufficient to override the somewhat unfavorable and undesirable aspects of having to determine the electrical properties by inference. Another averaging process technique that can provide a second independent check on the material properties involves the measurement of the phase constant β over the measurement path, by measuring the total phase shift between the transmitter and the receiver. Unfortunately, in most instances this may be extremely difficult to achieve because of the need to deploy an extra wire or wire pair in the measurement area to transmit the source phase reference to the receiver. This may not be a practical or valid option in most coal mine tunnel measurements because of the coupling of the coal seam mode to the cable guided mode, thereby confusing the measurement results. Thus, if this method is to be pursued, it will be necessary to investigate the feasibility of using either a precise frequency standard to establish the reference phase, or the use of independent communication links, either radio or optical, for transmitting the phase reference information to the receive location. The cost/benefits to be derived from such a measurement should also be assessed before proceeding with it.

Several other methods were proposed including the use of grounded electrode effective conductivity measurement methods used typically on the surface, probes placed in predrilled holes, transmission measurements between drilled bore holes, special inductive loop techniques, and the measurement of open and short circuit impedances of a two wire line placed in the immediate vicinity of coal pillars and/or the roof and floor. Problems were identified with each of the techniques, thereby requiring a more careful assessment of the desirability and practicality of making such a measurement. A potentially serious problem with most of these techniques is the fact that they sample the material for only a short penetration depth into the medium where the properties may be quite different from those of the bulk of volume not sampled. For example, Gabillard of France has observed that exposed surfaces in a mine lose their water content and the material dries out to significant penetration depths over a period of time, depths that are comparable to those being sampled by the above methods.

Thus, although important to determine the electrical properties independently, it was not possible to identify during the workshop a method satisfactory to a majority of the attendees, the one exception being the independent measurement of phase angle over the propagation path, which of course requires a means to obtain an independent source phase reference at the receiver. The results of Davis in performing such measurements do not seem to be applicable because of his utilization of an independent wire in the tunnel measurements, and the presence of a conducting support wire in the related surface-based airport measurements.

VII. THE SOUTH AFRICAN EXPERIENCE WITH MF RADIO

The presentation of, and subsequent discussions with, Austin of South Africa indicate that a large amount of significant work in the development of MF wireless radio for mines has been done in South Africa, work from which the U. S. mining industry can benefit. Unfortunately, most of this work has not been documented in the U. S. open literature. Thus, it would be highly desirable to encourage Austin and his associates to publish the results of their work, or to remain in close correspondence with the U. S. Bureau of Mines and its contractors regarding past achievements, and progress in the future stemming from things learned as a result of this workshop and their own independent investigations. Similarly, the U. S. Bureau of Mines should be encouraged to provide Austin with similar information on work being performed in the United States.

Of particular interest to the U. S. Bureau of Mines is the development of rugged, intrinsically safe, MF radio hardware that can be worn without inconvenience by the miners, and provide highly reliable radio communication to ranges of about 300 m. The SAR equipment in this latest advanced stage of development appears to satisfy all the driving forces and constraints imposed by the South African gold mine rescue type operations. These operations require not only high performance but extreme durability in a highly hostile environment, worn by a man who must work in extremely confined and uncomfortable conditions.

Particularly encouraging is the finding that Austin and his colleagues have come to many of the same conclusions as U. S. investigators with regard to the shape, size, and strength of the mobile personal transmit antenna worn by the miner. Their present radios also use a frequency synthesizer which allows flexible operation within the frequency band from 100 kHz up to 2 MHz, the band which is also of current interest to the U. S. mining industry. The conductivity of the rock in South African gold mines is typically on the order of 10^{-5} mhos per meter, which is significantly lower than that found in many U. S. coal mines, although some U. S. seams do approach this value. Nonetheless, Austin and his colleagues have found the units extremely useful in coal mine applications also, particularly when in the vicinity of conducting cables and the like. This finding has been corroborated by experimental work sponsored by the Bureau in U. S. coal mines. One of the main system design differences between the U. S. and South African MF radios is the use of single sideband AM modulation by the South Africans. According to Austin, this was the most logical and economical choice for their application within an environment that was not highly contaminated by impulsive electromagnetic noise. When in the vicinity of such noise, operational procedures are such that a slight movement of the person to a nearby position sufficiently free of the noise environment quickly allows the use of the radio. The radio also has a narrow band signaling feature, for use primarily during rescue operations where voice communications are not possible because of face masks or other constraints. Thus, it is conceivable that this tone signaling mode of operation would allow a man standing in an electrically noisy environment to be signaled by someone trying to get his attention, after which he could move away from the source of noise and carry on a two way radio voice conversation.

