STUDY OF THE PROPERTIES OF TURNSTILE ANTENNAS BY STANDING WAVE MODELLING OF CURRENT DISTRIBUTION

BY

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CERTIFICATE

This is to certify that this work has been done by me and it has not been submitted elsewhere for the award of any degree or diploma.

Signature of the Student

11.05.87

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ABSTRACT

A computerized study has been carried out to determine the characteristics of a turnstile antenna. The analysis is based on standing wave modelling of the current distribution on the constituent dipole elements of a turnstile array. In such modelling of the current distribution it is assumed that traveling wave of current and voltage along the dipole array will experience imperfect reflection at the terminations so that a standing wave pattern of current will be formed which will be modified over the conventional **sinusoidal** current distribution based on the assumption of open circuited terminals. The resulting effect is that a full wave stands on the dipole arms in a compressed manner without producing a null at the terminal. This gives a finite impedance result for one-wavelength and multiple-wavelength dipoles calculated by the induced e.m.f method. The characteristic impedance of the dipole is calculated from transmission line analogy which follows from biconical approximation of cylindrical dipoles. The antenna-length to radius ratio is found to be an important factor of the impedance of the antenna, because it is a controlling factor for the terminal reflection.

As a preliminary step to turnstile antenna a cross-dipole arrangement has been studied as the basic unit of a turnstile array. When the dipoles in the cross are fed in phase quadrature a circularly polarized field is obtained. If the phase difference is other than 90°, the circular polarization will be distorted and when the dipole currents are in phase a null is observed in

the midway between the cross-dipoles. Unequal strength of currents in the cross-arms will result in an elliptic polarization

Before advancing to a turnstile array a linear dipole array in the form of a pair of batwings has been studied. The broadside field pattern is found to be greatly intensified over that of a single dipole with slight modification in the input impedance. Two such arrays were crossed in phase quadrature to obtain the resulting turnstile characteristics. An approximately circular field pattern with increased gain is observed. When the ends of the dipole elements in the turnstile array are shorted by peripheral rings, the antenna characteristics are found be greatly improved. Improved gain over a very broad-band is the salient feature of the antenna. The computer program developed for studying the characteristics of a turnstile antenna can be readily used for designing a turnstile antenna.

CONTENTS

CHAPTER 1

GENERAL INTRODUCTION

1.1 Historical Review

Turnstile or batwing antenna is named after its construction like a turnstile composed of a pair of batwing like structures. At the present time, super-turnstile antennas [1] invented in the USA are most widely used all over the world for communication in the very high frequency (VHF) ranges.

This antenna was invented by R.W. Masters [1]. It is called a batwing antenna in the USA and "eineschmetterlings - antenne " in Germany in view of this shape. They are popularly used because of very high tolerance to wind velocity due to trussing construction and the most important is their capability of wideband operation.

In 1961, Ports and Rohrer [2] wrote "The antenna which has become the most popular for VHF television broadcasting purposes is the superturnstile or batwing antenna". This antenna came into use with the opening of the commercial television broadcasting in 1948. At the present time over 75 percent of the antenna used for VHF TV-transmitting purposes are super-turnstiles.

Various attempts have been made without much success to explain why this antenna exhibits such excellent wideband characteristics. Some said it was a kind of slot antenna, or a sort of branch or plate antenna, or an antenna excited by a traveling wave.

Berndt [3] tried to evaluate the desired modelling where every description was only qualitative and mere guess work, thereby lacking exactness. Sato and Kawakami [4] and Endo et al [5] tried to explain the characteristics of a turnstile antenna by using simultaneous equations of one degree with 22 unknowns, assuming that the current flowing in each conductor composing this antenna to be sinusoidal. The result was, however, not very satisfactory from the stand point of rigorousness.

Next Kawakami et al [6] rigorously calculated the characteristics of the batwing antenna using the Moment Method proposed by Harrington [7] and conducted some physical experiments. Whenever moment method is proposed Kawakami, Sato and Masters [6] deals with the Garlerkin method where the antenna current is developed using triangle function. The Garlerkin method can save computation time because the coefficient matrix is symmetrical, and the triangle function is widely used because it presents a better performance in calculation and accuracy etc. for an antenna element without sudden change in antenna current. In fact, tremendous involvement of calculation in all the references cited above is attributed to the scarcity of well defined current distribution functions for the straight dipole elements in a turnstile array.

In this thesis we first under took the task of obtaining the current distribution of a straight dipole antenna and applied it in the subsequent calculation of the properties of a turnstile array.

1.2 Present State of Art of the Research Topic

Like other array antennas the analysis of Batwing arrays are not simple. Even a proper electrical design and theoretical base are not available. Actually a unified current distribution for a linear antenna is still to be investigated. We anticipate that the approximate form of current distribution in wires or wire networks is sinusoidal with some deviation caused by the following factors. (i) Effects of size, shape, and proximity of conductors. (ii) Effects of discontinuities, such as sudden changes in the radius of the antenna, or sudden divergence of closely spaced wires. (iii) Effects due to the distribution of generators or to the impressed field. (iv) Effects due to radiation of power. (v) Effects due to the skin effect resistance of conductors. (vi) Effects of proximity between the antenna terminals. The sinusoidal approximation of the current distribution was first discovered by pocklington [8]. Later the sinusoidal distribution proposed by Schelkunoff [9] has some difficulties when the cylindrical dipole antenna is adjusted for integer electrical length. Though many of the tranmitting companies or agencies follow the crude approximation like sinusoidal distribution, investigation for an appropriate distribution of current on the antenna is advancing. Optimum theoretical designs of simple dipole antennas or Batwing or superturnstile antennas are yet to be investigated. But by synthesis procedure some companies or agencies are approaching towards the desired methodology and proper theoretical base.

1.3 Objective of the Research

The objective of this research is to study the characteristics of a turnstile array. Throughout the analysis in the incoming chapters it has been tried to establish a firm basis for the approximate evaluation of the current distribution of a dipole element and the resulting characteristics of a turnstile antenna calculated by arraying such dipole elements.

1.4 Brief Description of the Procedure/Methodology

The preliminary study has been carried out from the concepts derived for straight dipole antennas. In chapter - 2 the properties of straight dipole antennas were studied by standing wave modelling of the current distribution. The standing wave modelling approaches have touched the straight dipole antenna with a brief sequence such that cross-dipole and turnstile arrays have got appreciable consumption for principal analysis and evaluation.

The impedance functions are evaluated by the induced electromotive force(EMF) method. From the very conceptual study of straight dipole antenna, the properties and characteristics of batwing arrays could be readily speculated. Chapter 3 offers the properties of a cross dipole antenna that consists of two simple straight dipoles operated in quadrature and forms the basic unit of a turnstile array. The omnidirectional or horizontal polarization is well defined in this chapter. In chapter - 4 the turnstiling of dipole array is contemplated with theoretical modelling and numerical basing developed by the computer program for rapid inversion of a higher order impedance matrix. The shorted ring is an important element in the batwing arrays that highly increases its gain as well as radiation resistance rather than the arrays without such arrangement. The speciality of this methodology is the standing wave modelling of the current distribution of the dipole elements. This modelling is based on the concept of transmission line analogy applied to a cylindrical dipole with an exception from previous workers that we assumed imperfect reflection of the current and voltage waves at the ends of the dipole so as to comply with the continuity of the conduction current with the displacement current at the dipole ends. This resolves the infinite impedance behaviour of the dipole when its length approaches an integer multiple of a wavelength as is observed with a sinusoidal current distribution.

CHAPTER 2

PROPERTIES OF A STRAIGHT DIPOLE ANTENNA BY STANDING WAVE MODELLING OF THE CURRENT DISTRIBUTION

2.1 Introduction

Among the most common radiators is the dipole which consists of a straight circular cylindrical conductor excited by a voltage derived from a transmission line or directly from a generator. In most cases the exciting source is at the centre of the cylindrical dipole yielding a symmetrical dipole. The most important property of a cylindrical dipole is its current distribution. The current distribution received considerable studies. Schelkunoff [10] applied the biconical transmission line analogy to cylindrical dipoles and observed that the current wave experiences perfect end-reflection producing a sinusoidal standing wave pattern. However, the main drawback of the sinusoidal current distribution is that it predicts an infinite input impedance when the length of the antenna is equal to an integer multiple of a wavelength. A large number of papers [11] – [19] were published on the treatment of this infinite impedance catastrophe with the sinusoidal current distribution. In their analysis, they always assumed perfect end-reflection and tried to obtain finite impedance results at one wavelength or multiple wavelength resonances. However, there is no justification of being so much concerned with the perfect end-reflection. Because Maxwell's equations keep allocation for displacement current at the termination of conduction current for maintaining the continuity of energy flow. Hence a non-zero current can exist at the ends of the dipole. This concept leads to the present investigation on the standing wave modelling of the current distribution assuming imperfect end-reflection.

2.2 Transmission Line Analogy Applied to Dipole Antennas

By transmission line analogy a dipole radiator is assumed to be a typical form of transmission line where the load is the surrounding medium. In conventional analysis of transmission line, we think in terms of a current flowing in the conductors, equal and opposite in the two conductors if measured at any given transverse plane, and a voltage difference existing between the conductors. The transmission line, when it is unflared, the fields are guided by the parallel conductors bounding them and hence no leakage of energy by radiation is assumed to take place. What happens is that energy travels along the line from the source to the load. In a transmission line flared at one end as shown in Fig. 2.1(a), We can notice that the currents at the two parts of the flared end flow in the same direction. Then the energy is no longer confined but is stored or radiated in the surrounding medium. Time-dependence in most practical situations are concerned entirely or at least partially with e^{jwt} variation. With such e^{jwt} variation suppressed the voltage and current waves along a lossless transmission line are solutions of telegraphist's equations and given by

$$V(Z) = Ae^{jK(L-Z)} + Be^{-jK(L-Z)} \qquad \ldots\ldots \qquad 2.2(1)$$

and $$I(Z) = \frac{1}{Z_o}\left[Ae^{jK(L-Z)} - Be^{-jK(L-Z)}\right] \qquad \ldots\ldots \qquad 2.2(2)$$

Where Z_o is the characteristic impedance of the transmission line. A and B are magnitudes of incident and reflected voltages. L is the length of the line, K is the phase constant determined by the distributed inductance and capacitance of the line. In the forms given above the generator is located at the point Z=0 and the load is at the point Z=L.

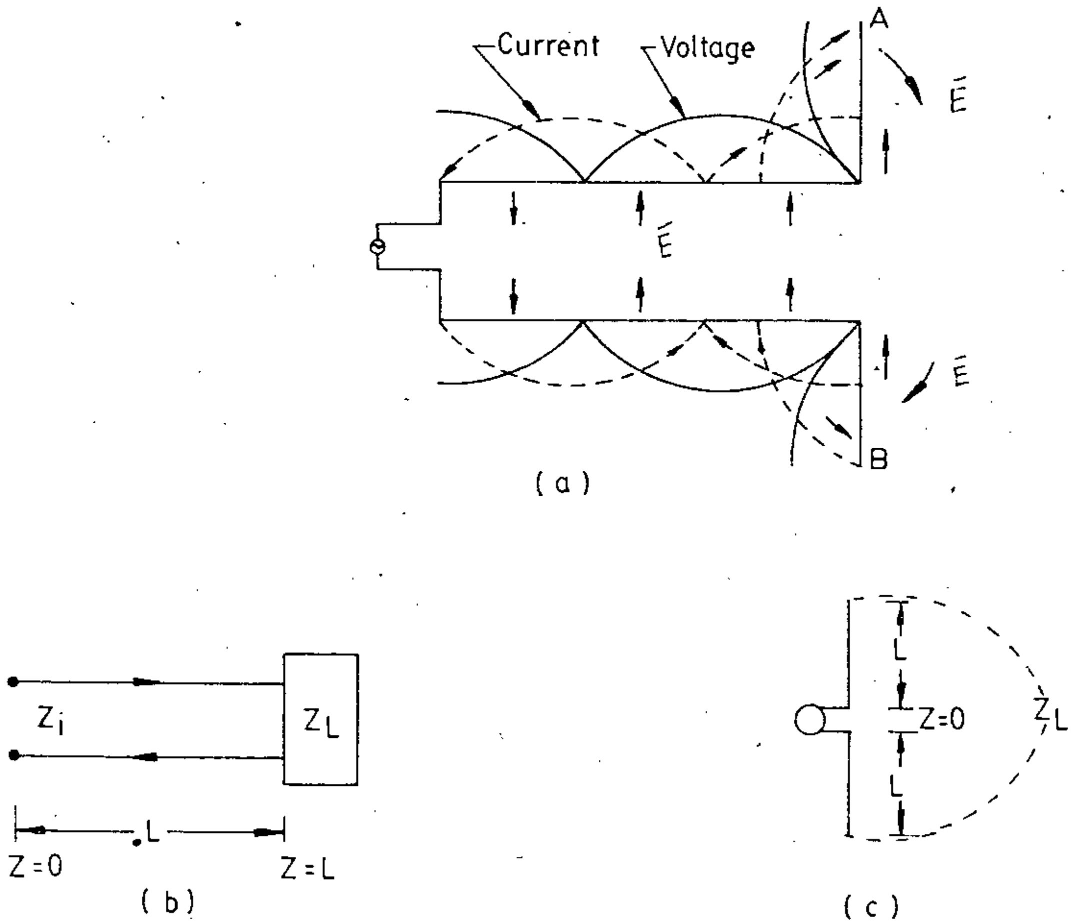

Fig. 2.1(a) Transmission line flared to form an antenna, (b) two wire line and (c) a dipole

In terms of the reflection co-efficient due to discontinuity at the load terminals, equation 2.2(2) can be written as

$$I(Z) = \frac{I_m}{2}\left[(1+\rho)\sin K(L-Z) - j(1-\rho)\cos K(L-Z)\right] \quad \ldots\ldots \quad 2.2(3)$$

Where $I_m = \frac{2jA}{Z_o}$ and ρ = B/A is the reflection coefficient.

In terms of the load impedance Z_L,

$$\rho = \frac{Z_L - Z_o}{Z_L + Z_o} \qquad \text{......... 2.2(4)}$$

For an open end $\rho = 1$ so that equation 2.2(3) takes the form

$$I(Z) = I_m \sin K(L-Z) \qquad \text{......... 2.2(4)}$$

This is the current distribution of an open end transmission line. From transmission line analogy this form of current distribution is applied to a cylindrical dipole antenna. However, in the analysis follows we shall assume $\rho \neq 1$ so that both sine and cosine terms will exist in the current distribution.

2.3 Determination of the End-Reflection of a Cylindrical Dipole

We have seen that unless the transmission line is shorted or opened a cosine distribution invariably exists with the sine distribution in the standing wave pattern of the current distribution. This modelling signifies itself an imperfect end-reflection as a result of which the reflection Co-efficient can be assumed other than unity. The methodology of the determination of the reflection Co-efficient is demontrated as follows. We have the electric field $\bar{E}$ in terms of scalar and vector potentials as

$$\bar{E} = -\nabla\phi \quad -jw\bar{A} \qquad \text{............ 2.3(1)}$$

Where, $\bar{E}$, electric field

Φ , scalar electric potential

$\bar{A}$, vector magnetic potential

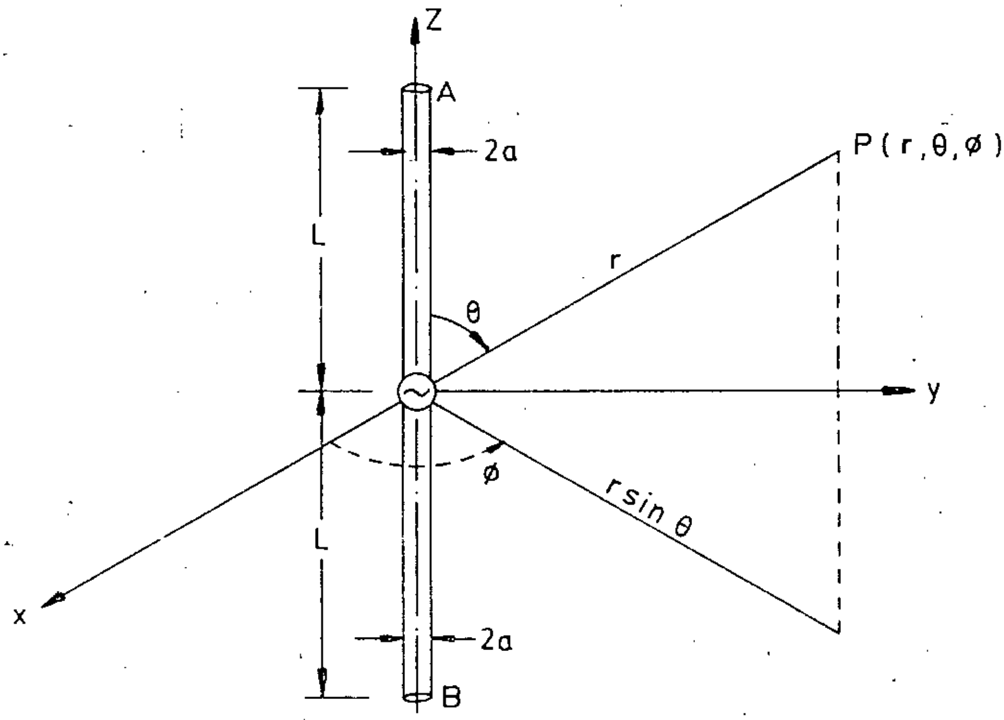

Fig. 2.2 Co-ordinate system of a cylindrical dipole antenna

For an isotropic homogeneous medium the two potentials are related by Lorentz's gauge,

$$\nabla \cdot \bar{A} = -j\omega\mu\varepsilon\Phi \qquad \ldots\ldots\ 2.3(2)$$

For a specified current distribution $\bar{i}$ the vector magnetic potential is given by

$$\bar{A} = \mu \int_v \bar{i} \frac{e^{-jkr}}{4\pi r} dv \qquad \ldots\ldots\ldots \quad 2.3(3)$$

Where ε is electric permittivity of the medium

μ is magnetic permeability of the medium

$k = w\sqrt{\mu\varepsilon}$

r is the distance between the source point and the observation point. In rectangular Co-ordinates

$$r = \sqrt{(x-x_o)^2 + (y-y_o)^2 + (Z-Z_o)^2} \qquad \ldots\ldots\ldots \quad 2.3(4)$$

Here (x_o, y_o, Z_o) is the source point and (x,y,z) is the observation point. Assuming that the current is flowing along the axis of the antenna and referring to the Co-ordinate system in Fig. 2.2 the Z-directed vector magnetic potential is given by

$$A_Z = \frac{\mu}{4\pi} \int_{-L}^{L} \frac{I(Z)e^{-jK\sqrt{x^2 + y^2 + (Z-Z_o)^2}}}{\sqrt{x^2 + y^2 + (Z - Z_o)^2}} dZ_o \quad \ldots \quad 2.3(5)$$

Where L is the half-length of the dipole antenna and I(Z) is the current distribution. With the help of equation 2.3(1) and 2.3(2) the Z-directed electric field can be written as

$$E_Z = \frac{1}{jw\mu\varepsilon}\left[\frac{\partial^2 A_Z}{\partial z^2} + k^2 A_Z\right] \qquad \ldots\ldots\ldots \quad 2.3(6)$$

Substituting the current distribution defined in equation 2.2(3) into the above equation, we have the solution for E_Z,

$$E_Z = \frac{I_m}{j8\pi\omega\varepsilon}\left[k(1+\rho)F_1(r,Z)+j(1-\rho)\;F_2(r\;,z)\right] \qquad \ldots\ldots \quad 2.3(7)$$

Where,

$$F_1(r,\;Z)=\frac{e^{-jkr_1}}{r_1}+\frac{e^{-jkr_2}}{r_2}-2\cos\;kL\;\frac{e^{-jkr_o}}{r_o}$$

$$F_2(r,Z)=(L-Z)\left(\frac{jk}{r_1^{\,2}}+\frac{1}{r_1^{\,3}}\right)e^{-jkr_1}+(Z+L)\left(\frac{jk}{r_2^{\,2}}+\frac{1}{r_2^{\,3}}\right)e^{-jkr_2}$$
$$-2k\sin\;kL\;\frac{e^{-jkr_o}}{r_o}$$

$$r = \sqrt{x^2+y^2}$$
$$r_1 = \sqrt{r^2+(Z-L)^2}$$
$$r_2 = \sqrt{r^2+(Z+L)^2}$$
$$r_o = \sqrt{r^2+\;Z^2}$$

The axial electric field can be readily used to obtain the potential difference between the two ends of the dipole as

$$V_{AB} = V(L)\;-V(-L)\;=\;-\int_{-L}^{L} E_Z\;\;dZ \qquad \ldots\ldots \quad 2.3(8)$$

Substituting 2.3(7) into 2.3(8) we get,

$$V_{AB} = -\frac{I_m}{j4\pi\omega\varepsilon}\left[K(1+\rho)I_1'\;+j(1-\rho)\;I_2'\right] \qquad \ldots\ldots \quad 2.3(9)$$

Where I_1' and I_2' are the integrals

$$I_1' = \int_o^L F_1(a,Z)\, dZ$$

$$I_2' = \int_o^L F_2(a,Z)\, dZ \qquad \ldots\ldots\ldots \qquad 2.3(10)$$

The terminal loading can be defined as

$$Z_L = \frac{V_{AB}}{I(L)} \qquad \ldots\ldots\ldots \qquad 2.3(11)$$

where I(L) is the tip current of the dipole

Again $Z_L = Z_o \frac{1+\rho}{1-\rho}$ .... 2.3(12)

The characteristic impedance Z_0 of a cylindrical dipole is given by [20]

$$Z_o = 120(\ln \frac{2L}{a} - 1) \qquad \ldots\ldots\ldots \qquad 2.3(13)$$

From 2.3 (9) , 2.3(11) and 2.3(12) we have,

$$Z_L = \frac{-j60\, I_2'}{K + \frac{60K}{Z_o} I_1'} \qquad \ldots\ldots\ldots \qquad 2.3(14)$$

and $$\rho = \frac{jI_2' + K\left(I_1' + \frac{Z_o}{60}\right)}{jI_2' - K\left(\frac{Z_o}{60} + I_1'\right)} \qquad \ldots\ldots\ldots \qquad 2.3(15)$$

At very low frequency $KL \to 0$, so that $\rho = 1$. That means at low frequencies $I(Z) \simeq I_m \sin K(L-Z)$. At high frequencies where $KL \to \infty$, $\rho = -1$ i.e. $I(Z) = -j\, I_m \cos K(L-Z)$. Thus both sine

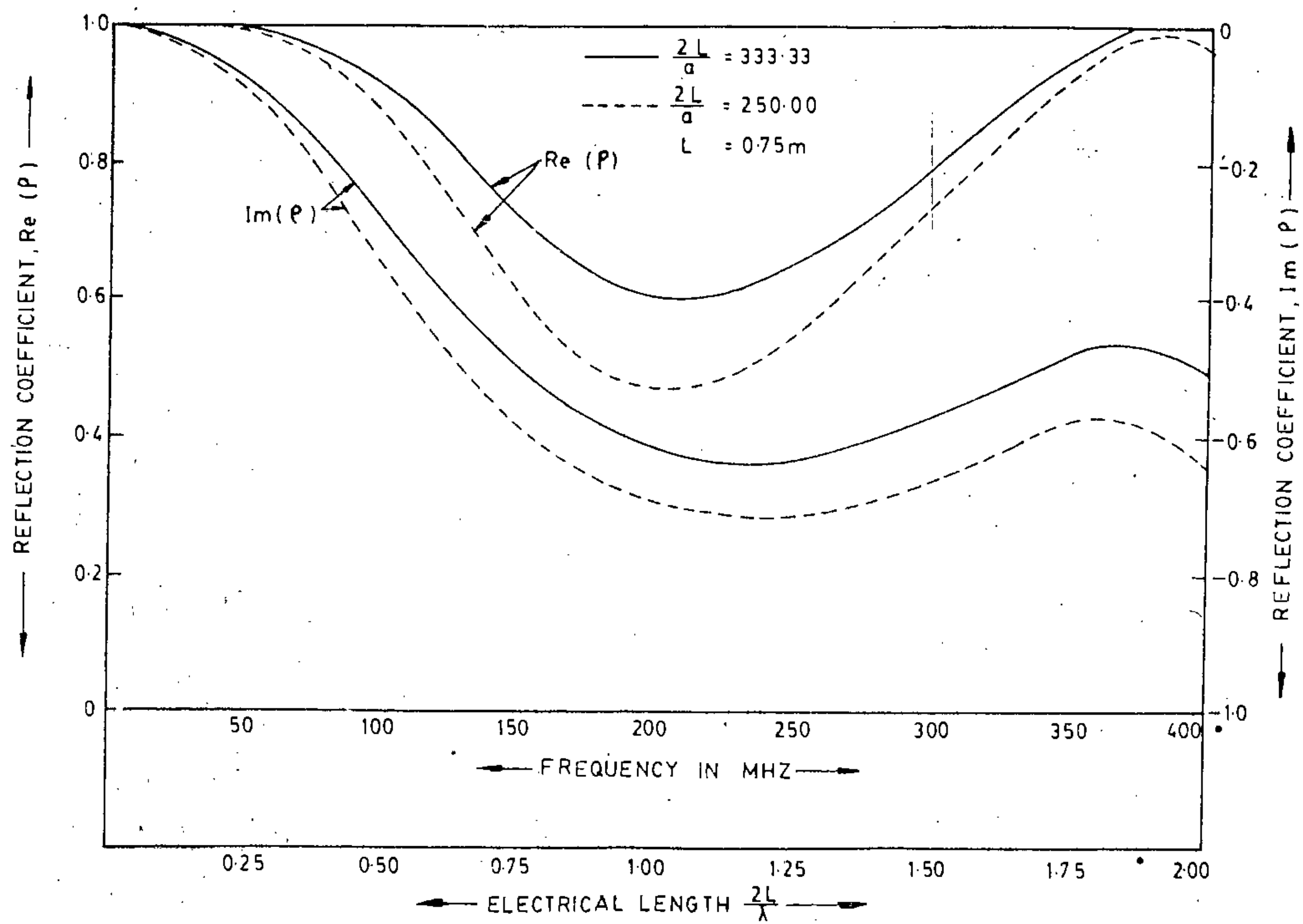

Fig. 2.3 End reflection coefficient versus frequency characteristics of a cylindrical dipole antenna

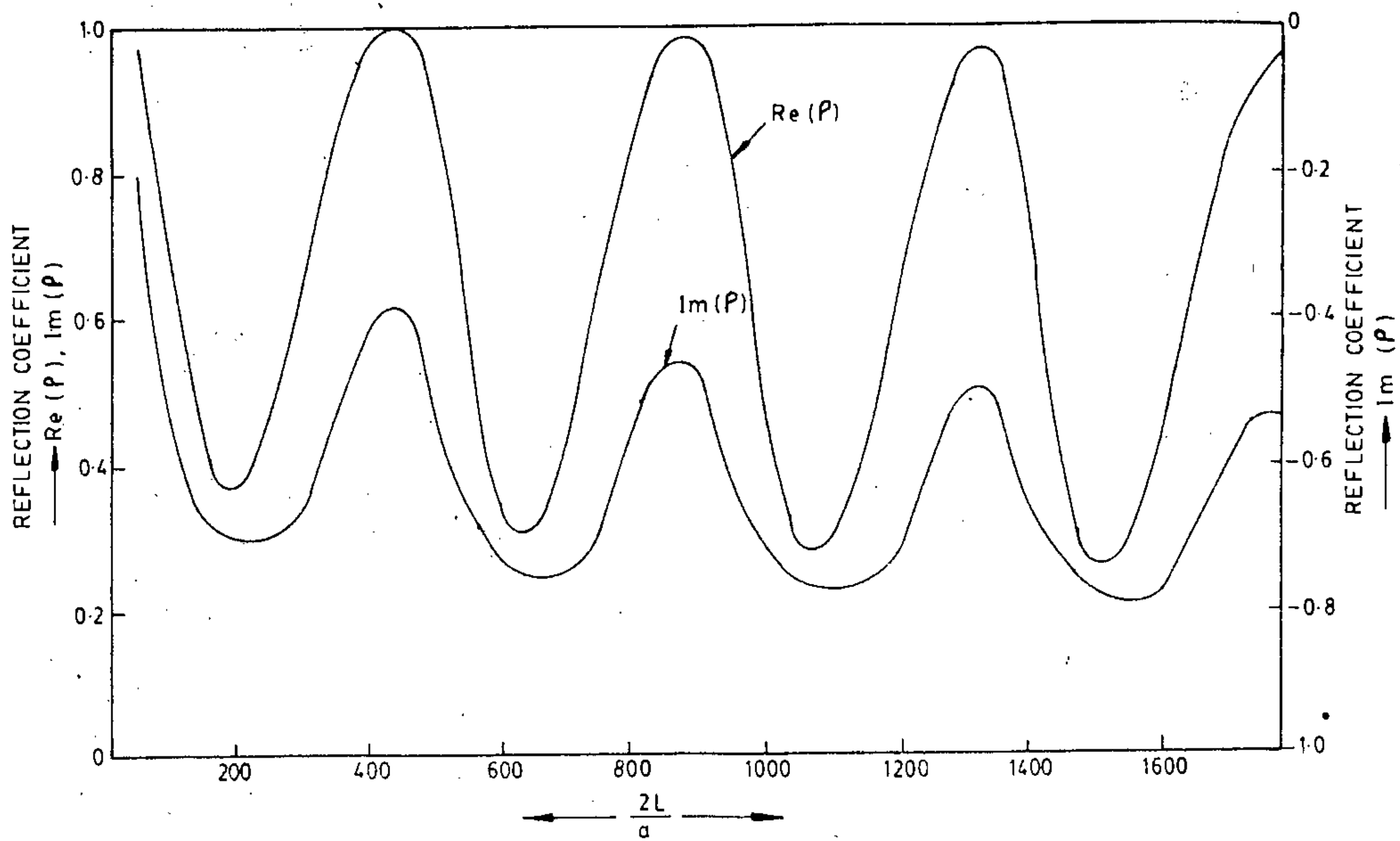

Fig. 2.4 End reflection coefficient versus length to radius ratio($\frac{2L}{a}$) characteristic of a cylindrical dipole antenna for f= 300 MHZ

and cosine distributions are simultaneously involved in a realistic current distribution.

The reflection Co-efficient given by equation 2.3(15) has been numerically calculated performing the integration by Gauss Legrendre method [21] . Fig. 2.3 shows the behavior of the end reflection with respect to the operating frequency. The electric length $\frac{2L}{\lambda}$ of the dipole is an important factor so that the end reflection has a gradual change with the variation of $2L/\lambda$ It is observed that for $\frac{2L}{\lambda}=0.5$, the end reflection tends toward unity indicating a sinusoidal current distribution. The wide variation of end reflection occurs in between $\frac{2L}{\lambda}$ = 0.5 to 1.75.

So far the length to radius ratio of the antenna is concerned the reflection Coefficient has a wide variation but of perodic in nature for both real and immaginary parts of it. The pictorial variation of end reflection versus the dipole length to radius ratio ($\frac{2L}{a}$) is plotted in Fig. 2.4.

2.4 Current Distribution of a Cylindrical Dipole by Standing Wave Modelling

In the case of a short conductor which is lossless when a current is made to flow through it, the amplitude of the wave does not change and what is observed is only a change of phase. If the length of the wire is long, on the other hand the amplitude of the wave as it travels along the conductor would decrease due to two reasons: (1) the radiative loss along the length of the

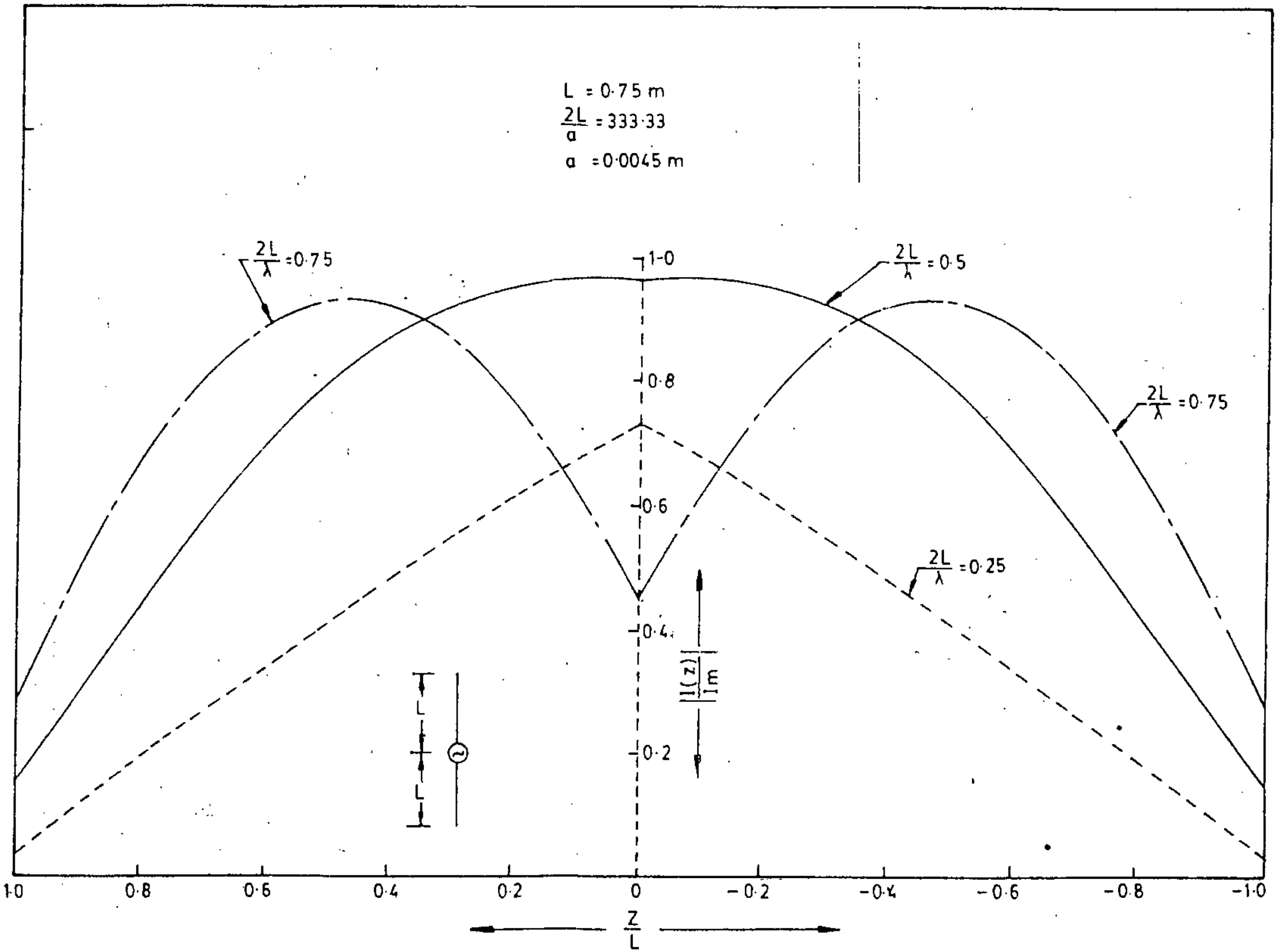

Fig. 2.5 Standing wave distribution of current on a cylimdrical dipole antenna for 2L/λ =.25, .5 and 0.75

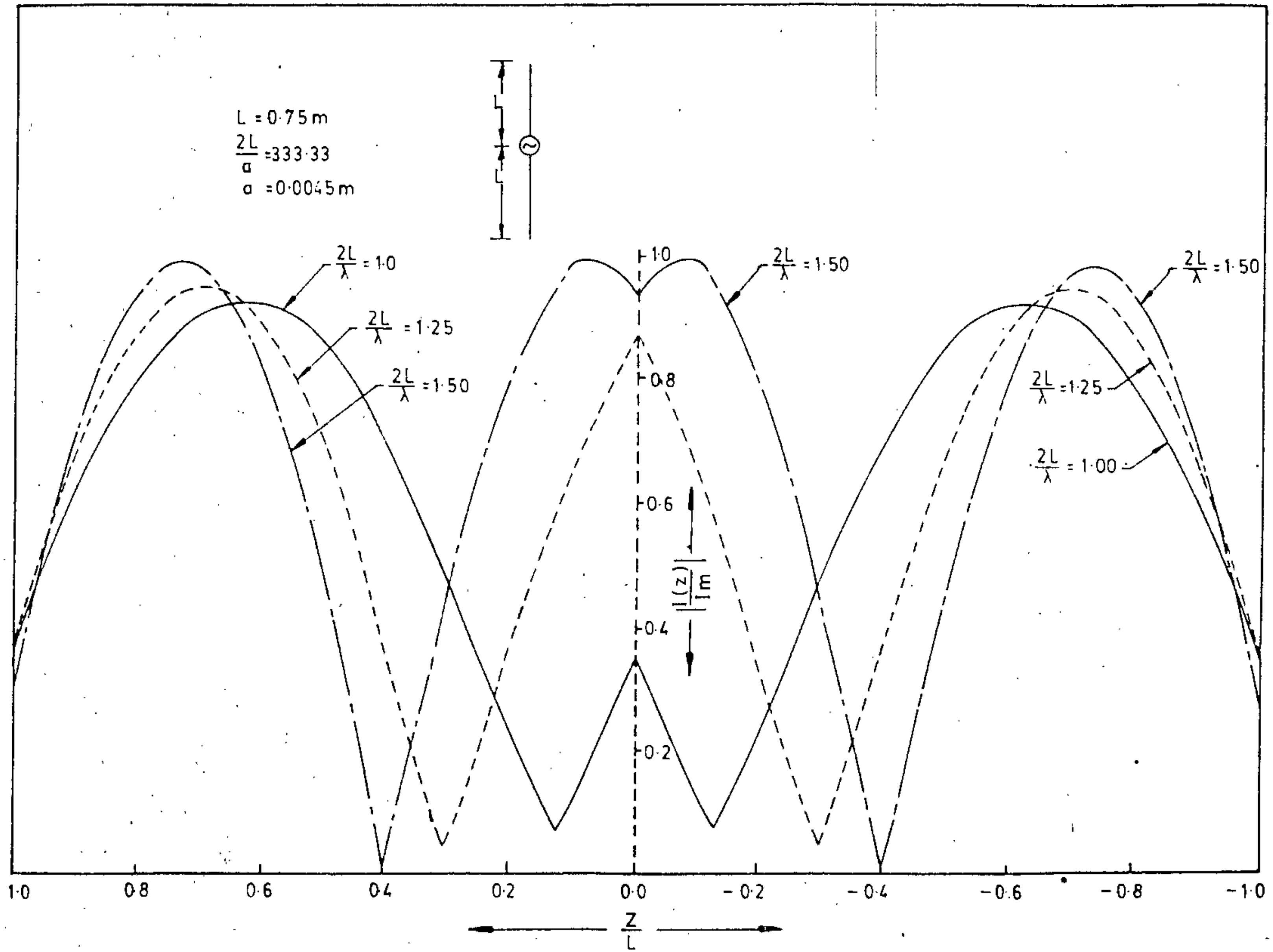

Fig. 2.6 Standing wave distribution of current on a cylimdrical dipole antenna for $2L/\lambda = 1.0, 1.25, 1.5$

conductor and (2) the dissipative loss due to skin effect which also increases as the length of the conductor increases. Current distribution is the most important property of a cylindrical dipole. Using the reflection Coefficient derived in section 2.3, the current distribution along the surface of the dipole is calculated by equation 2.2(3). The numerical results are plotted in Figs. 2.5 - 6. The results suggest that for short antennas i.e. $2L/\lambda \leqslant .25$ a triangular current distribution is satisfactory. This is also predicable from a sinusoidal current distribution. In the present case the salient feature is that the current is non-vanishing at the tips of the dipole due to imperfect end reflection. As in Fig. 2.6 for $\frac{2L}{\lambda} = 1$ a full wave stands on the antenna in a compressed manner.

2.5 The Field Pattern of a Straight Dipole Antenna

The radiation pattern of an antenna is a graphical representation of the radiation of the antenna as a function of direction. When radiation is expressed as field strength, the radiation pattern is a field strength pattern. The coordinate system generally used in the specification of antenna field pattern is the spherical co-ordinate system (r, θ, ϕ). The shape of the radiation pattern is independent of r, as long as r is chosen sufficiently large. When this is true, the magnitude of the field strength in any direction varies inversely with r.

The distribution of current along the two halves of the dipole is given by

$$I(Z) = \frac{I_m}{2}\Big[(1+\rho)\sin K(L-Z) - j(1-\rho)\cos K(L-Z)\Big]; \quad (o \leqslant Z \leqslant L)$$

$$= \frac{I_m}{2}\Big[(1+\rho)\sin K(L+Z) - j(1-\rho)\cos K(L+Z)\Big]; \quad (-L \leqslant Z \leqslant 0)$$

...... 2.5(1)

For the far field consideration,

$$E_\theta = \eta H_\varphi = \int_{-L}^{L} dE_\theta = \frac{j30ke^{-jkr}}{r}\sin\theta \int_{-L}^{L} I(Z)e^{jkZ\cos\theta}\,dZ$$

..... 2.5(2)

Using the current distribution described in 2.5(1), the field pattern of the cylindrical dipole antenna is given by

$$E_\theta = \frac{j60\ I_m}{r}\,e^{-jkr}\,P(\theta) \qquad \text{........ 2.5(3)}$$

$$\text{Where } P(\theta) = \frac{1+\rho}{2}\cdot\frac{\cos(KL\cos\theta) - \cos KL}{\sin\theta} - j\,\frac{1-\rho}{2}\cdot\frac{\sin KL - \cos\theta\,\sin(KL\cos\theta)}{\sin\theta}$$

.... 2.5(4)

θ is the angle between the axis of the dipole and the radius vector to the point where the field strength is measured. The absolute value of the field pattern factor $|P(\theta)|$ will give the actual magnitude of the pattern.