During the discussion, Austin noted that he found that in one gold mine it was the most favorable loop orientation to be coplanar, but in the horizontal plane, not the vertical. This finding is counter to the behavior noted in U. S. coal mines. However, it can possibly be explained by the presence of vertically oriented fracturing or perhaps igneous inclusions or dikes of higher conductivity located on both sides of the loops, thereby forming a vertical parallel plane waveguide.

In addition, communication of data has been accomplished over the mine power lines extending from three kilometers in the mine up to the surface, and also up drill strings. Austin also related what he found to be the "substation effect," of being able to better couple into power cables at transformer substation locations where the individual conductors in the cables splayed out from the cable protective covering to make appropriate connections in the substation. The transmission of data over the mine power lines to the surface was also found to exhibit unpredictable adverse behavior which was untraceable because of the vastness and complexity of the power network on which the signals were superimposed. Worsening this situation was the inability to know when and where a mine operational change was initiated which did not change the power operation of the electrical network, but did change the MF propagation characteristics of the cable supported guided wave. The South African units have also been used to provide voice communication in vertical hoist shafts as deep as 1700 meters to personnel standing either just to the side of the shaft and/or within the cage. D. J. R. Martin commented that communication to cages in 1000 meter deep hoist shafts has been accomplished in the U. K. by using equipment operating at 27 MHz and at 170 MHz, the latter being of current choice because of the availability of commercial hardware. Finally, Austin stressed the need to define some sort of frequency control or allocation plan for the use of radio and wire line signals in mines, so that future chaos in this area could be avoided. This is an undertaking that is presently being pursued jointly by the U. S. Bureau of Mines, the IEEE, and the coal industry, and interaction in this area should be pursued between the two countries.

VIII. UTILITY OF THE EFFECTIVE CONDUCTIVITY PARAMETER

The validity of the use of the so-called effective or apparent conductivity was discussed. This was treated in the context of the through-the-earth signal transmission loss measurements being performed between the surface and coal seams for assessing the performance effectiveness of trapped miner location equipment in the U. S. coal fields. The principal quantity of interest is the actual transmission loss, which is the quantity that is being directly measured. Thus, precautions were issued against attaching too much significance to any effective conductivity derived from these loss measurements on the basis of a homogeneous earth model. Precautions were also issued against using effective conductivity values determined as a result of any four-probe surface-based measurements. It was felt that this effective conductivity parameter should be used only with great caution and precision with regard to what it represents, and what it does not represent, and to particularly refrain from using it as a means to explain difficult to interpret measurement results.

IX. SUBSURFACE TO SURFACE SIGNAL STRENGTH MEASUREMENTS - TRAPPED MINER LOCATION

It was noted that some of the work related to detection and location of trapped miners still seems to display interest in obtaining location accuracies that could be described as better than will actually be required during most mine rescue operations. During most mine rescue operations, detection of the presence of a miner will essentially be tantamount to his location. For example, the coal mine rescue problem represents an area search in which miners may be dispersed over a widely spaced planar area. In many cases just the determination of a miner's gross whereabouts, namely in what section of the mine is he in, will provide valuable information that will most likely allow him to be rescued, because the efforts of rescue teams can then be directed toward those areas where miners are known to be alive. In some cases a finer location may be required, to within a pillar or two, and the present trapped miner system will more than provide that capability in a coal mine environment.

In metal/nonmetal mine environments represented by large disseminated ore bodies or vein types, the mine is stratified more vertically than it extends in area horizontally. The objective in these mines will be different, namely to determine what level the miners are trapped on in most cases. The lateral position of the miners may not be as critical to know. This may most logically be determined by means of transmitters coded in terms of tones or digital codes which represent the location of the miner with respect to specific or approximate levels in the mine. The problem of determining exact location from an electromagnetic transmitter in many metal and nonmetal mines is compounded by the fact that such mines usually contain a large number of metal conductors deployed along the same paths that the electromagnetic signals must traverse to the surface. As a result, the presence of these conductors will distort and enhance the fields produced by these transmitters, thereby causing considerable ambiguity in terms of geometrical location of the transmitter, while at the same time extending markedly the range over which a signal can be received. Consequently the detection and location strategy to be used in metal/nonmetal mines will most likely evolve to be something quite different from that used in coal mine environments. Attention should therefore be addressed to this reality and its associated problems at all times to ensure the development of the most effective and practical metal/nonmetal trapped miner detection and location system.