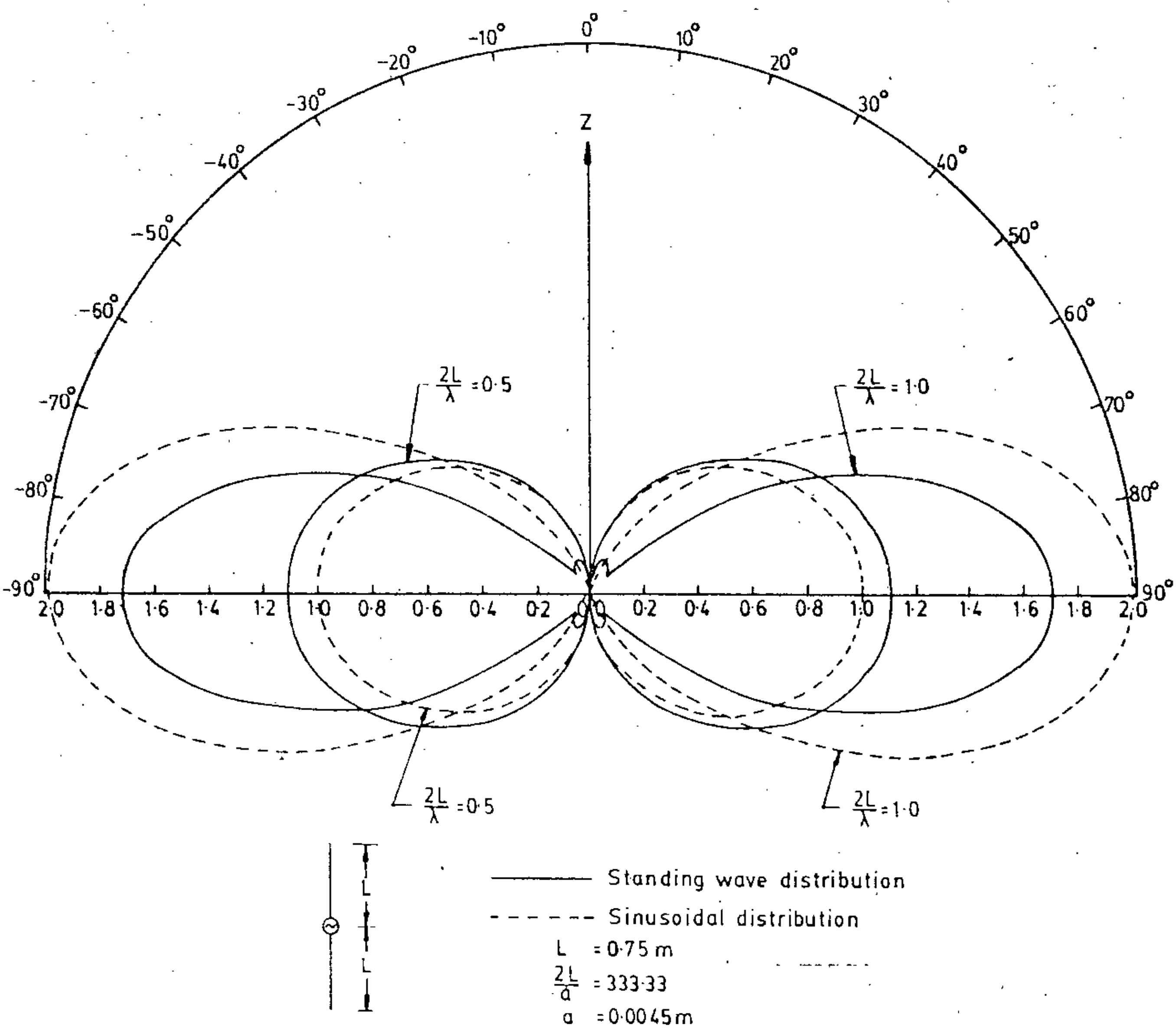

Fig. 2.7(a) Electric field pattern in the E-plane for a straight cylindrical dipole for $2L/\lambda$ = 0.5 and 1.0

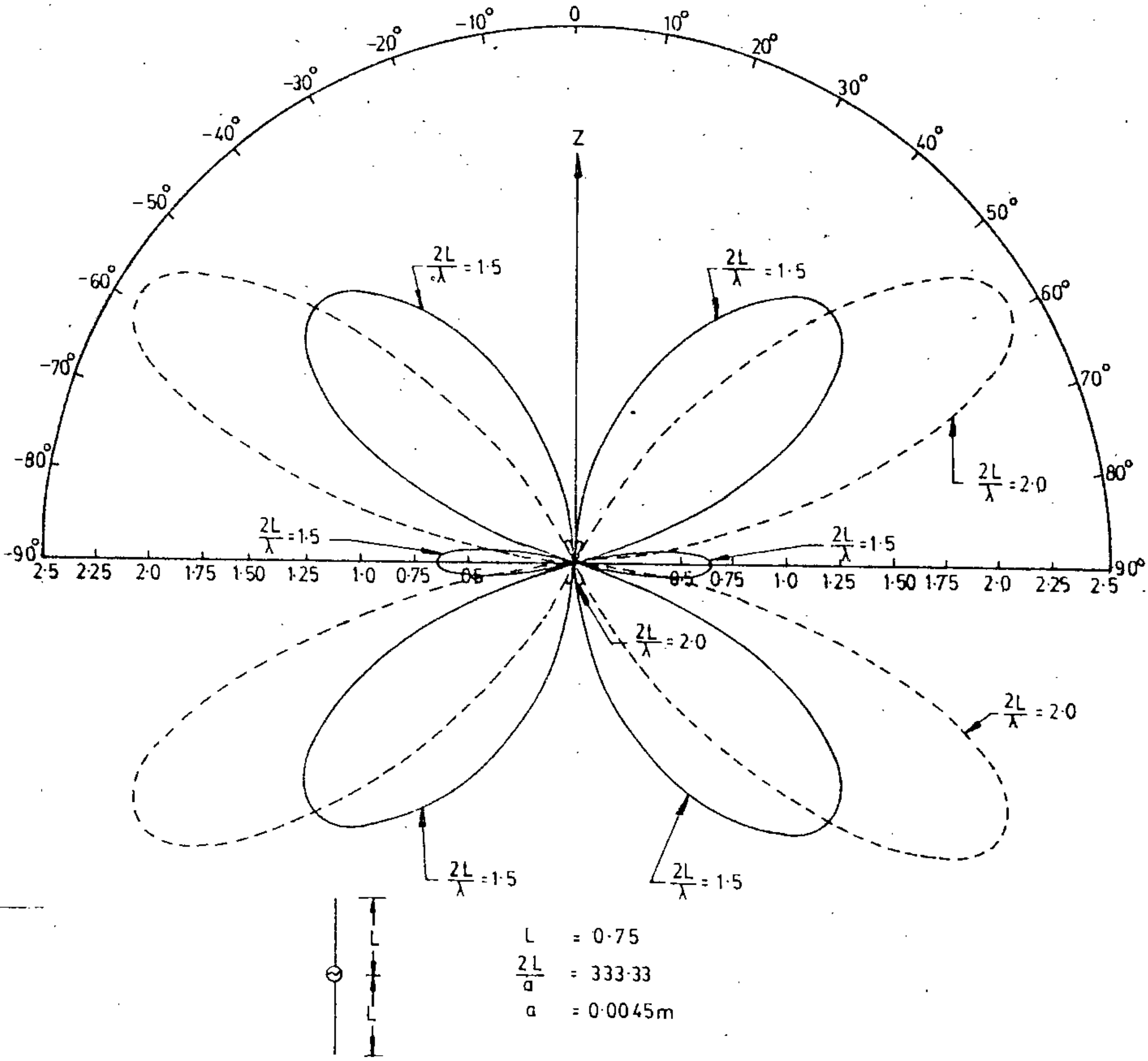

Fig. 2.7(b) Electric field pattern in the E-plane for a staright cylindrical dipole considering the imperfect reflection at the end

Figs. 2.7(a)-(b) describe the plot of the magnitude of the field pattern of the cylindrical dipole using standing wave current distribution. For comparison, field patterns for sinusoidal current distribution are also shown with dotted lines. The field pattern is symmetrical about the antenna axis. For sinusoidal distribution of current the field pattern has two leaves upto $\frac{2L}{\lambda} = 1$ i.e. upto where the antenna length becomes equal to one wave length. When the length of the antenna exceeds this limit, the number of leaves in the field pattern increases. For $m = \frac{2L}{\lambda} = 1, 2, 3$ .... etc, the number of leaves becomes 2m and for $m = \frac{2L}{\lambda} = 1.5, 2.5, 3.5$ ..etc. the number of leaves will be 4m. For standing wave modelling of current distribution, there is the tendency of creating leaves with the same sense except for one-wave-length dipole where the number of leaves are six. Among six leaves, four leaves are very weak. The general tendency is that when $m = \frac{2L}{\lambda} = 0.5,\ 1.5,\ 2.0$ .... etc. the number of leaves becomes 4 m.

2.6 The Impedance of a Straight Dipole Antenna

The antenna input impedance presented to the feed line is consequently an important parameter to know in order to design efficient coupling networks that will give maximum power transfer or in abstracting the maximum amount of receiving energy available from the antenna. In electromagnetic wave propagation the notion of impedance has been extended to include the ratio of transverse components of electric and magnetic fields. The antenna is the device that integrates the electric and magnetic fields to produce the voltages and currents required to actuate electrical devices. In this transition from fields to circuits it is necessary to examine critically the notion of impedance so that the meaningful expression involving both field and circuit quantities can be derived. The input impedance of a long cylindrical dipole may be calculated by the induced emf method. Applied to circuits this method consists of assuming a certain current flow, computing the voltages induced by the assumed current in the various parts of the circuit, and then summing these back-voltages around the circuit to determine the voltage that must be impressed at the input terminals to produce the assumed current. The induced emf method was originally introduced [22] for the computation of radiation resistance. But since it deals with the radially near fields of the antenna, it can also be used to calculate mutual impedance between antennas and the reactance of a single antenna. Using the standing wave model of the current distribution, the dipole input impedance by definition is given by,

$$Z_i(0) = -\frac{1}{I(0).\ I(0)^*} \int_{-L}^{L} E_Z(a, Z)\ I^*(\,|Z|\,)\ dZ$$

.... 2.6(1)

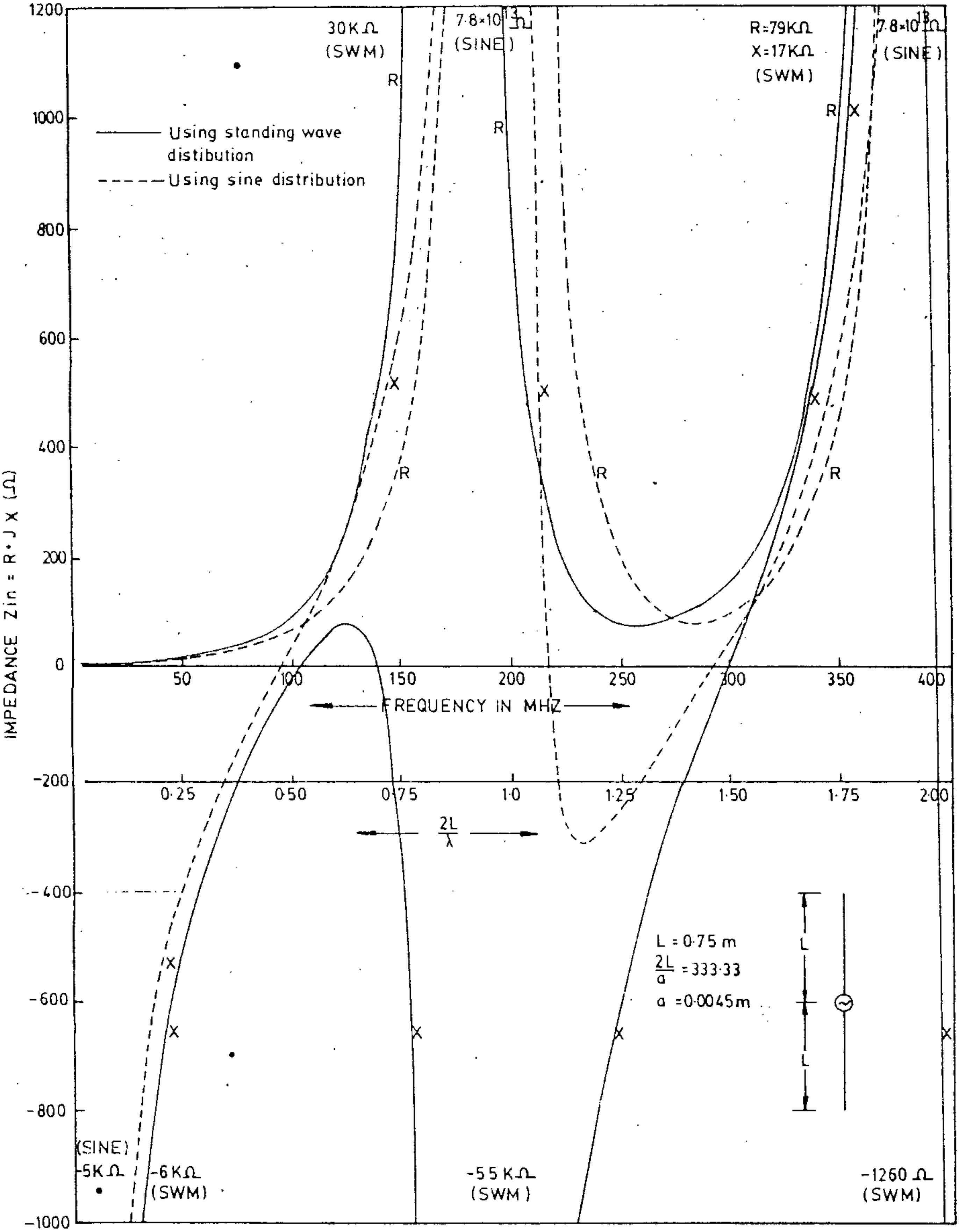

Fig. 2.8(a) Impedance versus frequency characteristics of a cylindrical dipole antenna for 2L/a = 333.33(solid lines for standing wave model of current distribution and dotted line for sine distribution)

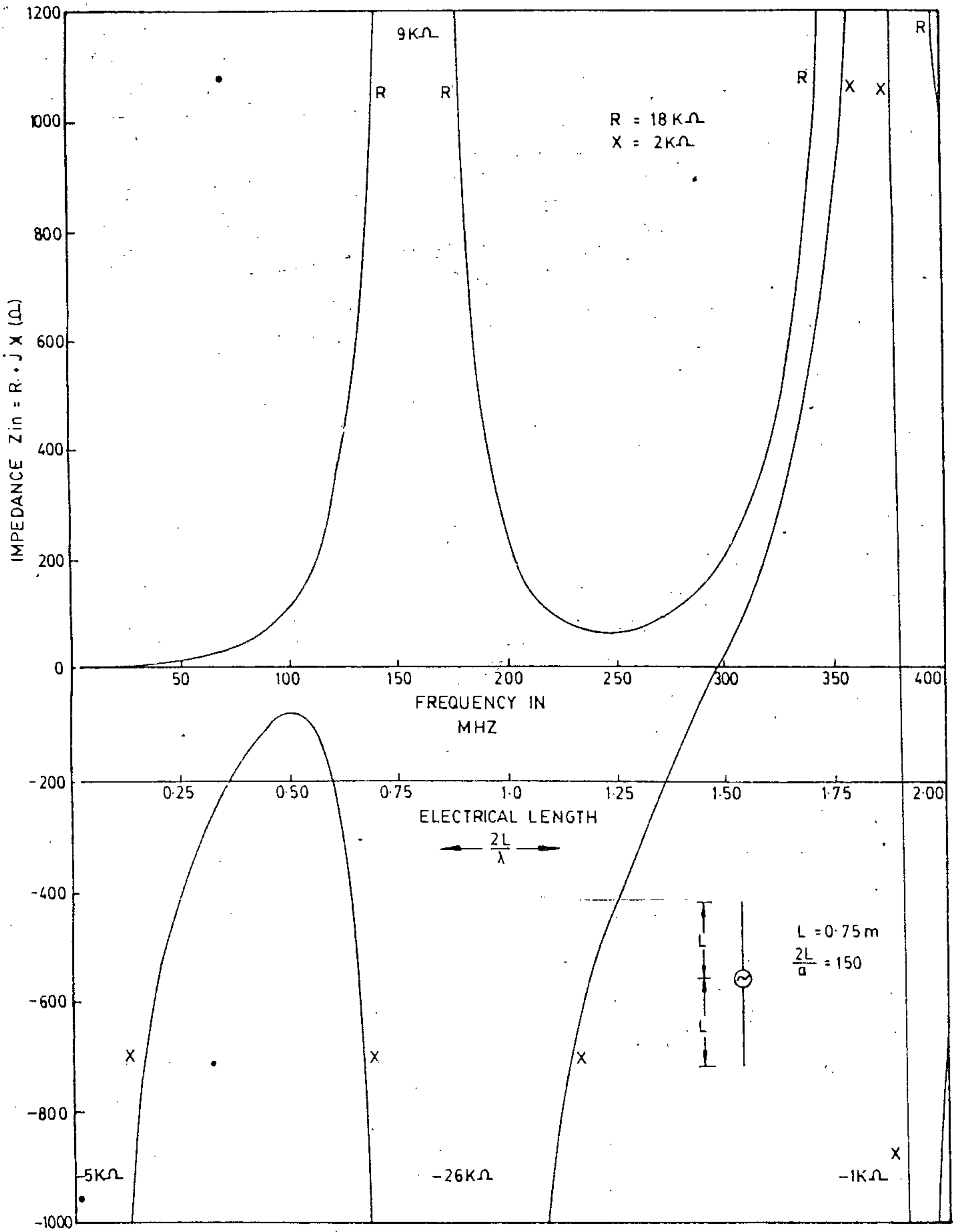

Fig. 2.8(b) Impedance versus frequency characteristics of a cylindrical antenna for 2L/a = 150, using standing wave model of current distribution

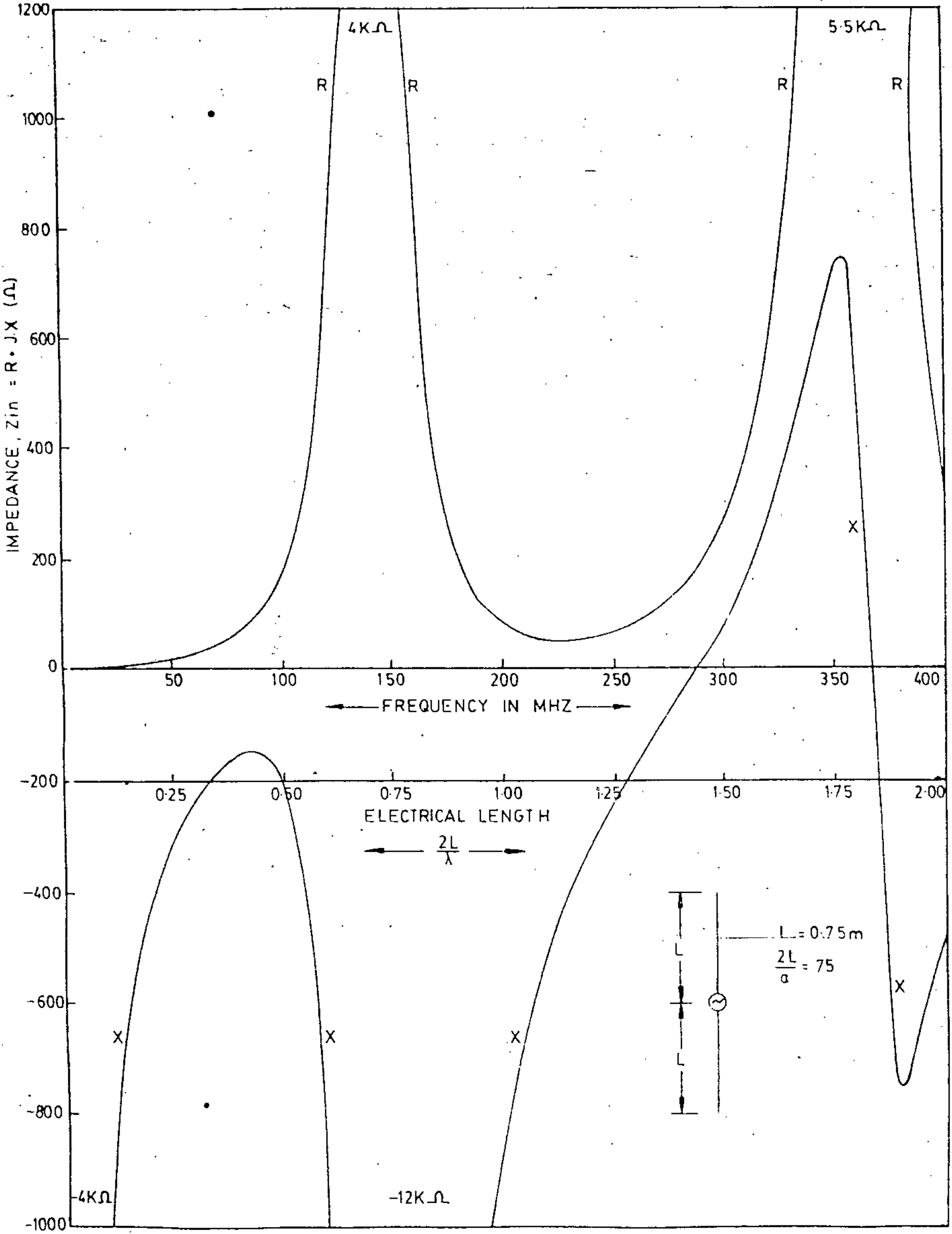

Fig.2.8(c) Impedance versus frequency characteristics of a cylindrical antenna for 2L/a = 75, using standing wave model of current distribution

Substituting for E_Z the axial electric field derived in 2.3(7), the dipole input impedance is given by

$$Z_{in} = \frac{j60}{KC(0).C(0)^*} \int_0^L F(a,Z)\, C^*(Z)\, dz = R + jx \quad \ldots\ldots \; 2.6(2)$$

Where $C(0) = -j(1-\rho)\cos KL + (1+\rho)\sin KL$

$C(Z) = -j(1-\rho)\cos K(L-Z) + (1+\rho)\sin K(L-Z)$

$F(a,Z) = K(1+\rho)F_1(a, Z) + j(1-\rho)F_2(a,Z)$

Figs. 2.8(a), (b) and (c) show the impedance results calculated by 2.6(1). When the length of the antenna becomes one wave-length or a multiple of wave-length the input impedance becomes infinite for sinusoidal distribution of current. But the standing wave current saves the problem from such catastrophic results giving finite values of impedance at these resonaces of the antenna. The resistive part of the input impedance has contributions from all physical mechanisms that gives rise to a loss of energy from the antenna. In this case it is the radiation resistance of the dipole. The antenna length-to-radius ratio, 2L/a plays an important role in determining the impedance of the dipole as can be observed from comparison of Figs. 2.8(a), (b) and (c).

2.7 Discussion

The description that has been presented in the previous articles are the properties of linear antennas having considerable dimension. We have concerned about the straight cylindrical dipole. The investigations were carried out by choosing the standing wave modelling and sinusoidal distribution of current. The catastrophic result obtained from sinusoidal distribution of current has been greatly eliminated by using the standing wave model of the current distribution. Otherwise the results have qualitative agreement with those predictable by a sinusoidal current distribution.

CHAPTER 3

PROPERTIES OF A CROSS - DIPOLE ANTENNA

3.1 Introduction

Cross Dipole is an elementary arrangement radiating continuously in all directions and there is no direction in which the radiation would equal zero. If an omnidirectional pattern is desired at short wave, two dipoles at right angles to each other are used and they are fed with quadrature phase currents. This results in an approximately omnidirectional pattern in the horizontal plane. Among the non-directional multi-unit antennas, mention should be made of the turnstile antenna utilized in telecasting stations. At VHF such turnstiling of crossed dipoles is used to get omnidirectional pattern.

First of all, the antennas should radiate uniformly in all directions in the horizontal plane since TV-towers are in the center of the region which they serve. In the horizontal plane, the directivity should be sufficiently large in order that the radiation should be concentrated in the horizontal plane. Moreover, the antenna should radiate electromagnetic energy with a horizontal E-field polarization due to the fact that the sources of interference on ultrashort waves are industrial installations of all kinds which interfere with an electromagnetic field having mainly vertical polarization.

The dipoles in a cross-dipole arrangement are excited by seperate feeders connected in parallel. Moreover, one feeder is a quarter wave-length longer than the other ensuring a 90^{o} phase difference of the currents.

3.2 Field Pattern of a Cross Dipole

A cross-dipole antenna cosists of two radiators mounted at right angles to each other in the same horizontal plane. The radiators carry equal currents but in phase quadrature. This kind of phasing could be effected by having an extra quarter wave feeder in the connection to one antenna known as a quarter wave loop. In any direction 'θ' from the axis of a small dipole at a distance point 'P' the field magnitude is propertional to $\sin\theta$ and that due to the quadrature dipole is proportional to $\cos\theta$. The resultant is therefore a constant independent of θ. That is we get an all round horizontally polarized wave from a small cross-diple. Moreover, the two dipole fields being at quadrature, there will be no mutual coupling between them.

Let us assume the currents through the dipoles are of standing wave distribution having maximum stengths I_{m_1} and I_{m_2} and a phase difference α. Then by considering the first dipole as reference the current distributions of the two dipoles can be given as

$$I_1(Z) = \frac{I_{m1}}{2}\left[(1+\rho)\sin K(L-Z) - j(1-\rho)\cos K(L-Z)\right]$$

and $$I_2(Z) = \frac{I_{m2}}{2}\left[(1+\rho)\sin K(L-Z) - j(1-\rho)\cos K(L-Z)\right] . e^{-j\alpha} \qquad \ldots\ldots.3.2(1)$$

For the first dipole the electric field in the far zone at a point $P(r, \theta, \phi)$ is given by,

$$E_{\theta_1} = \frac{j \eta I_{m1}}{2 \pi r} e^{-jkr} \quad P_1(\theta) \qquad \ldots\ldots\ldots \quad 3.2(2)$$

Where the pattern factor $P_1(\theta)$ has been derived in 2.5(3) may be repeated here,

$$P_1(\theta) = \frac{1+\rho}{2} \frac{\text{Cos}(KL \text{ Cos}\theta) - \text{Cos } KL}{\text{Sin}\theta} - j \frac{1-\rho}{2} \frac{\text{Sin}KL - \text{Cos}\theta \text{ Sin}(KL \text{Cos}\theta)}{\text{Sin}\theta}$$

$$\ldots\ldots\ldots \quad 3.2(3)$$

similarly the electric field of the second dipole in the far zone at $P(r, \theta, \phi)$ is given by ,

$$E_{\theta_2} = \frac{j \eta I_{m2}}{2 \pi r} e^{-jkr} P_2(\theta) e^{-j\alpha} \qquad \ldots\ldots\ldots \quad 3.2(4)$$

Where

$$P_2(\theta) = \frac{1+\rho}{2} \frac{\text{Cos}(KL\text{Sin}\theta) - \text{Cos}KL}{\text{Cos}\theta} - j \frac{1-\rho}{2} \frac{\text{Sin}KL - \text{Sin}\theta \text{ Sin}(KL \text{ Sin}\theta)}{\text{Cos}\theta}$$

$$\ldots\ldots\ldots \quad 3.2(5)$$

The total electric field observed at the point $P(r, \theta, \phi)$ is given by ,

$$E_\theta = E_{\theta 1} + E_{\theta 2} = \frac{j \eta e^{-jkr}}{2 \pi r} \left[I_{m1} P_1(\theta) + I_{m2} P_2(\theta) e^{-j\alpha} \right] \ldots 3.2(6)$$

Let us define $\frac{I_{m1}}{I_{m2}}$ or $\frac{I_{m2}}{I_{m1}} = r$ such that $r \geqslant 1$. Then

$$P(\theta) = P_1(\theta) + rP_2(\theta) e^{-j\alpha}, \quad I_{m1} \leqslant I_{m2}$$

and $P(\theta) = rP_1(\theta) + P_2(\theta) e^{-j\alpha}$, $I_{m1} \geqslant I_{m2}$ ..... 3.2(7)

Since the half-wave length dipole pattern is slightly sharper than that for a small dipole the θ- plane pattern of the cross dipole with half-wave length element is not quite circular but departs from a circle by about $\pm$ 5 percent. Starting from the inphase current distribution, the achievement of rotating and a horizontal polarization, uniform in all direction field patterns are approached by a fractional step of quadrature phase. Fig. 3.1 - 3.3 indicate the field pattern of a cross dipole fed in phase-quadrature under matched condition. For a comprehensive study we have considered the field pattern for various electrical lengths of the dipole. The results arising from the feeding of cross dipole in non-quadrature phase and with non-uniform distribution of current may be cited. Fig. 3.3 describes the field pattern of the same when fed in 90° phase difference under mismatched condition i.e. $I_{m_1} \neq I_{m_2}$. In both the cases when $I_{m1} > I_{m2}$ or $I_{m2} > I_{m1}$, the polarization is elliptic rather than circular. Fig. 3.4 shows the field distribution when the dipoles are fed in phase. Null is observed at $\theta = 135^{\circ}$ and at $\theta = -45^{\circ}$ measured from the axis of the reference dipole. With the increase of phase difference this null portion is filled and finally for a phase difference of 90° a symmetrical field pattern is obtained . Thus the field term is controllable by adjusting magnitude and phases of the currents in the two dipoles forming the cross dipole.

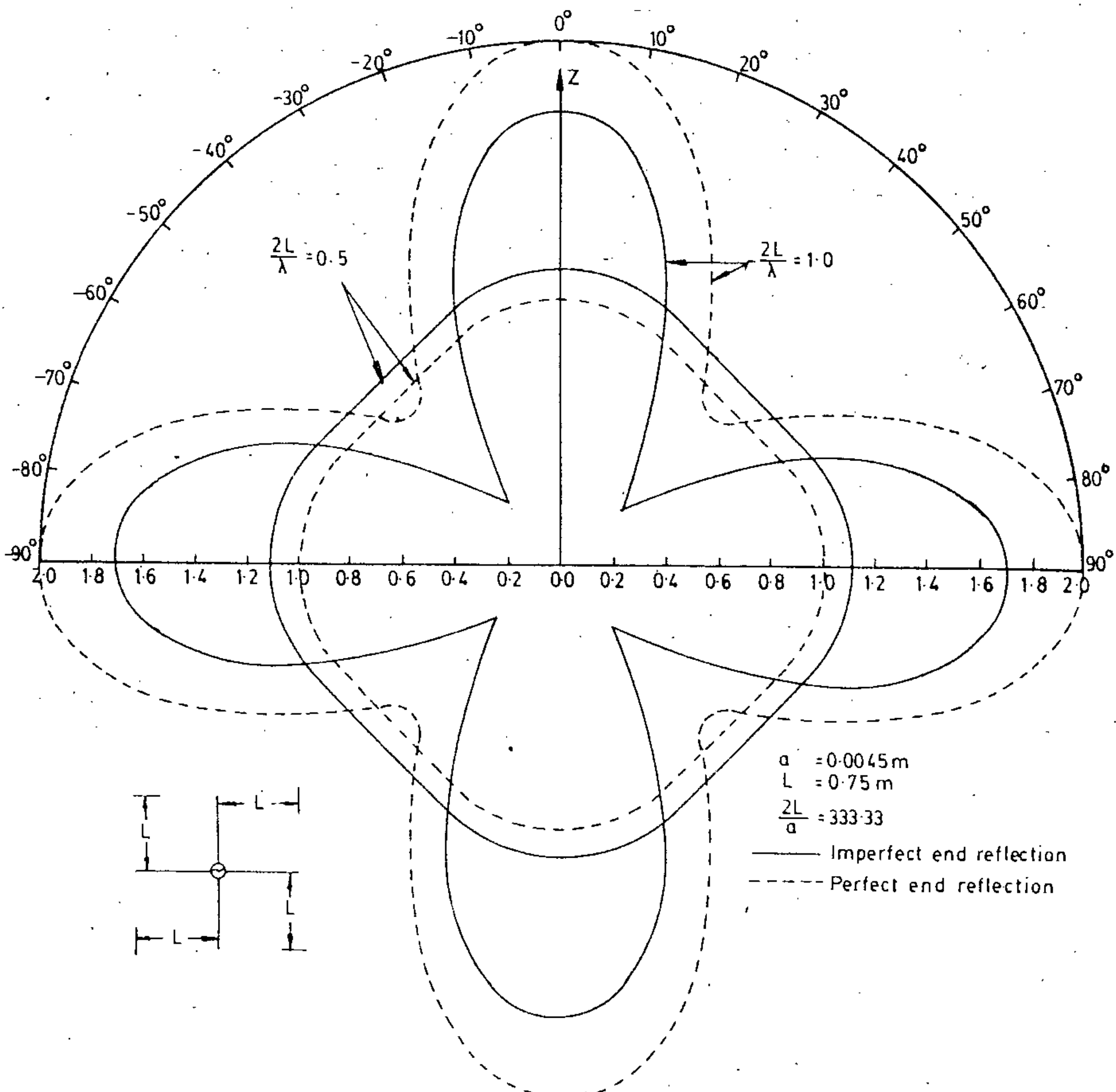

Fig. 3.1 Field pattern of a cross dipole fed with $\alpha = 90^{o}$ and $I_{m1} = I_{m2}$

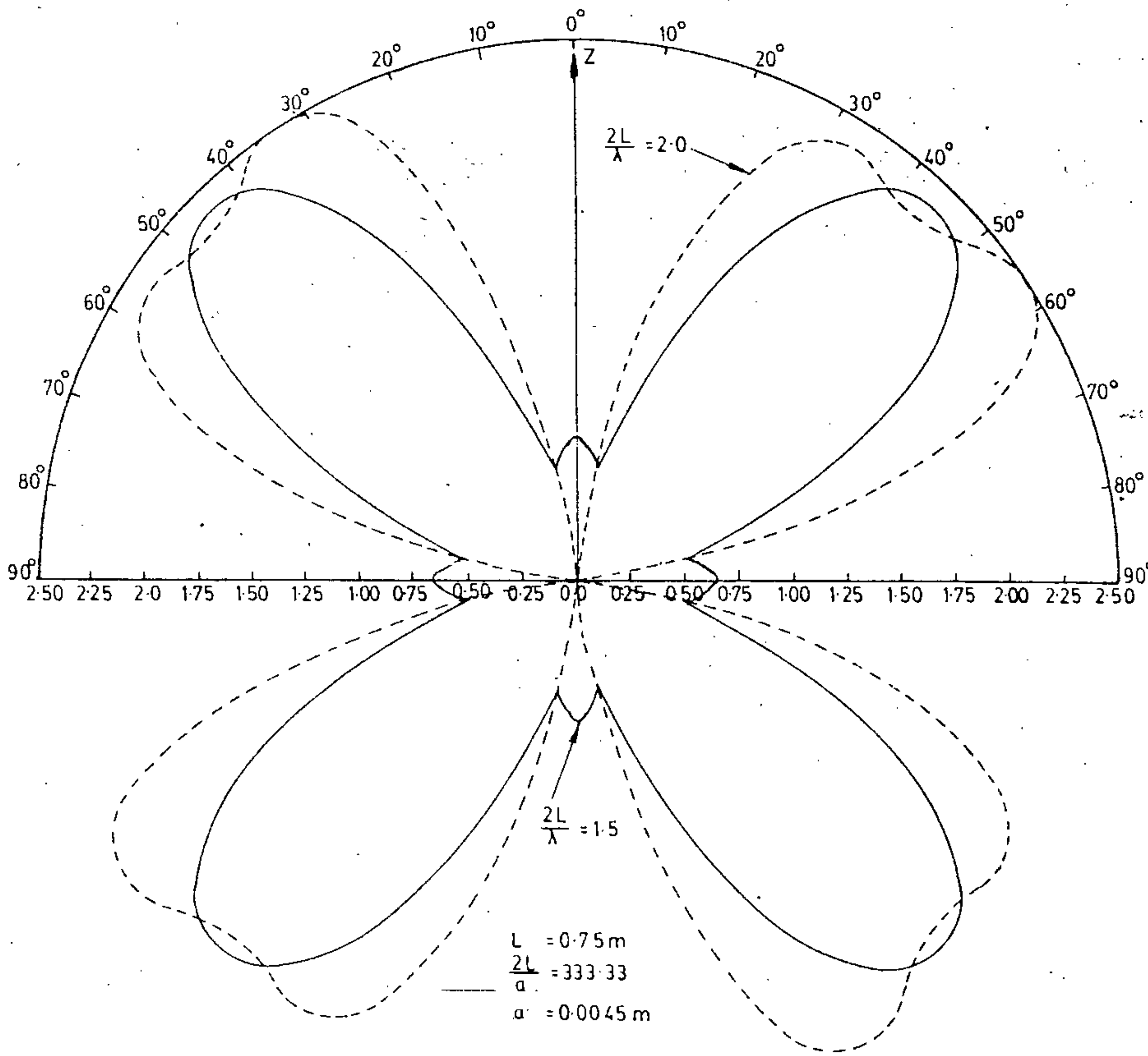

Fig. 3.2 Field pattern of a cross dipole fed with $\alpha = 90^{o}$ and $I_{m1} = I_{m2}$

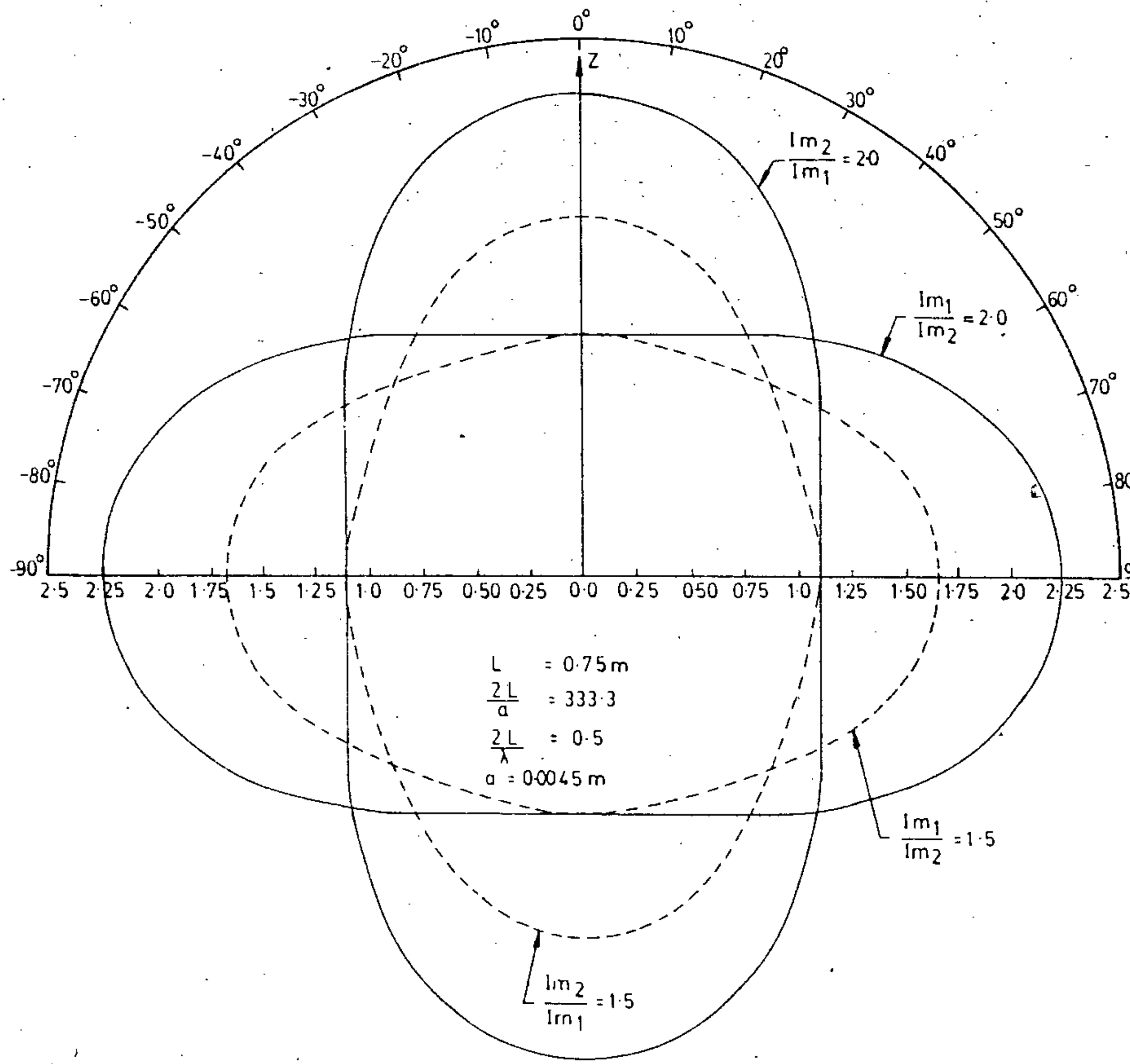

Fig. 3.3 Field pattern of a cross dipole with $\alpha = 90^{\circ}$, $2L/\lambda = 0.5$ and $I_{m1} \neq I_{m2}$

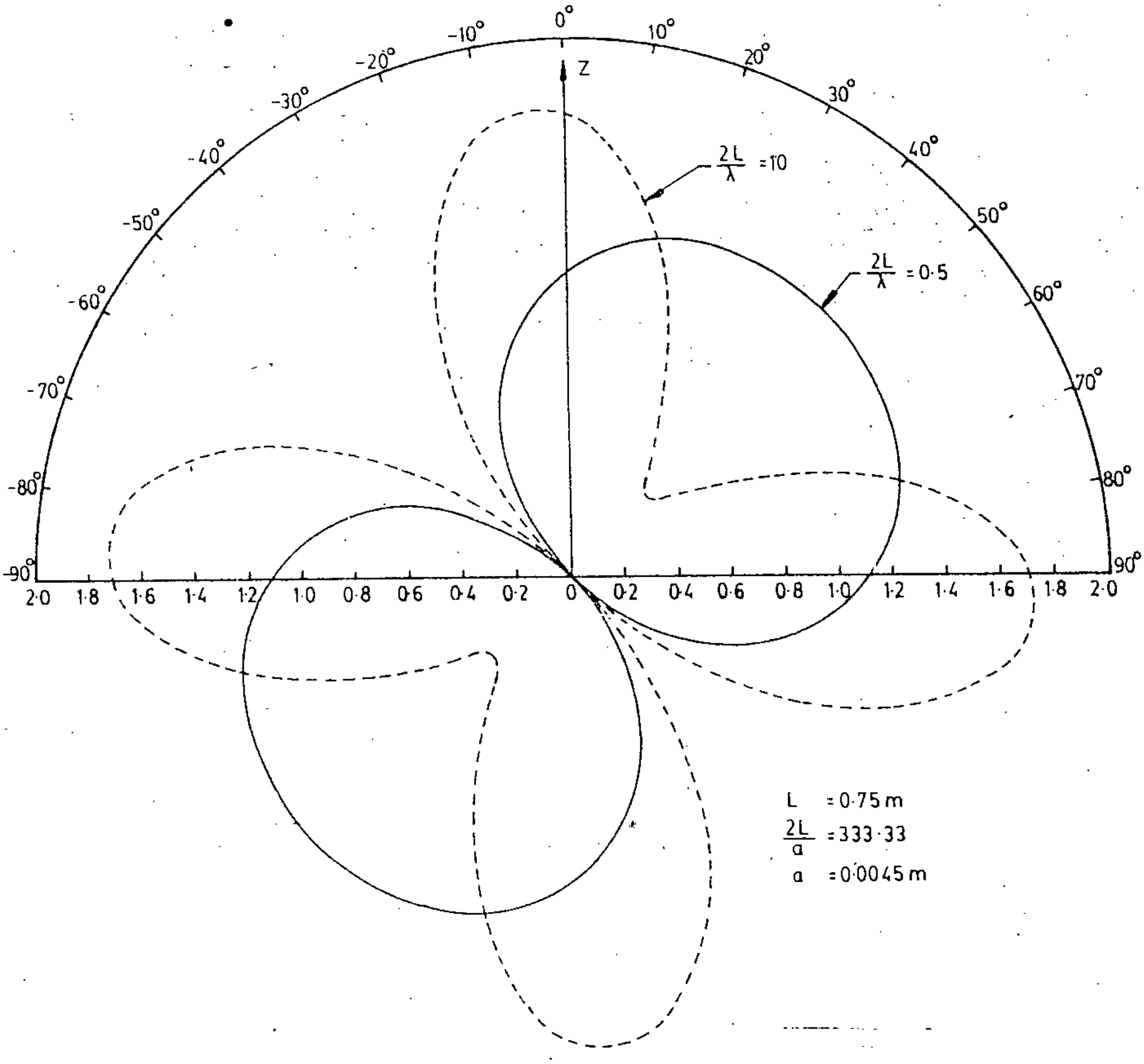

Fig. 3.4 Field pattern of a cross dipole with $\alpha = 0$ and $I_{m1} = I_{m2}$

3.3 Impedance of a Cross-Dipole

In order that the currents on the half-wave length dipoles be in phase quadrature, the dipoles may be connected to separate nonresonant lines of unequal length. Impedance matching is an important feature in the antenna engineering. The mismatched line will cause a standing wave on the feeding cable so that maximum transfer of energy is restricted. In our analysis the feeding systems are assumed matehed with the radiator so that maximum transfer of energy is availed. Suppose, for example, that the terminal impedance of a dipole in a cross arrangement is 100+j0 ohms. Then by connections 100 - ohm lines, as in the schematic diagram of Fig. 3.5(a), with the length of one line 90 electrical degrees longer than the other, the dipoles will be driven with currents of equal magnitude and in phase quadrature. By connecting a 50 - ohm line between the junction point P of the two 100 -ohm lines and the transmitter, the entire transmission line system is matched.

Another method of obtaining quadrature currents is by introducing reactance in series with one of the dipoles [23]. Suppose, for example, that the length and diameter of the dipoles in Fig. 3.5(b) result in a terminal impedance of 100 - j100 ohms. By introducing a series reactance(inductive) of + j100 ohms at each terminal of dipole 1 as in Fig.3.5(b), the terminal impedance of this dipole becomes 100 +j100 ohms. With the two dipoles connected in parallel, the currents are

$$I_1 = \frac{V}{100+j100} \text{ and } I_2 = \frac{V}{100-j100} \qquad \text{........ 3.3(1)}$$

Where $\dot{V}$ = impressed emf.

I_1 = current at terminals of dipole 1

I_2 = current at terminals of dipole 2

So that I_1 and I_2 are equal in magnitude, but I_2 leads I_1 by 90^o. The two impedances in parallel yield

$Z = \frac{1}{Y}$ = 100 + j0 so that a 100 - ohm line will be properly matched when connected to the terminal FF(Fig.3.5(b)).