X. PERSONAL ANTENNAS AND WAVEGUIDE MODE EXCITERS

Considerable attention was given by several speakers during the workshop to defining the limitations and potentials of portable unit antennas for MF band mine wireless communications. Universal agreement was obtained on the point that loop antennas were far superior to electric dipole antennas for the application and frequency band of interest. The primary mechanism for the transfer of energy from one transmitter to a remotely located receiver is not the conventional radiation process, but one involving the coupling of energy to the waveguiding medium and the sensing of that energy by the receiving unit. South African and U. S. efforts have been aimed at obtaining the largest practical source strengths to provide completely wireless communications as a primary goal. They have thus concluded that air-core type bandolier antennas could most easily meet this need within the limits of intrinsic safety. It appears, on the other hand, that the Europeans are primarily concerned with achieving much smaller portable units, mainly as a result of their desire and ability to use the radios in the vicinity of leaky feeders, and because of a high miner resistance to the use of the more clumsy and constraining air-core bandoliers. Therefore, much smaller, lower-power, ferrite antennas integrated in the radio unit have been used by the Europeans. However, recent work in the U. S. (ADL) appears to support the possibility that an equal strength ferrite-core loop antenna of significantly smaller and more convenient dimensions than an air-core loop bandolier may be able to provide the same performance. Since ferrite materials, in the form of magnesium zinc ferrite, are now available that may allow the achievement of such performance in a practical manner, some effort should be devoted to designing and fabricating a prototype unit to verify whether in fact this is now possible. It appears that similar efforts will also be initiated by the South African investigators.

Parallel plate waveguide mode exciters of both the grounded horizontal long wire and vertical wire varieties were addressed. They may be capable of achieving higher performance than vertical plane loops for certain fixed station installations. These mode exciters are similar to those considered for use with the Sanguine ELF communication system in the earth ionosphere waveguide. They seem to have sufficient merit to warrant further analysis to determine if some of these techniques can help solve either operational or emergency warning types of mine communications requirements.

Gabillard of France has had much experience in evaluating such configurations in French mines. Significant transmission experiments have been performed by Gabillard with vertical wire grounded antennas and long wire horizontal antennas in gypsum mines. These experiments led to the development of reliable means of providing solid electrical grounds to rods implanted in the roof of such mines. Gabillard has also noticed that the electrical characteristics of exposed mine walls, roofs, and floors can exhibit significant changes as the moisture in these walls is lost due to prolonged exposure to the ventilated mine environment. Thus, the operation of radio equipment connected to these walls, and the measuring of material properties at short distances into the materials of these walls, are likely to produce significantly different behavior and electrical properties than would be obtained if it were possible to sample the material in the bulk, away from the regions where it comes into contact with the tunnel cavities. Gabillard's experience also extends into the accurate location of magnetic dipole sources beneath the surface. The wide experience of Gabillard in the area of mine communications and location should be extremely useful to the U. S. Bureau of Mines in the pursuit of its objectives. Therefore, the Bureau should endeavor to obtain any relevant publications in the open literature by Gabillard, and to encourage continued communication and interchange between him and the U. S. Bureau of Mines on past, current, and future work in this area.

XI. RADIO TRANSMISSION ALONG A METAL ROD IN A CONDUCTING MEDIUM

The work of Wait and Hill in this area appears to be worthy of further investigation, particularly since it appears that propagation along bore hole casings and/or drill strings could be accomplished with the expenditure of relatively little power. Considerable discussion followed this talk, principally with regard to a South African application on the transmission of data up a drill string of a "raise-type" of cutting head subjected to an extremely harsh mechanical environment. Of particular interest were the means to best couple into the drill string transmission line while not exposing the antenna or coupler to the destructive forces of the drilling process. Ferrite loaded loops and capacitive type couplers were suggested by the participants. A ferrite loaded signaling source could also be used as a means to locate such a highly expensive drilling bit on those occasions when the drill string is sheared due to excessive torques placed on the bit.

XII. LOCATION OF TUNNELS AND ACTIVE BURIED SOURCES

It appears that the cross bore hole transmission method of locating tunnels presented by Laine, and the accurate location of small earth penetrators equipped with active magnetic dipole sources presented by Caffey, may prove useful to the Bureau of Mines for the location of lost bore holes, gas wells, occlusions, coal gasified volumes, etc., and for position location in either vertical, or horizontal within the seam, drilling related operations.

XIII. GUIDED RADAR USING PARALLEL LEAKY FEEDERS

The Canadians, represented by Beal and McKay, have been involved in leaky feeder work for some years, and thus have accumulated considerable experience in this area. They have recently applied this technology to a guided radar system capable of detecting obstacles for providing security and surveillance around penitentiaries, industrial sites, etc., or for the detection of the presence of moving objects. As part of this work they have developed testing fixtures and tests for the accurate measurement of the coupling loss of leaky feeder cables. This method appears to be more accurate and practical than devices and test procedures developed by others to date. The performance of the guided radar system appears to be quite favorable. As such it should probably receive a brief examination with respect to the possibilities for its use in coal mines or metal mines for identifying and/or detecting the location or presence of locomotives and/or cars or other objects, perhaps in critical areas, as a means of achieving safer and more efficient mine operating conditions. Such a system may also be applicable to the measurement of the position or detection of a hoist at critical locations in a shaft.