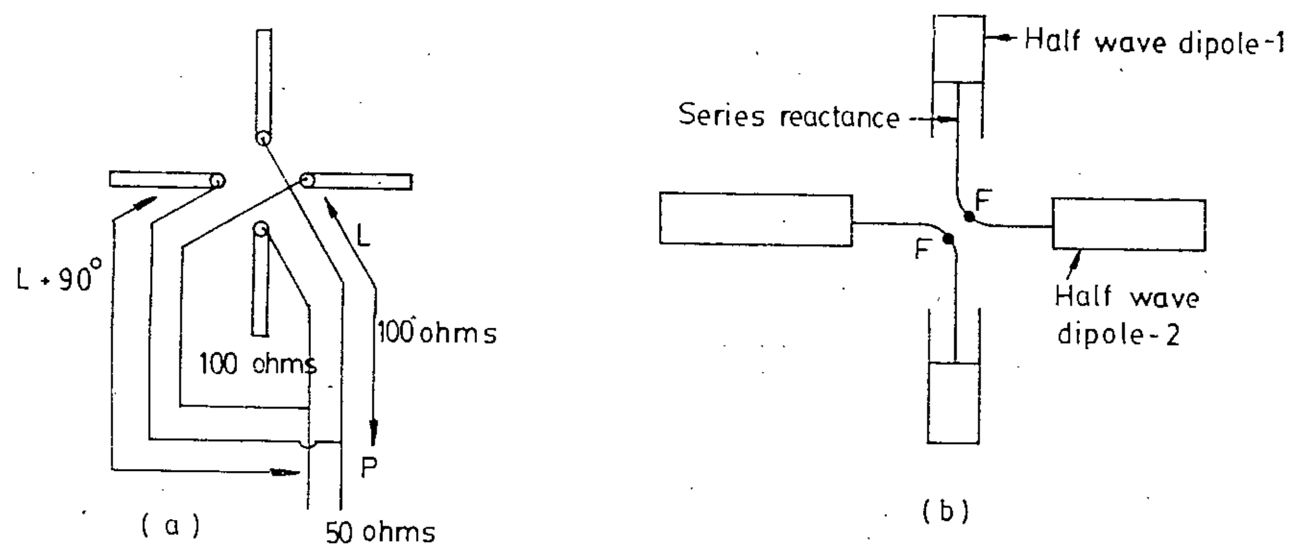

Fig. 3.5(a) Cross Dipole Antenna arrangement, (b) Impedance matching

The impedance of a cross dipole for the SWM of current distribution is described in Fig. 3.6. The crossing of dipoles reduced the impedance with a value that is almost the fifty percent of a single dipole.

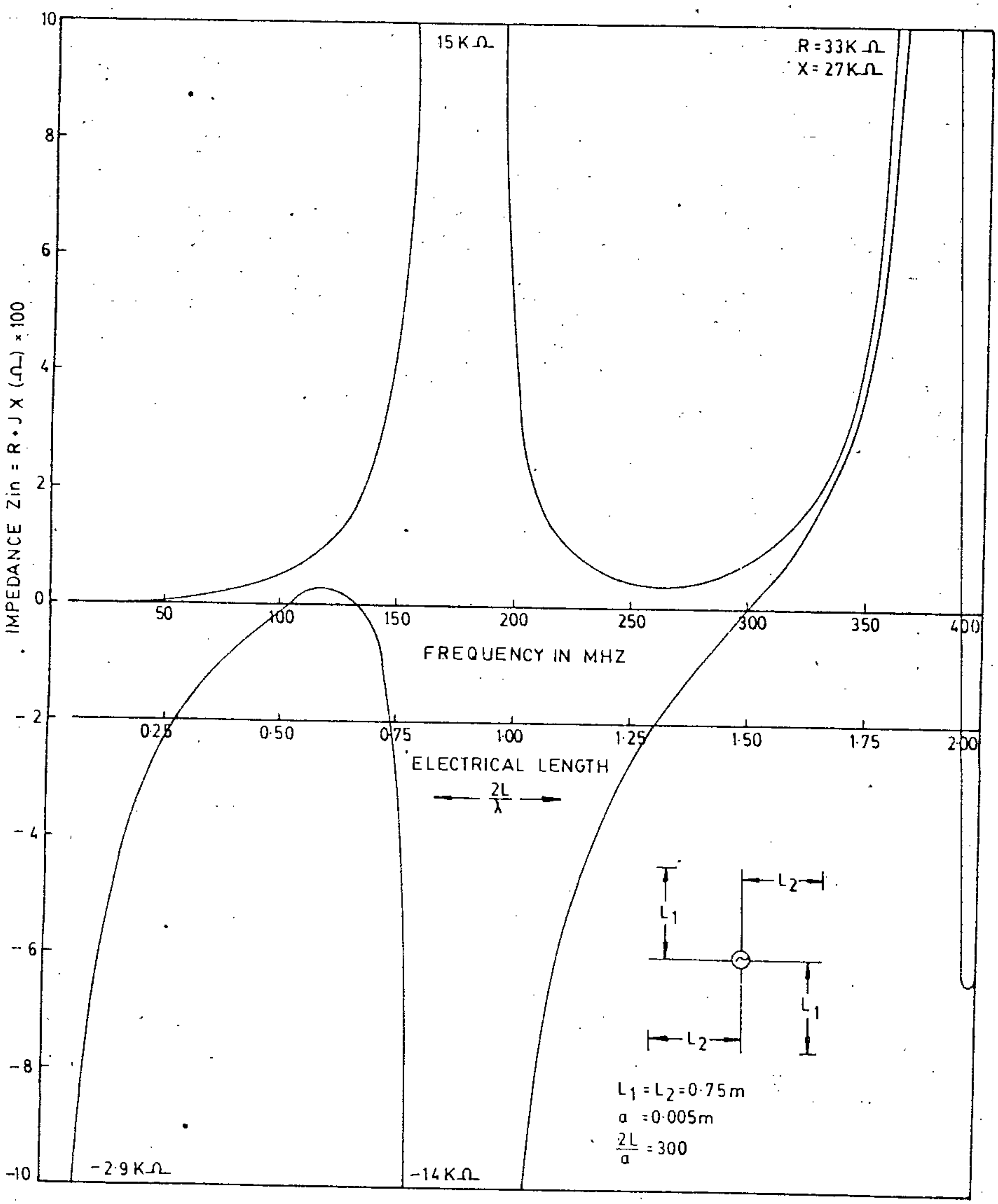

Fig. 3.6 The impedance versus frequency characteristics of a cross dipole using standing wave modelling of current distribution

3.4 Discussion

In order to obtain a low SWR over a considerable band-width, the cross dipole described above has to be modified in the next chapter to form a turnstile antenna. When the two feeders are connected either in series or in parallel, the variation of the input impedance with frequency is reduced compared with the input impedance of a single element. This antenna has excellent characteristics not only over 6 MHZ of one TV channel, but also over several adjoining TV- channels. Its various characteristics has been investigated exhaustively, exploring the frequency range from 5 MHZ to 400 MHZ on purpose to clarify its characteristics. A further modification of cross dipole will stand the analysis in a physical reality and will be discussed in the next chapter.

CHAPTER 4

BATWING ANTENNA ARRAYS AND THEIR TURNSTILING

4.1 Introduction

The antenna which has become the most popular for VHF television broadcasting purposes is the super turnstile, or batwing antenna. This antenna has got extensive use with the opening of the commercial television broadcasting field in 1948. At the present time, over 75 percent of the antennas used for VHF transmitting purposes are superturnstiles or batwings. The turnstile antenna radiator consists of a welded framework of steel pole at 90^{o} intervals. This configuration is referred to as one bay. The antenna radiators are a development of the standard dipole. If the arms of dipoles are enlarged at the end or made cone-shaped, the bandwidth is greatly increased, with no sacrifice in the radiation. It is possible to further increase the bandwidth of the antenna by paralleling two quarter-wave stubs at the feed point. Some of the characteristics of the batwing radiator, which is the heart of the turnstile antenna system are theoretically calculated here by the induced emf method and transmission line analogy. A complete feature of using a shorted-end ring around the array of dipoles is described. A tremendous gain in every respect is available with this structure. In order to give still better bandwidth characteristics and uniform current distribution the batwing structure is sometimes considred as a solid sheet construction. An antenna having solid sheets of material for radiators would offer high wind loading and difficulty in mounting. In order to overcome these difficulties and still maintain the bandwidth, the sheet construction is maintained and a sufficient number of horizontal bars are utilized to provide essentially the same electrical characteristics. This arrangement gives a VSWR of about 1.1 or less over about 30 percent bandwidth, which makes it convenient as a mast-mounted television transmitting antenna for frequencies as low as about 50 MHZ. Unlike the simple turnstile there is relatively little radiation in the axial direction (along the mast) and only one bay is required to obtain a field intensity approximately equal to the maximum field from a single half-wave dipole with the same power input. For decreased beam-width in the vertical plane the superturnstile bays are stacked at intervals of about a wave-length between centers.

4.2 Characteristics of a Linear Dipole Array Forming a a Batwing Structure.

Arrangements in a straight line of more than two elements become much more significant after it is studied as a two-element array. It may be regarded as the first step in forming a more highly directive antenna than a single dipole. A Multi-element array having every element driven from a parallel line are considered to form a batwing array under the construction like a pair of batwings. Fig. 4.1 indicates the geometry of a Batwing array. The linear elements are arranged in the YZ plane and are fed by a transmission line shorted at the ends so that every antenna is getting the signal directly from the transmission line.

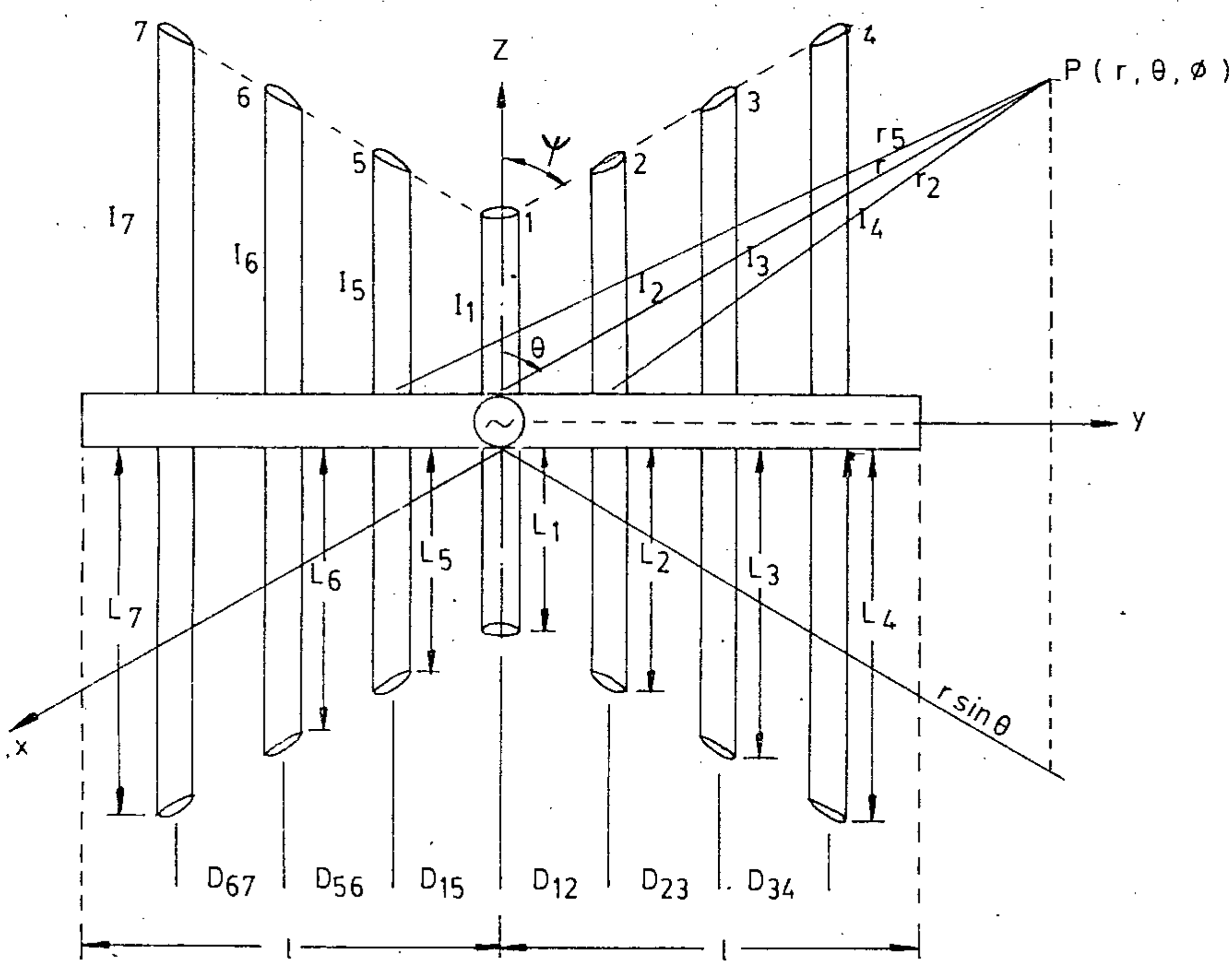

Fig. 4.1. The basic 7-element Batwing array

To determine the voltage applied on each dipole, we are to consider the voltage wave along the feeding line

$$V(Y) = V_{+} e^{jk(\ell - Y)} + V_{-} e^{-jk(\ell - Y)} \quad \ldots\ldots\ldots \; 4.2(1)$$

Where ℓ is the length of the feeding line

V_+ is the incident voltage

and V_- is the reflected voltage.

As the feeder is shorted at both ends at a length ℓ, then imposing the condition $V(\ell) = 0$, we have $V_+ = -V_-$ and hence equation 4.2(1) becomes,

$$V(Y) = V_o \sin K(\ell - Y) \qquad \ldots\ldots\ldots \qquad 4.2(2)$$

Where $V_o = 2jV_+$, is the maximum value of the impressed voltage.

When two or more antennas are used in an array, the driving-point impedance of each antenna depends upon the self-impedance of that antenna and in addition upon the mutual impedance between the given antenna and each of the others. Using the emf method and SWM of current distribution on each element the mutual impedance between any two straight cylindrical dipoles 'i' and 'j' of length 'L_i' and 'L_j' is given by

$$Z_{ij} = Z_{ji} = -\frac{2}{I_i(0) \cdot I_j^*(0)} \int_0^{L_j} E_{iZ}(D_{ij}, Z) I_j^*(Z)\, dZ \ldots 4.2(3)$$

Where, $I_i(0) = \frac{I_{mi}}{2}\left[(1+\rho_i)\sin KL_i - j(1-\rho_i)\cos KL_i\right]$

$$I_j(0) = \frac{I_{mj}}{2}\left[(1+\rho_j)\sin KL_j - j(1-\rho_j)\cos KL_j\right]$$

$$I_j(Z) = \frac{I_{mj}}{2}\left[(1+\rho_j)\sin K(L_j - Z) - j(1-\rho_j)\cos K(L_j - Z)\right]$$

and $E_{iZ}(D_{ij}, Z) = \frac{I_{mi}}{j8\pi w\epsilon} F_{iZ}(D_{ij}, Z)$

ρ_i and ρ_j are the end reflection coefficients of the elements i and j respectively. Substituting these results in the equation 4.2(3) we have the generalized expression for mutual impedance given by,

$$Z_{ij} = Z_{ji} = \frac{j60}{K.\ C_i(0).C_j^*(0)} \int_0^{L_j} F_{iZ}(D_{ij}, Z) C_j^*(Z)\, dZ \qquad \ldots\ldots\ 4.2(4)$$

Where $C_i(0) = (1+\rho_i)\text{Sin } KL_i - j(1-\rho_i)\text{Cos } KL_i$

$C_j(0) = (1+\rho_j)\text{Sin } KL_j - j(1-\rho_j)\text{Cos } KL_j$

$C_j(Z) = (1+\rho_j)\text{Sin}K(L_j - Z) - j(1-\rho_j)\text{Cos } K(L_j - Z)$

$F_{iZ}(D_{ij}, Z) = K(1+\rho_i)F_1(D_{ij}, Z) + j(1-\rho_i)F_2(D_{ij}, Z)$

The input impedance of the overall system and the current in each element are calculated in the following manner:

The applied voltages in each driving elements are from 4.2(2),

$$V_1 = V_o \text{ Sin } K(\ell - D_{11})$$
$$V_2 = V_o \text{ Sin } K(\ell - D_{12})$$
$$V_3 = V_o \text{ Sin } K(\ell - D_{13})$$
$$V_4 = V_o \text{ Sin } K(\ell - D_{14})$$
$$V_5 = V_o \text{ Sin } K(\ell - D_{15})$$
$$V_6 = V_o \text{ Sin } K(\ell - D_{16})$$

and $V_7 = V_o \text{ Sin } K(\ell - D_{17}) \qquad \ldots\ldots\ 4.2(5)$

Let the corresponding currents at the feed points of the various elements be I_1, I_2, I_3, I_4, I_5, I_6 and I_7 respectively. Applying kirchoff's law we can write,

$$
\begin{aligned}
V_1 &= Z_{11} I_1 + Z_{12} I_2 + \quad \ldots\ldots\ldots + Z_{17} I_7 \\
V_2 &= Z_{21} I_1 + Z_{22} I_2 + \quad \ldots\ldots\ldots + Z_{27} I_7 \\
V_3 &= Z_{31} I_1 + Z_{32} I_2 + \quad \ldots\ldots\ldots + Z_{37} I_7 \\
&\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \\
V_7 &= Z_{71} I_1 + Z_{72} I_2 + \quad \ldots\ldots\ldots + Z_{77} I_7 \quad \ldots\ldots \quad 4.2(6)
\end{aligned}
$$

The above simultaneous equations may be written in matrix notation like,

$$
\begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_7 \end{bmatrix} = \begin{bmatrix} Z \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ \vdots \\ I_7 \end{bmatrix} \quad \ldots\ldots\ldots \quad 4.2(7)
$$

and may be solved by the matrix inversion method,

$$
\begin{bmatrix} I_1 \\ I_2 \\ \vdots \\ I_7 \end{bmatrix} = \begin{bmatrix} Z \end{bmatrix}^{-1} \begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_7 \end{bmatrix} \quad \ldots\ldots\ldots \quad 4.2(8)
$$

Whdre $[Z]^{-1}$ is the inverse of the impedance matrix $[Z]$ given by

$$
Z = \begin{bmatrix} Z_{11} & Z_{12} & \ldots\ldots & Z_{17} \\ Z_{21} & Z_{22} & \ldots\ldots & Z_{27} \\ \vdots & \vdots & & \vdots \\ Z_{71} & Z_{72} & \ldots\ldots & Z_{77} \end{bmatrix} \quad \ldots\ldots\ldots \quad 4.2(9)
$$

To find the overall input impedance we have to know the overall current drawn from the supply cable. It is to be noted that the feeding cable will carry the radiating current as well as the transmission line current already fed to each dipole.

We can write,

$$I = I_{Tr} + I_r \qquad \ldots\ldots\ldots \qquad 4.2(10)$$

Where I = total driving current from the supply cable

I_{Tr}=transmission line component of current

I_r= dipole or radiation component of current.

Whatever energy is supplied to the system, the radiated energy must be less than or equal to it. Assuming the optimum transmission of energy we may write,

$$V_1 I_r = V_1 I_1 + V_2 I_2 + V_3 I_3 + \ldots\ldots + V_7 I_7$$

i.e

$$I_r = I_1 + \frac{V_2}{V_1} I_2 + \frac{V_3}{V_1} I_3 + \ldots\ldots + \frac{V_7}{V_1} I_7 \qquad \ldots\ldots \qquad 4.2(11)$$

The radiation component of current may be regarded as the principal contribution to the input impedance. As we are mainly concerned about radiation, the input radiation impedance Z_r may be written as

$$Z_r = \frac{V_1}{I_r} = \frac{V_1^2}{V_1 I_1 + V_2 I_2 + \ldots\ldots + V_7 I_7} \qquad \ldots\ldots \qquad 4.2(12)$$

Transmission line has a great influence on the effective input impedance when the ends are shorted, the terminal loading, $Z_L = 0$.

Taking into consideration of the loading on the feeder by the different dipole elements, the input impedance of the array is given by, $Z_{in} = Z_1 || Z_{T1}/2$ .... 4.2(13)

Where $Z_1 = \frac{V_1}{I_1}$, $Z_{T1} = Z_{OT} \frac{Z_{i2} \text{Cos} KD_{12} + jZ_{OT} \text{Sin} KD_{12}}{Z_{OT} \text{Cos} KD_{12} + jZ_{i2} \text{Sin} KD_{12}}$

Z_{OT} is the characteristic impedance of the feeder.

$Z_{i2} = Z_2 || Z_{T2}$, $Z_2 = \frac{V_2}{I_2}$

$$Z_{T2} = Z_{OT} \frac{Z_{i3} \text{Cos} KD_{23} + jZ_{OT} \text{Sin} KD_{23}}{Z_{OT} \text{Cos} KD_{23} + jZ_{i3} \text{Sin} KD_{23}}$$

$Z_{i3} = Z_3 || Z_{T3}$, $Z_3 = \frac{V_3}{I_3}$

$$Z_{T3} = Z_{OT} \frac{Z_{i4} \text{Cos} KD_{34} + jZ_{OT} \text{Sin} KD_{34}}{Z_{OT} \text{Cos} KD_{34} + jZ_{i4} \text{Sin} KD_{34}}$$

In the present Case, $Z_{i4} = Z_4 = \frac{V_4}{I_4} = 0$

$\therefore$ $Z_{T3} = jZ_{OT} \tan (KD_{34})$ .... 4.2(14)

The characteristic impedance of the feeder is given by [24] for $2a << d$,

$Z_{OT} = 276 \log_{10} \frac{d}{a}$ .... 4.2(15)

Where a = radius of each conductor of the feeder

d = separation between the conductors

We have taken $Z_{OT} = 300$ ohms.

Referring to the geometry of the dipole array of Fig. 4.1, taking the driving point of the dipole array as the phase center, the radiation field at the point 'P' due to all the antenna elements is given by,

$$E_\theta = \sum_{i=1}^{7} E_{\theta i} = j60 \sum_{i=1}^{7} \frac{I_{mi}}{r_i} e^{-jkr_i} P_i(\theta) \quad \text{....4.2(16)}$$

Where,

$$r_1 \cong r - D_{11} \sin\theta \sin\phi$$

$$r_2 \cong r - D_{12} \sin\theta \sin\phi$$

$$r_3 \cong r - D_{13} \sin\theta \sin\phi$$

$$r_4 \cong r - D_{14} \sin\theta \sin\phi$$

$$r_5 \cong r + D_{15} \sin\theta \sin\phi$$

$$r_6 \cong r + D_{16} \sin\theta \sin\phi$$

$$r_7 \cong r + D_{17} \sin\theta \sin\phi$$

$P_i(\theta)$ is the pattern factor of the individual dipole defined as,

$$P_i(\theta) = \frac{1+\rho_i}{2} \cdot \frac{\cos(KL_i \cos\theta) - \cos KL_i}{\sin\theta} - j\,\frac{1-\rho_i}{2} \cdot \frac{\sin KL_i - \cos\theta \sin(KL_i \cos\theta)}{\sin\theta}$$

The far field consideration will cause

$r_i \simeq r$ and the phase factors become

$$S_1 = e^{jKD_{11} \sin\theta \sin\phi} \simeq 1$$

$$S_2 = e^{jKD_{12} \sin\theta \sin\phi}$$

$$S_3 = e^{JKD_{13} \sin\theta \sin\phi}$$

$$S_4 = e^{jKD_{14} \sin\theta \sin\phi}$$

$$S_5 = e^{-jKD_{15} \sin\theta \sin\phi}$$

$$S_6 = e^{-jKD_{16} \sin\theta \sin\phi}$$

$$S_7 = e^{-jKD_{17} \sin\theta \sin\phi} \qquad \ldots\ldots \quad 4.2(17)$$

The resultant far-field at the point P due to all the antenna elements is given by

$$E_\theta = \frac{j60}{r} e^{-jkr} (I_{m1} P_1 S_1 + I_{m2} P_2 S_2 + \ldots\ldots + I_{m7} P_7 S_7) \quad \ldots\ldots\ldots \quad 4.2(18)$$

Where P_1, P_2 ... etc. are the respective pattern factors and S_1 , S_2 ...... etc are the respective phase factors defined in 4.2(17). I_{m1} , I_{m2} ..... etc are the respective maximum currents in the dipoles and can be evaluated as follows:

I_1, I_2 ..... etc are the feeding point currents in the dipoles.

Therefore at the feed point,

$$I_i = \frac{I_{mi}}{2} \left[(1+\rho_i) \operatorname{Sin} KL_i - j(1-\rho_i) \operatorname{Cos} KL_i \right]$$

$$\text{i.e.} \quad I_{mi} = \frac{2I_i}{(1+\rho_i) \operatorname{Sin} KL_i - j(1-\rho_i) \operatorname{Cos} KL_i} \quad \ldots\ldots \quad 4.2(19)$$

The pattern factor of this batwing array can be normalized with the maximum current on the first dipole,

$$\therefore \quad P(\theta) = P_1 S_1 + \frac{I_{m2}}{I_{m1}} P_2 S_2 + \ldots\ldots + \frac{I_{m7}}{I_{m1}} P_7 S_7 \ldots\ldots 4.2(20)$$

In order to obtain a radiation pattern with polarization in a horizontal plane the array should radiate in the broadside with $\phi = 0$. The phase factors then become unity and the resulting field pattern is given by

$$P(\theta) = P_1 + \frac{I_{m2}}{I_{m1}} P_2 + \frac{I_{m3}}{I_{m1}} P_3 + \ldots\ldots + \frac{I_{m7}}{I_{m1}} P_7 \ldots\ldots 4.2(21)$$

Let us assume the dipoles in the batwing array are equally spaced and a complete feature of the length of the dipole are determined from the half-angle of tapering made at the center dipole. If L_1 is the half-length of the center dipole, the length of the other dipoles are given by [Fig.4.1]

$$L_2 = L_1 + \Delta L = L_5$$
$$L_3 = L_1 + 2\Delta L = L_6$$
$$L_4 = L_1 + 3\Delta L = L_7 \qquad \ldots\ldots\ldots \qquad 4.2(22)$$

Where $\Delta L = \frac{D}{\tan\psi}$

D = separation between two consecutive dipoles

ψ = half-angle of the tapering of the array.

A convenient method of measuring the directive properties of an antenna is to define a gain function, $G(\theta, \phi)$, where θ and ϕ are the colatitude and the azimuthal angles respectively of the associated spherical Co-ordinate system. The gain of an antenna is defined as the ratio of the maximum radiation intensity at the peak of the main beam to the radiation intensity in the same direction that would be produced by an isotropic radiator with the same input power. For antennas that have no internal losses, the gain is the same as the directivity. The gain of an antenna is given by,

$$G = \frac{4\pi r^2 S(\theta, \phi)}{W} \qquad \ldots\ldots\ldots \qquad 4.2(23)$$

Where $S(\theta,\varphi)$ is the radiated power density pattern and W is the average power radiated by the antenna.

W is given by

$$W = \frac{1}{2} \text{ Re} \left[I_r \cdot I_r^* \cdot Z_r \right] \text{ watts} \quad \ldots\ldots\ldots \quad 4.2(24)$$

and $S(\theta,\varphi) = \frac{1}{2}\eta \, \text{Re}(E_\theta \cdot E_\theta^* - E_\varphi \cdot E_\varphi^*) \text{ watt/m}^2$ ..... 4.2(25)

η is the intrinsic impedance of air, $\eta = 120\pi$ ohms and,

E_θ and E_φ are respectively the theta and phai components of the electric field at far distance from the antenna.

For the dipole array there is no phai-polarization, so that the gain function reduces to,

$$G = \frac{120}{R_r} \cdot \frac{(I_{m1}P_\theta) \cdot (I_{m1} P_\theta)^*}{I_r \cdot I_r^*} \quad \ldots\ldots\ldots \quad 4.2(26)$$

where $R_r = R_e \left[Z_r \right]$ is the radiation resistance and P_θ is the pattern factor of the array. The gain in decibels may be denoted by Gdb,

$$\text{Gdb} = 10 \log_{10}(G_{max}) \quad \ldots\ldots\ldots \quad 4.2\ (27)$$

where G_{max} is the maximum gain occuring at $\theta = 90^\circ$

Using the above equations the properties of the seven-element batwing array were numerically calculated. Fig.4.2 describes the field pattern of a batwing array in the horizontal plane of polarization. The array is operated at a certain frequency say 150 MHZ for model study, where the gain has an appreciable figure. The field pattern of a single dipole operated at the same frequency is also depicted with broken lines for comparison. It is observed that the batwing array has got almost three times more intensified field strength at $\theta = 90^\circ$, in comparison with a simple dipole. Fig. 4.3 indicates the impedance and gain variation for a batwing array.

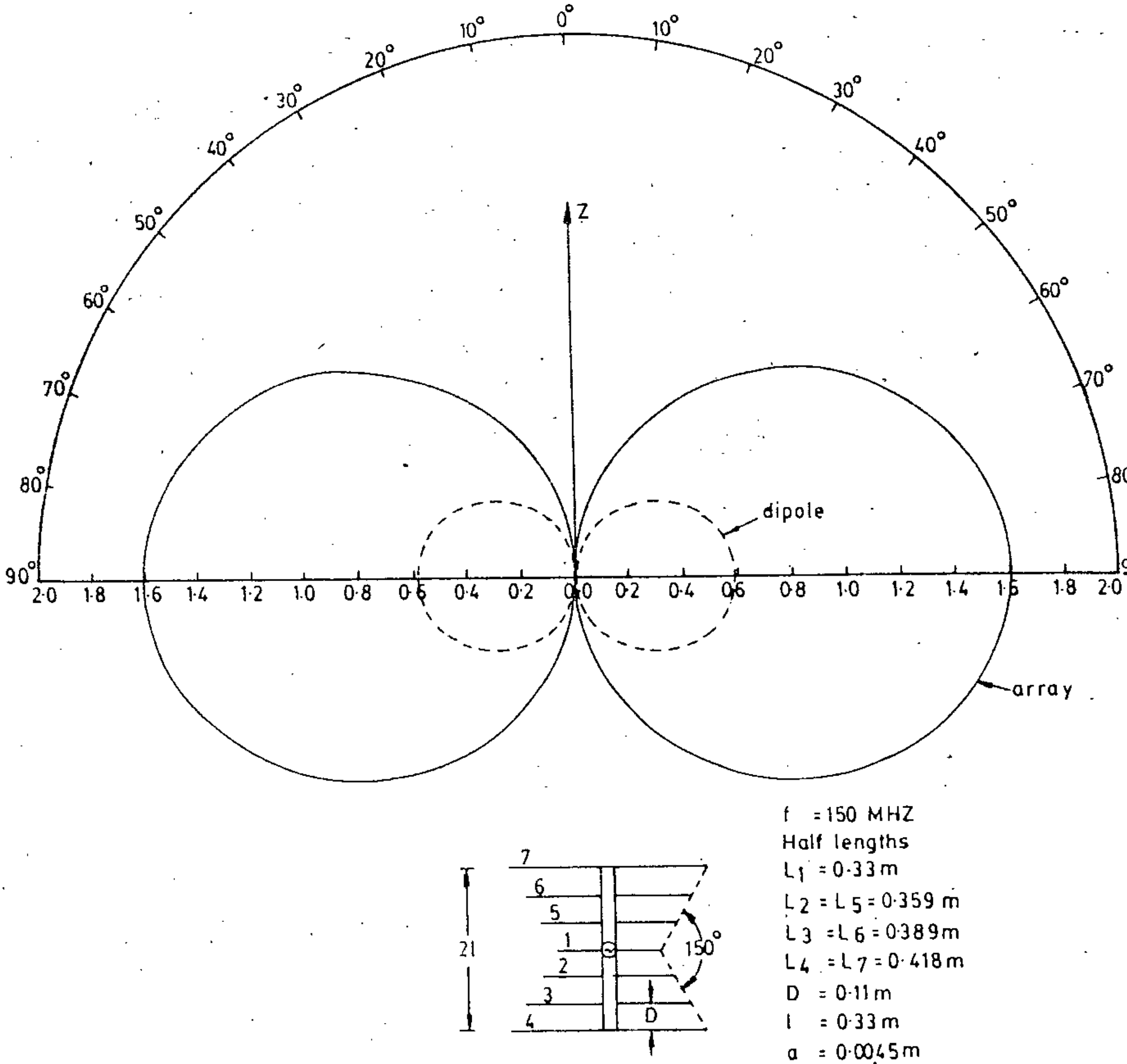

Fig. 4.2 Broadside field pattern of a batwing array

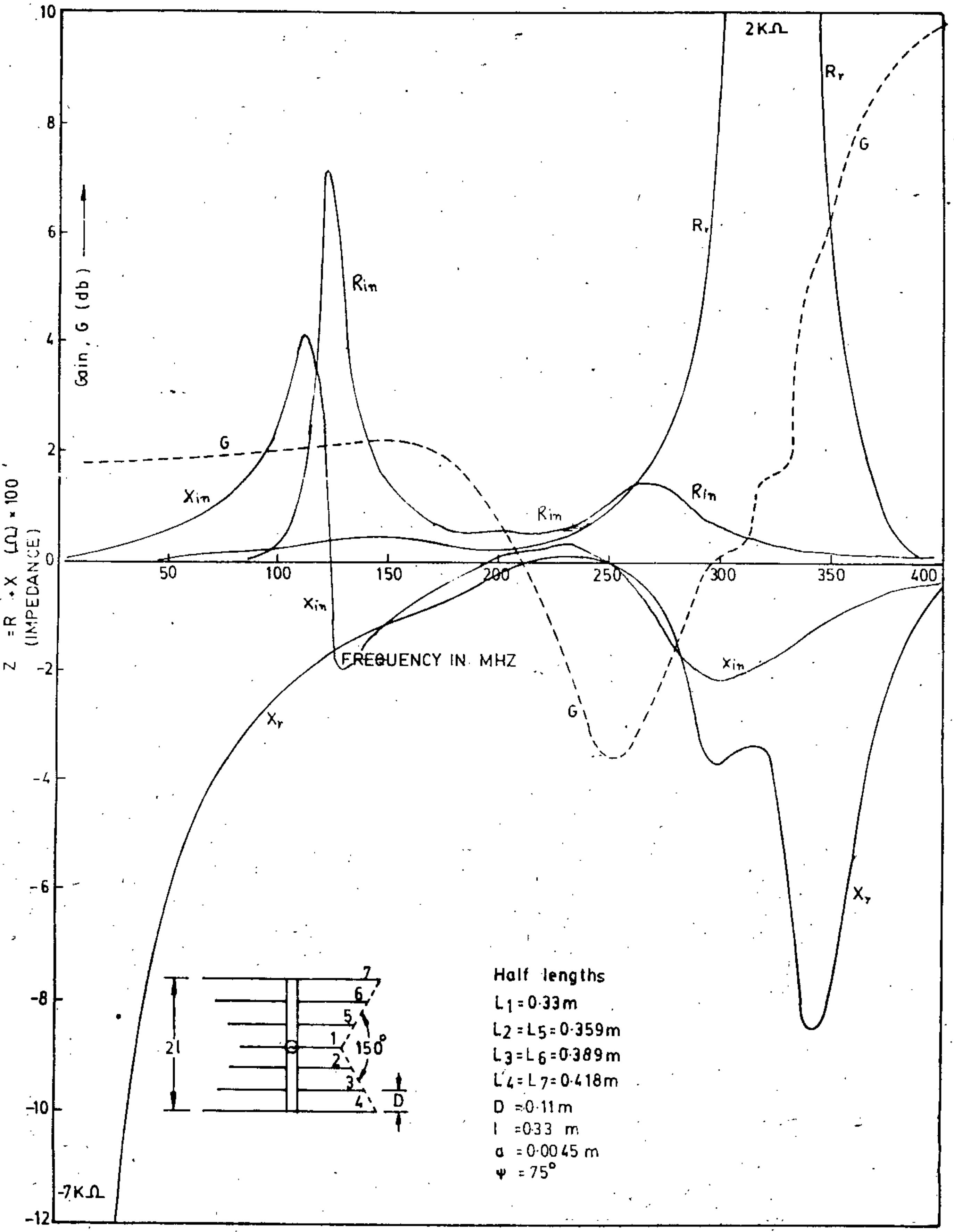

Fig. 4.3 Impedance and gain versus frequency characteristics of a batwing array, $Z_{OT} = 300\,\Omega$.

The analysis is carried out for a frequency range from 5 MHZ to 400 MHZ. The gain is almost constant upto the frequency 175 MHZ and then has a stop-band between 220 MHZ to 300 MHZ. Above 300 MHZ the gain increases sharply and then occupies another wide bandwidth.

4.3 Characteristics of a Batwing Array with the Ends Shorted by a Peripheral Ring.

A theoretical establishment of a batwing array having all the dipole elements shorted in the periphery is not available with a specific modelling. A shorted-end ring placed over the perimeter is motivated for improvement of the mechanical and electrical characteristics of the antenna. When the ends of a dipole are shorted by a wire the terminal voltage is assumed to be zero. By the use of transmission line concept, the standing wave voltage and current on the dipole may be written as,

$$V(Z) = Ae^{jK(L-Z)} + Be^{-jK(L-Z)}$$

and

$$I(Z) = \frac{Ae^{jK(L-Z)} - Be^{-jK(L-Z)}}{Z_c} \qquad \ldots\ldots\ldots \qquad 4.3(1)$$

where

Z_c is the characteristic impedance of the dipole; A and B are magnitudes of the incident and reflected voltages respectively.

When a ring is used, the ends of the elements are shorted so that their terminal voltages will be zero, i.e.

$$V(L) = 0 \qquad \ldots\ldots\ldots \qquad 4.3(2)$$

Approaching to a single dipole we may apply the condition 4.3(2) into 4.3(1), so that

$$B = - A \quad \ldots\ldots\ldots \quad 4.3(3)$$

Substituting 4.3(3) into 4.3(1) we get

$$I(Z) = I_m \operatorname{Cos}K(L-Z) \quad \ldots\ldots\ldots \quad 4.3(4)$$

Whdre $I_m = \frac{2A}{Z_c}$, the maximum value of the current on the dipole.

Therefore when a shorted-end ring is used in the periphery of the array the only standing wave current on the elements will be a cosine wave enhancing the charge accumulation towards the ring. Using the cosine distribution described in 4.3(4), the axial electric field component of a dipole is given by

$$E_Z = \frac{j\, I_m}{4 \pi w \varepsilon} F_2(r, Z) \quad \ldots\ldots \quad 4.3(5)$$

where $F_2(r, Z) = (L-Z)\left(\frac{jK}{r_1^2} + \frac{1}{r_1^3} \right) e^{-jkr_1} + (Z+L) \cdot$

$$\left(\frac{jK}{r_2^2} + \frac{1}{r_2^3} \right) e^{-jkr_2} - 2 K \sin KL \frac{e^{-jkr_o}}{r_o}$$

$$r = \sqrt{x^2 + y^2}$$
$$r_1 = \sqrt{r^2 + (Z-L)^2}$$
$$r_2 = \sqrt{r^2 + (Z+L)^2}$$
$$r_o = \sqrt{r^2 + Z^2}$$

Using the emf method and the cosine distribution, the mutual impedance between any two straight dipoles 'i' and 'j' is given by

$$Z_{ij} = Z_{ji} = - \frac{2}{I_i(0) . I_j^*(0)} \int_0^{L_j} F_{iZ}(D_{ij}, Z) I_j^*(Z) dz \quad .... 4.3(6)$$

where $Z_{ij} = Z_{ji}$ = mutual impedance between the i th and the j th element.

$I_i(0) = I_{mi} \; \mathrm{Cos} KL_i$

$I_j(0) = I_{mj} \; \mathrm{Cos} KL_j$

$I_j(Z) = I_{mj} \; \mathrm{Cos} K(L_j - Z)$

D_{ij} = separation between the i th and j th element

L_i, L_j = half-length of i th and j th element respectively

Substituting the above results in the equation 4.3(6) we have the mutual impedance for the precribed distribution,

$$Z_{ij} = Z_{ji} = - \frac{j\,60}{\mathrm{Cos}\; KL_i \;\mathrm{Cos}\; KL_j} \int_0^{L_j} F_2\,(D_{ij}, Z)\; \mathrm{Cos}\; K(L_j - Z) dZ$$

.... 4.3(7)

Using the similar procedure described in chap.2, we have the theta component of the electric field at far distance from the antenna of a single dipole,

$$E_{\theta i} = \frac{j \eta e^{-jkr}}{2\pi r} I_{mi} \cdot \frac{\mathrm{Sin} KL_i - \mathrm{Cos}\,\theta \; \mathrm{Sin}(KL_i \mathrm{Cos}\,\theta)}{\mathrm{Sin}\,\theta} \quad ....4.3(8)$$

where $\eta = 120\pi$ intrinsic impedance of air,

The pattern factor of the dipole is defined as

$$P_i(\theta) = \frac{\mathrm{Sin} KL_i - \mathrm{Cos}\,\theta \; \mathrm{Sin}(KL_i \; \mathrm{Cos}\,\theta)}{\mathrm{Sin}\,\theta} \quad .... 4.3(9)$$

The pattern factor of the batwing array with peripheral ring can also be normalized in the like manner as equation 4.2(20) and for horizontal polarization, it is given by

$$P(\theta) = P_1 + \frac{I_{m2}}{I_{m1}} P_2 + \cdots\cdots\cdots + \frac{I_{m7}}{I_{m1}} P_7 \quad \cdots\cdots \qquad 4.3(10)$$

where P_1, P_2 .... etc are the pattern factors of the respective dipole,

and I_{m1} , I_{m2} .... etc are the maximum currents on the dipoles.

A tremendous increase of radiation happens when a cosine distribution stands on the antenna elements. Fig. 4.4 shows the field pattern of a batwing array with a peripheral ring. In comparison with this pattern, the pattern for a single dipole with the same operating frequency is negligible. This intensification of the field strength will cause more radiation. Fig. 4.5 describes the impedance and gain variation for a batwing array with peripheral ring. Here the radiation resistance is resonably increased causing greater radiation. The resonance at 21o MHZ will not cause any serious affect on the radiation. High gain is observed over a wide range of frequencies from 5 MHZ to 260 MHZ which covers most of the TV-channels.

4.4 A Turnstile Bay of Two Batwing Arrays

The batwing antenna radiator consists of a welded framework of steel tubes in the form of batwing. Some of these are mounted around a tubular steel pole at 90^{o} intervals.