XIV. ELECTROMAGNETIC HAZARDS TO BLASTING

An ad hoc session was allocated to this subject at the beginning of the second day because of its increasing importance to the U. S. mining industry as the use of radios underground becomes more widespread. Thompson of the Franklin Institute presented the results obtained to date regarding the potential hazards of unintentional firing or detonation of blasting caps by radio equipment used in their vicinity in mine tunnel environments. A great deal of discussion ensued regarding the special problems and techniques applicable to the wide range of frequency bands being considered for underground mines, and the possible sources of radio energy, both discrete and distributed, capable of producing fields of intensity great enough to cause a detonation.

The limited amount of time available for discussion of this topic did not allow resolution of the many questions remaining. It appears that this particular problem needs to be addressed from a number of directions, and that the expertise represented by many of the workshop participants will be valuable to achieving the goal of defining and minimizing the hazards. Thus, it may be highly beneficial to convene a one or two day meeting of individuals having expertise and experience in all areas related to this particular problem. Such a meeting will help to define and identify those areas where experience and results exist, those areas having gaps in data or confidence, and the likely steps that need to be taken to satisfactorily resolve these issues in a practical manner. These issues are likely to range from the quantitative assessment of danger and/or standards, to operational limitations or procedures that must be imposed on the users of such equipment. Agreement could also be reached on those areas where the greatest amount of progress can be obtained in the least amount of time, and provide the greatest overall benefits.

ADDITIONAL DISCUSSIONS

Concerning paper "Experimental Results Obtained in a Mine on a Network of Leaky Feeders" (printed on pp. 2-11).

QUESTION BY R.L. LAGACE

What is the cost of cable compared with that of conventional cables, and is it mass produced for purchase as opposed to special order?

REPLY (COMMUNICATED) BY P. DEGAUQUE AND R. GABILLARD

The cable is not mass produced but only to special order. Then its price depends on the ordered cable length. For example, for a length greater than 1 km, it costs about 1$/m.

QUESTION BY R.H. SPENCER

Is inner conductor copper/steel or copper?

REPLY (COMMUNICATED) BY P. DEGAUQUE AND R. GABILLARD

Until now, the inner conductor is copper, but in order to increase the tensile strength, we shall use coppered steel wires.

QUESTION BY J.R. WAIT

Could you indicate if the impedance of the transmitting antenna is modified for different locations in the tunnel?

REPLY (COMMUNICATED) BY P. DEGAUQUE AND R. GABILLARD

We have not measured the variation of the impedance of the transmitting antenna. Of course, it will be interesting to make these experiments to compare the variation of the input impedance obtained with a single loop and with a coil wound around bars of ferrite, as used in the actual transmitter.

QUESTION BY N. MACKAY

What is the velocity of propagation of the cable?

REPLY (COMMUNICATED) BY P. DEGAUQUE AND R. GABILLARD

The relative permittivity of the inner dielectric material is $\varepsilon_2 \sim 1.5$, corresponding to a wave velocity of 0.82 c, ($c = 3 \times 10^8$m/s). But the wave velocity inside the cable also depends on the surface impedance of the braid and of the coupling between the cable and the gallery. Therefore, we obtain $v \simeq 0.75$ c.

COMMUNICATED REPLY BY Q.V. DAVIS TO COMMENTS BY N. MACKAY ON PAGE 86

Absence of all cables other than the leaky feeder would obviously be ideal, but the extra effort of providing a satisfactory phase stable telemetry link is not worthwhile. There are two aspects to be considered regarding the effects of reference cables in the tunnel. Firstly, does the actual presence of a cable seriously modify the fields? In our case we think not, because amplitude measurements made without cables in the tunnel are very similar in character to those made in the presence of a passive cable.

Secondly, does the presence of the reference cable affect the actual measurements made, by virtue of its movement with the transmitter or receiver to which it is coupled? We have been very careful to check this matter by taking readings, then moving the cable and re-measuring. We have doubled the cable back on itself, put in loops, run it well past the measurement point and approached from the opposite direction, and so on, without observing change. This is probably so because we have successfully suppressed propagation on the outer screen near the aerial by means of ferrite rings.

The aerial itself was as small as possible consistent with the required sensitivity and was held on a long pole. We also satisfied ourselves that movement of the operator holding the "fishing rod" did not affect the readings, nor did the presence of other people in the tunnel more than a few metres from the aerial.