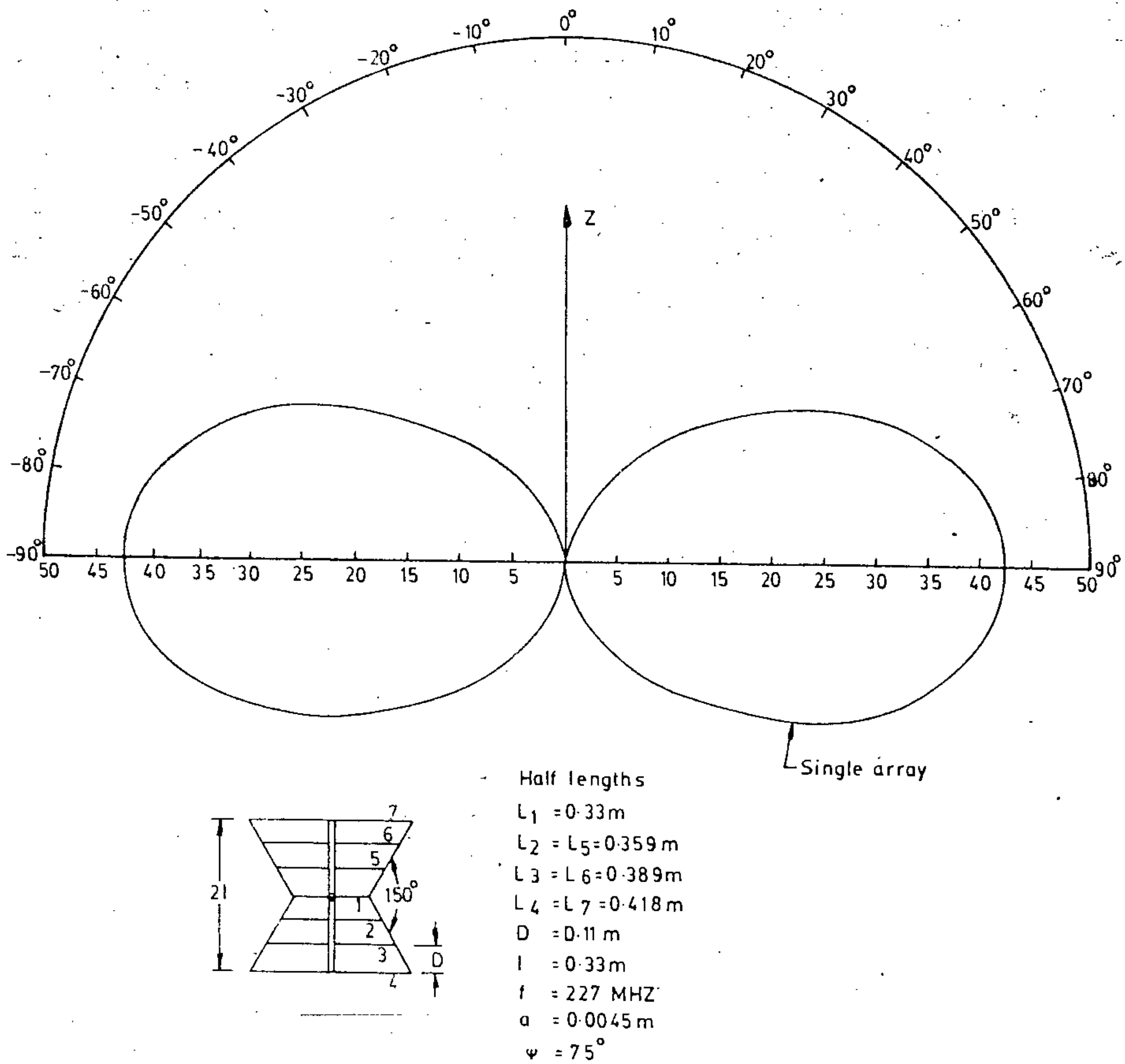

Fig. 4.4 Field pattern of a batwing antenna when a shorting ring is used in the periphery of the array

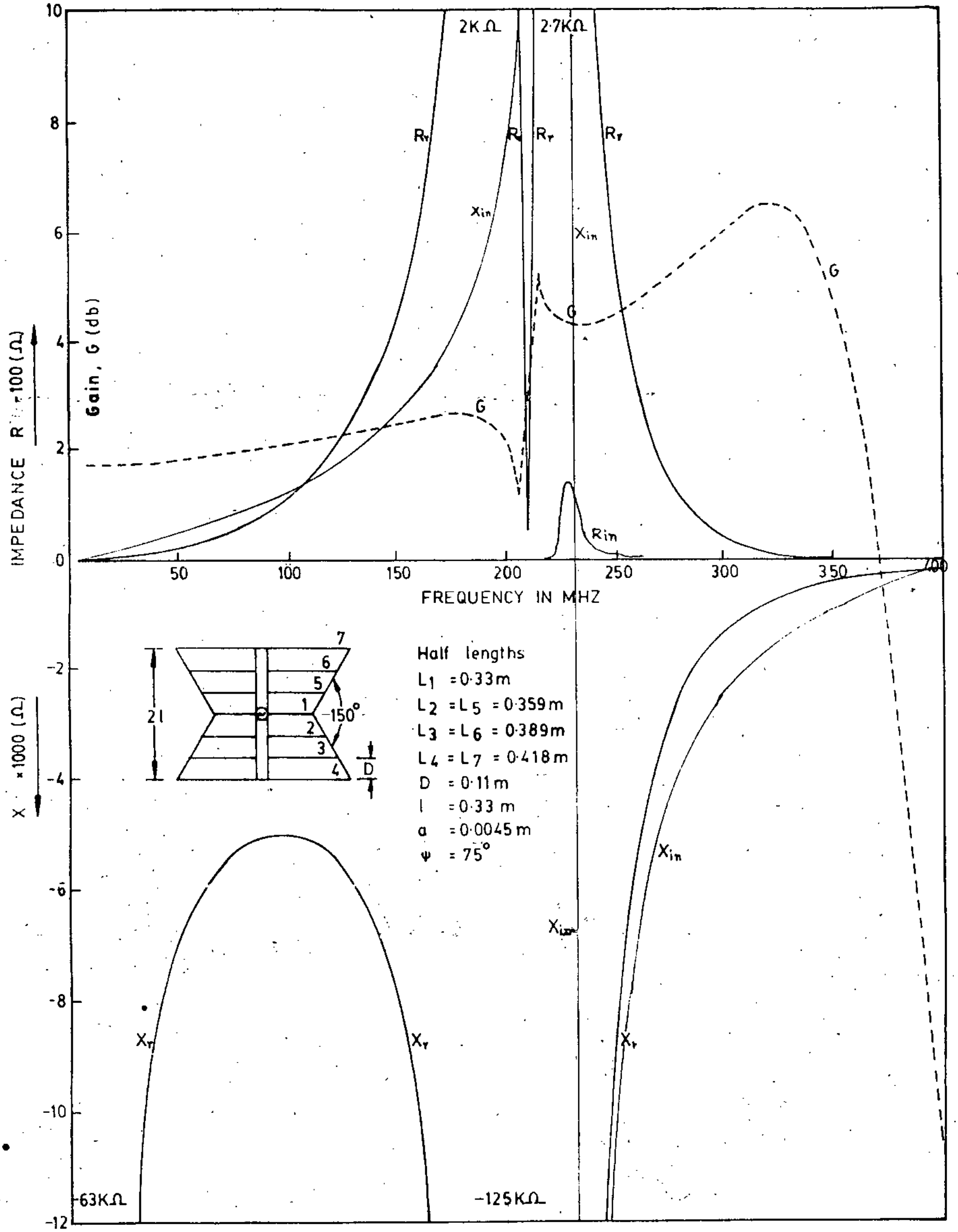

Fig.4.5 Impedance and gain versus frequency characteristics of a batwing antenna with a peripheral ring, Z_{OT} = 300 Ω.

This configuration is referred to as one-bay fed in opposite pairs in order to improve the circularity. In addition, the arrays are fed in phase quadrature, commonly referred to as turnstiling. The isometic view of one turnstiling bay is shown in Fig. 4.6. This requirement demands that the antenna be fed by two transmission lines, which must be kept electrically similar in order to ensure the proper current and phase relationship. The pair of radiators array are referred to as "north-south" and "east-west" to denote the two groups of radiators which are fed by separate transmission lines. A turnstile antenna may be conveniently mounted on a vertical mast. The mast is coincident with the axis of the turnstile.

The two batwing arrays constructing the turnstile antenna are identical in all respect. The identical single radiators carry equal currents but in phase quadrature. Therefore field pattern for a single batwing array will also be in phase quadrature. The total electric field pattern for a single bay turnstile antenna is given by,

$$P(\theta) = P_{NS}(\theta) - jP_{EW}(\theta) \qquad \ldots\ldots \qquad 4.4(1)$$

Where $P_{NS}(\theta)$, field pattern of the north-south pair

$P_{EW}(\theta) = P_{NS}(90^o - \theta)$, field pattern of the quadrature (east-west) pair

and $j = \sqrt{-1}$

We will study here the two cases of array furnished with and without shorted end rings at the periphery. The field patterns

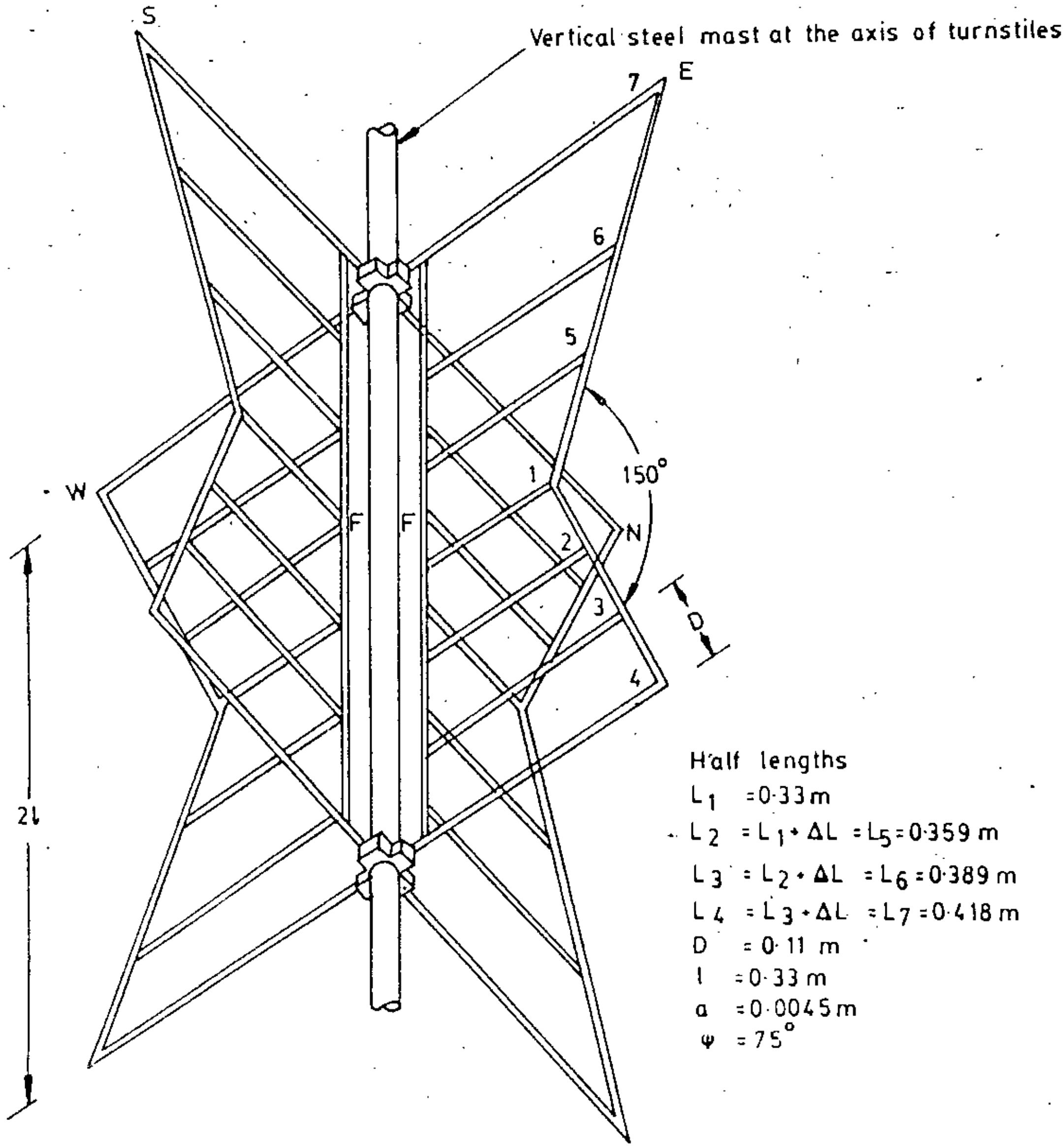

Fig. 4.6 Construction of a single-bay turnstile antenna

for batwing array without peripheral ring are discussed in art 4.2. Using the same rule followed by equation 4.4(1) we may derive the field pattern of a cross-batwing array i.e. a turnstile array without shorted-end ring at the Periphery. Fig. 4.7 describes such a pattern almost circularly polarized in the horizontal θ-plane. This specimen pattern is studied at 275 MHZ and may also be obtained as a good and resonable pattern over a wide range of operation. The gain of such a bay is quite resonable and more improvement follows from the single batwing array discussed in article 4.2. Fig. 4.8 indicates the impedance and gain variation of a turnstile array without ring at the periphery. A good and resonable radiation resistance is observed after 250 MHZ, where both are increasing linearly upto 325 MHZ. After 325 MHZ radiation resistance decreases but gain increases.

A partial treatment on the use of end ring is established in the art. 4.3. The ring at the periphery is used to enhance the radiation and that is observed when the current distribution tur to cosine wave distribution. Fig. 4.9 shows the field pattern of a turnstile array using the shorted-end rings. The field pattern of a single dipole or a cross-dipole is insignificant compared to the pattern described in Fig. 4.9. A tremendous amplification results when such an arrangement is made. The current of the ring has no contribution to the field in the broad-side. A simultaneous treatment on the gain and impedance

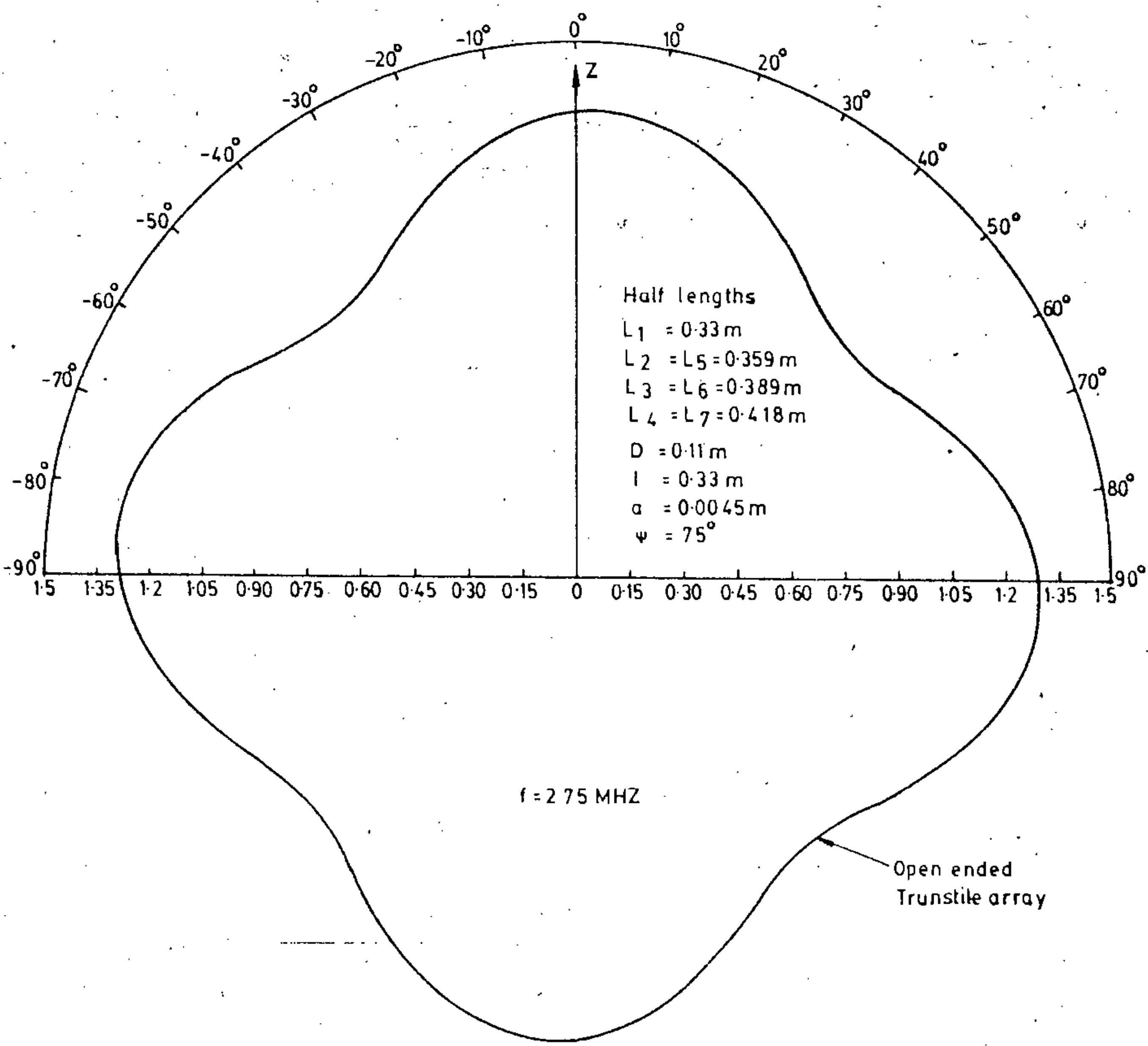

Fig. 4.7 Broadside field pattern of a turnstile array without a peripheral ring

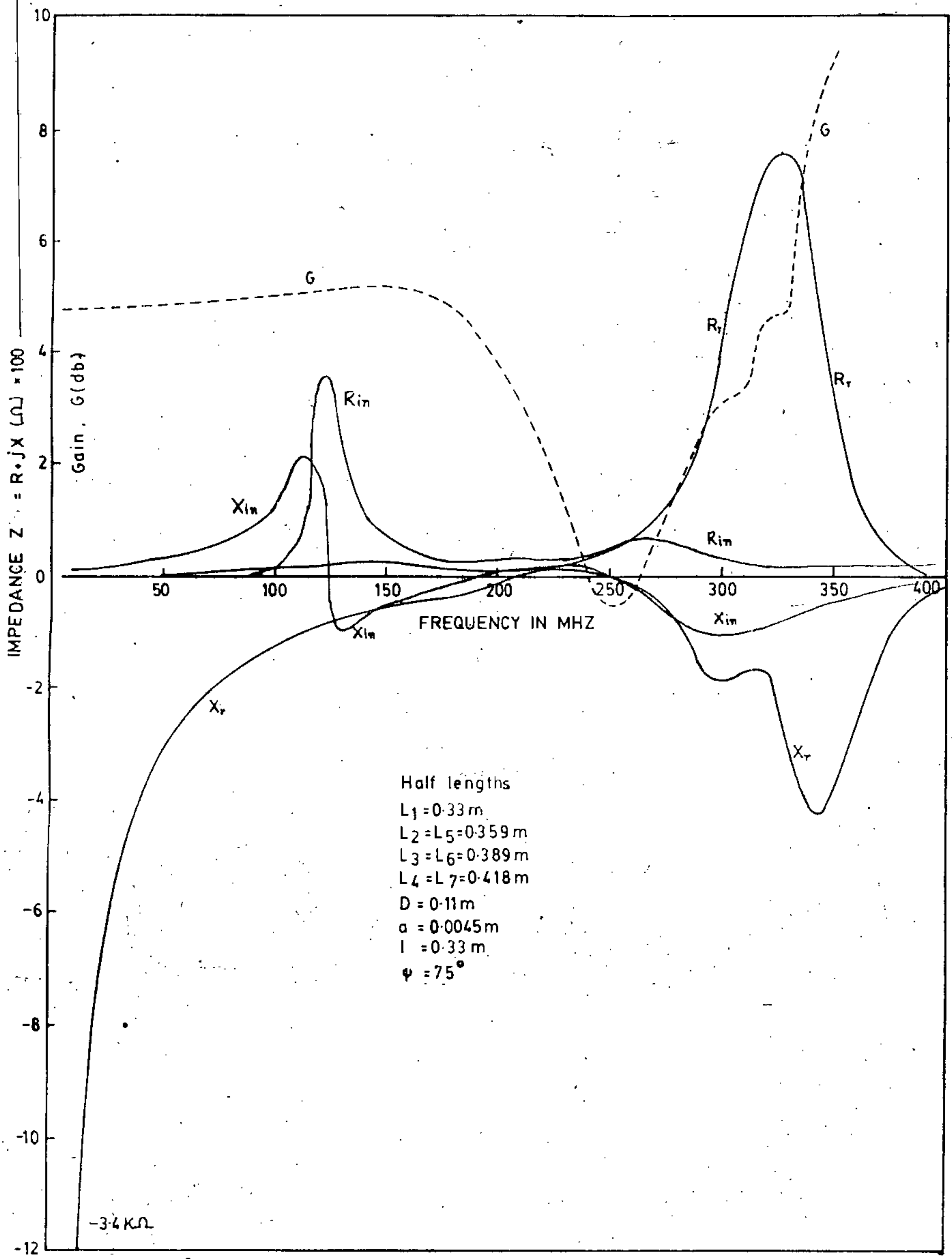

Fig. 4.8 Impedance and gain versus frequency characteristics of a turnstile array without a peripheral ring, Z_{OT} = 300 Ω.

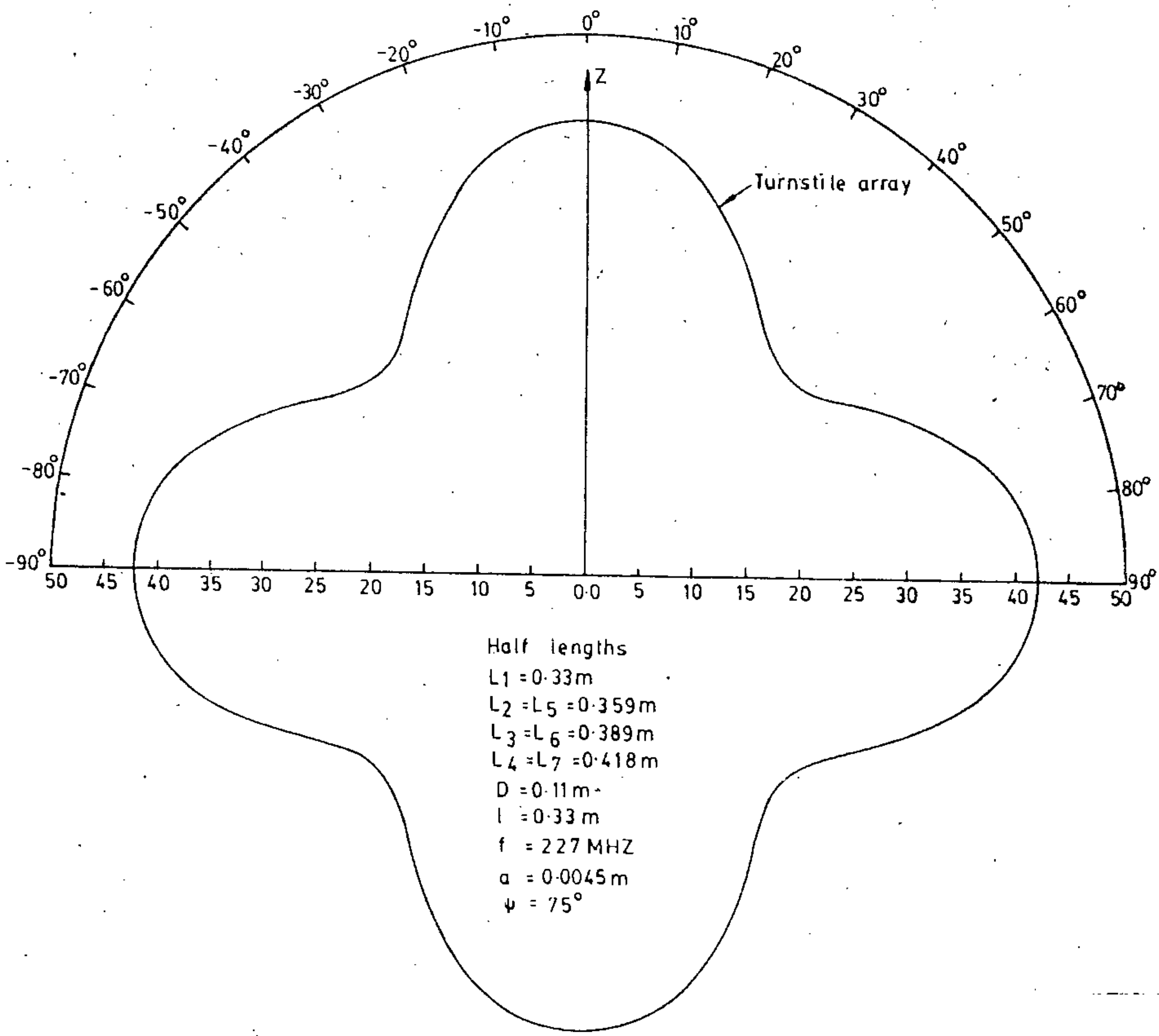

Fig. 4.9 Broadside field pattern of a turnstile antenna when a shorting ring is used in the periphery of the array

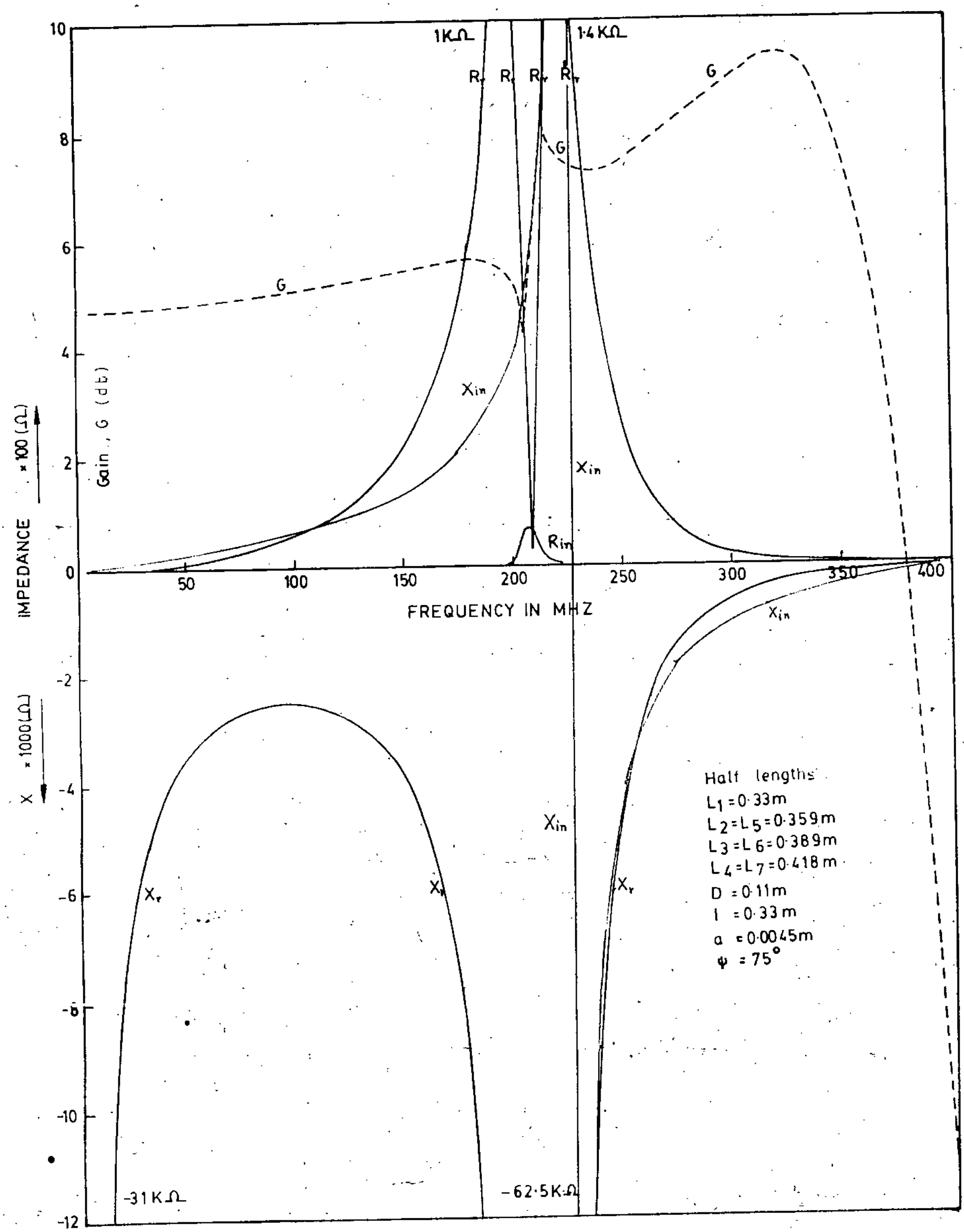

Fig. 4.10 Impedance and gain versus frequency characteristics of a turnstile array with a peripheral ring, Z_{OT}= 300 Ω.

characteristics of turnstile antenna with shorted rings at periphery are discussed in Fig. 4.10. The specified selection and arrangement of radiators give a very high and resonable gain from 5 MHZ to 370 MHZ. The salient feature of this arrangement is the wide band operation with reasonably high radiation resistance.

4.5 Discussion

In this chapter, the characteristics of turnstile antennas, are analyzed theoretically with the aid of transmission line analogy applied to each dipole element. To increase the horizontal plane directivity, sevaral turnstile units called bays can be stacked at about one wave-length intervals in order to obtain a very low SWR over a considerable bandwidth. This arrangement is called a "Superturnstile" antenna. Batwing or turnstile do not have a satisfactory impedance characteristics when mounted in front of a screen or tower face. A disadvantage of the turnstile antenna is the multiplicity of connections required in the antenna. Four points of connection are required at each bay, one to each batwing and hence more for Superturnstile antenna where several bays exist. This multiplicity of connections can result in substantial repair time in case of trouble with the antenna. The study of Superturnstile antenna is limited here upto a single bay turnstile. The superturnstile antenna offers a convenient means of obtaining maximum effective radiated power for VHF television broad cast station. It is possible to reduce the VSWR less than 1.1 across the channel for which it is desired.

It is possible to connect the transmitter output to the two antenna feed lines by several methods. The standard method is to feed the outputs of the aural and visual transmitters through a bridge diplexer. The output from the diplexer is fed through two transmission lines to the antenna. From the analysis of a turnstile array given above it is evident that very high gain with considerable radiation resistance can be obtained by properly designing the antenna. The peripheral ring shorting the tips of the dipole -elements improves the antenna characteristics considerably.

CHAPTER 5

GENERAL DISCUSSION AND CONCLUSION

In the above a theoretical analysis has been carried out for determining the characteristics of a turnstile antenna. The basic unit of any array antenna is the dipole antenna which still requires a rigorous study. As the current distribution of a cylindrical dipole antenna became a controversial issue the theoretical base for a dipole array remains controversial. In the past various numerical techniques were developed for studying the salient features of the current wave on a cylindrical dipole. In most cases it was assumed that the current wave has a null at the dipole-end indicating a perfect end reflection. In the present analysis we have shown that this is not truly a necessary condition. Rather an imperfect reflection can reasonably exist at the dipole ends creating a standing wave pattern of the current on the arms of the dipole. The end reflection can be readily determined. The resulting effect is that a full wave of current stands on the dipole in a compressed manner. The impedance results become modified over those for a pure sinusoidal current distribution. A finite impedance value of the dipole is observed when the electrical length of the dipole is an integer. Other characteristics of the dipole have qualitative agreement with those for a sinusoidal current distribution. One interesting feature is observed that the antenna length-to-radius ratio plays an important role in the end-reflection as well as in the radiation resistance and reactance of the dipole. For a sinusoidal current distribution the radiation resistance remains uneffected with the variation of this ratio, rather the reactance is modified.

Simulation of all the characteristics is performed on the basis of computer programming which involves the rapid calculation of current, self and mutual impedances among the various antenna elements in a turnstile array . The basic arrangement of turnstile antenna is the cross dipole in which two dipoles at right angles are supplied with currents in quadrature phase for a desired omnidirectional pattern. Crossed dipole with equal lengths having equal and unequal magnitude of currents was studied for polarization in the horizontal plane for various electrical lengths. In every case the polarization in the horizontal plane is unsymmetrical except for quadrature phase of currents with identical magnitudes. The dipoles in a cross-dipole or turnstile arrangement are excited by separate feeders connected in parallel and hence a reduction of input impedance is observed. Therefore the gain over a single dipole is increased and the operating bandwidth is also increased. The assumption followed in cross-dipole arrangement is that the feeding system is matched with the antenna so that maximum transfer of energy and low SWR can be realized . The cross-dipole antenna gives an excellent characteristic over several adjoining TV-channels.

The batwing array which is the heart of superturnstile antenna is extensively discussed here with and without using shorted rings at the periphery. It is the first step in forming a highly directive antenna than a dipole. Stacking of dipoles in the linear array is followed by an arbitrary arrangement and every characteristic is followed with this configuration. A specimen study has been produced here and an optimization can be achieved in various ways of changing the geometry of the array.

The impedance of open ended batwing array and its corresponding turnstiling gives a great improvement in radiation resistnace so that more radiation is achievable with a wide band operation of it. Gain is greatly increased and an almost circular or horizontal polarization is achieved.

Shorted-end batwing arrays gives a modified and renovent distribution of current standing on the dipoles. The concept is not outside the desired selection of distribution of current on the elements. The transformation observed in the standing wave current is the abrupt change into a cosine distribution. The field pattern that is observed by the procedure afore-mentioned is tremendously increased so that a dipole and even an array of dipoles forming open-ended batwing or turnstile arrays have an insignificant field strength in comparison with that. The gain and impedance have a tremendous improvement over the arrangement described before so that a broad band operation may be realized. For our specimen study the band is quite broad from 5 MHZ to 400 MHZ .

Everywhere the subject matter is studied within two wavelengths at the required operating frequency. Arrays exceeding two wavelengths are usually suspected to support surface waves that may be subject of study of a future project with or without using mode theory. In this context a superturnstile consisting of several turnstile bays may of considerable research interest. A rigorous treatment by integral equation technique or by moment method for evaluating the current distribution along the dipoles for batwing or turnstile arrays may explore further electro-magnetic phenomena.Finally it can be said that the present study gives a thorough understanding about a turnstile antenna and a study of this nature is not available in public literature.

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C              -------------------------------------
C                        MAIN PROGRAM
C     PROPERTIES OF MULTIELEMENT DIPOLES ARRAY APPROACHING TOWARDS
C     A TURNSTILING BAY
C     NUMBER OF ELEMENTS=7
      EXTERNAL FUNC,FUNC1,FUNC2
      DIMENSION ANT1(28),ANT2(28),D(28),ZA(7,14),ZI(7,7),VD(7),XA(7),
     +Z(7,7),E(28),F(28),ZD(7),ATR(7),ANT(7)
      COMPLEX V,AIG(28),Z12(28),FUNC,DIG(28),Z,ZA,ZI,XA,
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     +CS,S,S1,S2,S3,S4,S5,S6,S7,
     +FUNC1,FUNC2,AIS1(7),AIS2(7),DIS1(7),DIS2(7),R(7),E,F,CII,CJJ,
     +CKK,PK1,PK2,PK3,PK4,PK5,PK6,PK7,QK1,QK2,QK3,QK4,QK5,QK6,QK7,
     +PAT1,PAT2,PAT3,PAT4,PAT5,PAT6,PAT7,CUR1,CUR2,CUR3,CUR4,CUR5,
     +CUR6,CUR7,RE1,RE2,Q12,
     +PATT1,PATT2,PATT3,PATT4,PATT5,PATT6,PATT7,SS1,SS2
     +,SS3,SS4,SS5,SS6,SS7,CSS,SSM,SCK,VI,ZR,ZI5,ZI3,ZI2,ZI2,ZT1,ZIN,
     +CZIN,CZR,CCC,CTT,BBT,ZZ1,ZZ2,ZZ3
      COMMON V, ANTL1,ANTL2,Q12,W,ANTR,ANL,RE1,RE2
      OPEN(UNIT=9,FILE='OUT',STATUS='NEW')
C     DATA ANT1/.33,.33,.33,.33,.33,.33,.33,.503,.503,.503,.503,.503,
C    +.503,.676,.676,.676,.676,.676,.85,.85,.85,.85,.503,.503,.503,.676
C    +,.676,.85/
C     DATA ANT2/.33,.503,.676,.85,.503,.676,.85,.503,.676,.85,.503,.676
C    +,.85,.676,.85,.503,.676,.85,.85,.503,.676,.85,.503,.676,.85,.676
C    +,.85,.85/
C     DATA D/.0045,.33,.66,.99,.33,.66,.99,.0045,.33,.66,.66,.99,1.32,
C    +.0045,.33,.99,1.32,1.65,.0045,1.32,1.65,1.98,.0045,.33,.66,.0045
C    +,.33,.0045/
C     DATA ANT/.33,.503,.676,.85,.503,.676,.85/
      DATA ATR/.0045,.0045,.0045,.0045,.0045,.0045,.0045/
      V=(0.0,1.0)
      PAI=3.1415927
      SEP=0.05
      RAS=0.01
      ZOT=300.0
      ZOO=276.0*ALOG10(SEP/RAS)
      DO 122 NN=1,85
      FR=5.0E06*FLOAT(NN)
C     FR=227.0E06
      FRR=FR/(1.0E06)
      W=2.0*PAI*FR/(3.0E8)
      DD=0.11
      ALP=75.0
      ALPR=PAI*ALP/180.0
      TAA=TAN(ALPR)
      DEL=DD/TAA
      ANT(1)=0.33
      ANT(2)=ANT(1)+DEL
      ANT(3)=ANT(2)+DEL
      ANT(4)=ANT(3)+DEL
      ANT(5)=ANT(2)
      ANT(6)=ANT(3)
      ANT(7)=ANT(4)
      DO 61 M=1,7
61    ANT1(M)=ANT(1)
      DO 106 M=8,13
106   ANT1(M)=ANT(2)
      DO 62 M=14,18
62    ANT1(M)=ANT(3)
      DO 26 M=19,22
26    ANT1(M)=ANT(4)
      DO 63 M=23,25
63    ANT1(M)=ANT(5)
      DO 36 M=26,27
```

```
38    ANT1(M)=ANT(5)
      ANT1(28)=ANT(7)
      DO 64 M=1,7
64    ANT2(M)=ANT(M)
      DO 46 M=8,13
46    ANT2(M)=ANT(M-6)
      DO 55 M=14,18
55    ANT2(M)=ANT(M-11)
      DO 56 M=19,22
56    ANT2(M)=ANT(M-15)
      DO 66 M=23,25
66    ANT2(M)=ANT(M-18)
      DO 67 M=26,27
67    ANT2(M)=ANT(M-20)
      ANT2(28)=ANT(7)
      D(1)=ATR(1)
      D(2)=DD
      D(3)=2.0*DD
      D(4)=3.0*DD
      D(5)=DD
      D(6)=D(3)
      D(7)=D(4)
      D(8)=D(1)
      D(9)=DD
      D(10)=D(3)
      D(11)=D(3)
      D(12)=D(4)
      D(13)=4.0*DD
      D(14)=D(1)
      D(15)=DD
      D(16)=D(4)
      D(17)=D(13)
      D(18)=5.0*DD
      D(19)=D(1)
      D(20)=D(13)
      D(21)=D(18)
      D(22)=6.0*DD
      D(23)=D(1)
      D(24)=DD
      D(25)=D(3)
      D(26)=D(1)
      D(27)=DD
      D(28)=D(1)
      ALL=0.35
      APP1=W*(ALL-D(1))
      APP2=W*(ALL-D(2))
      APP3=W*(ALL-D(3))
      APP4=W*(ALL-D(4))
      APP5=W*(ALL-D(5))
      APP6=W*(ALL-D(6))
      APP7=W*(ALL-D(7))
      VO(1)=SIN(APP1)
      VO(2)=SIN(APP2)
      VO(3)=SIN(APP3)
      VO(4)=SIN(APP4)
```

```
      VO(5)=SIN(APP5)
      VO(6)=SIN(APP6)
      VO(7)=SIN(APP7)
      DO 16 K=1,7
      ANTR=ATR(K)
      ANL=ANT(K)
      ZO(K)=120.0*(ALOG(2.0*ANL/ANTR)-1.0)
      CALL TINTG2(FUNC1,0.0,ANL,AIG1(K),INDER,DIG1(K))
      CALL TINTG2(FUNC2,0.0,ANL,AIG2(K),INDER,DIG2(K))
      R(K)=(V*AIG2(K)+W*(AIG1(K)+ZO(K)/60.0))/((V*AIG2(K))-W*(ZO(K)/
     +60.0+AIG1(K)))
16    CONTINUE
      DO 19 I=1,7
19    E(I)=R(1)
      DO 91 I=8,13
91    E(I)=R(2)
      DO 29 I=14,18
29    E(I)=R(3)
      DO 92 I=19,22
92    E(I)=R(4)
      DO 39 I=23,25
39    E(I)=R(5)
      DO 93 I=26,27
93    E(I)=R(6)
      E(28)=R(7)
      DO 49 I=1,7
49    F(I)=R(I)
      DO 94 I=8,13
94    F(I)=R(I-6)
      DO 59 I=14,18
59    F(I)=R(I-11)
      DO 95 I=19,22
95    F(I)=R(I-15)
      DO 69 I=23,25
69    F(I)=R(I-18)
      DO 96 I=26,27
96    F(I)=R(I-20)
      F(28)=R(7)
      DO 10 I=1,28
      ANTL1=ANT1(I)
      ANTL2=ANT2(I)
      D12=D(I)
      RE1=E(I)
      RE2=F(I)
      P1=W*ANTL1
      Q1=SIN(P1)
      Q11=COS(P1)
      P2=W*ANTL2
      Q2=SIN(P2)
      Q22=COS(P2)
      CII=-V*(1.0-RE1)*Q11+(1.0+RE1)*Q1
      CJJ=-V*(1.0-RE2)*Q22+(1.0+RE2)*Q2
      CKK=CONJG(CJJ)
      Q12=CII*CKK
      CALL TINTG2(FUNC,0.0,ANTL2,AIG(I),INDER,DIG(I))
```

```
      Z12(I)=V*60.0*AIG(I)/(W*E12)
10    CONTINUE
      DO 103 I=1,7
      DO 103 J=1,7
103   Z(J,I)=(0.0,0.0)
      Z(1,1)=Z12(1)
      Z(1,2)=Z12(2)
      Z(1,3)=Z12(3)
      Z(1,4)=Z12(4)
      Z(1,5)=Z12(5)
      Z(1,6)=Z12(6)
      Z(1,7)=Z12(7)
      Z(2,1)=Z12(2)
      Z(2,2)=Z12(8)
      Z(2,3)=Z12(9)
      Z(2,4)=Z12(10)
      Z(2,5)=Z12(11)
      Z(2,6)=Z12(12)
      Z(2,7)=Z12(13)
      Z(3,1)=Z12(3)
      Z(3,2)=Z12(9)
      Z(3,3)=Z12(14)
      Z(3,4)=Z12(15)
      Z(3,5)=Z12(16)
      Z(3,6)=Z12(17)
      Z(3,7)=Z12(18)
      Z(4,1)=Z12(4)
      Z(4,2)=Z12(10)
      Z(4,3)=Z12(15)
      Z(4,4)=Z12(19)
      Z(4,5)=Z12(20)
      Z(4,6)=Z12(21)
      Z(4,7)=Z12(22)
      Z(5,1)=Z12(5)
      Z(5,2)=Z12(11)
      Z(5,3)=Z12(16)
      Z(5,4)=Z12(20)
      Z(5,5)=Z12(23)
      Z(5,6)=Z12(24)
      Z(5,7)=Z12(25)
      Z(6,1)=Z12(6)
      Z(6,2)=Z12(12)
      Z(6,3)=Z12(17)
      Z(6,4)=Z12(21)
      Z(6,5)=Z12(24)
      Z(6,6)=Z12(25)
      Z(6,7)=Z12(27)
      Z(7,1)=Z12(7)
      Z(7,2)=Z12(13)
      Z(7,3)=Z12(18)
      Z(7,4)=Z12(22)
      Z(7,5)=Z12(25)
      Z(7,6)=Z12(27)
      Z(7,7)=Z12(28)
      DO 15 I=1,7
```

```
      DO 15 J=1,7
15    ZA(I,J)=Z(I,J)
      CALL INVERT(ZA,7,ZI)
      CALL MATV(ZI,VD,XA,7)
      VV=VD(1)*VD(1)
      VI=VD(1)*XA(1)+VD(2)*XA(2)+VD(3)*XA(3)+VD(4)*XA(4)+VD(5)*XA(5)+
     +VD(6)*XA(6)+VD(7)*XA(7)
      ZR=VV/VI
      CZR=ZR/2.0
      ZZ1=VD(1)/XA(1)
      ZZ2=VD(2)/XA(2)
      ZZ3=VD(3)/XA(3)
      ZT3=V*ZOT*TAN(W*DD)
      ZI3=(ZZ3*ZT3)/(ZZ3+ZT3)
      ZT2=ZOT*(ZI3*COS(W*DD)+V*ZOT*SIN(W*DD))/(ZOT*COS(W*DD)+
     +V*ZI3*SIN(W*DD))
      ZI2=(ZZ2*ZT2)/(ZZ2+ZT2)
      ZT1=ZOT*(ZI2*COS(W*DD)+V*ZOT*SIN(W*DD))/(ZOT*COS(W*DD)+V
     +*ZI2*SIN(W*DD))
      ZIN=(ZZ1*ZT1/2.0)/(ZZ1+ZT1/2.0)
      CZIN=ZIN/2.0
C     DO 555 MM=1,73
C     THETA=8.0*FLOAT(MM-1)
      THETA=90.0
      ARD=THETA*PAI/180.
      CD=COS(ARD)
      SD=SIN(ARD)
      IF(SD.EQ.0.0)SD=1.0E-05
      IF(CD.EQ.0.0)CD=1.0E-05
      AL1=ANT2(1)*W
      AL2=ANT2(2)*W
      AL3=ANT2(3)*W
      AL4=ANT2(4)*W
      AL5=ANT2(5)*W
      AL6=ANT2(6)*W
      AL7=ANT2(7)*W
      PL1=CD*AL1
      PL2=CD*AL2
      PL3=CD*AL3
      PL4=CD*AL4
      PL5=CD*AL5
      PL6=CD*AL6
      PL7=CD*AL7
      CC1=COS(PL1)
      CC2=COS(PL2)
      CC3=COS(PL3)
      CC4=COS(PL4)
      CC5=COS(PL5)
      CC6=COS(PL6)
      CC7=COS(PL7)
      C1=COS(AL1)
      C2=COS(AL2)
      C3=COS(AL3)
      C4=COS(AL4)
      C5=COS(AL5)
```

```
      C6=COS(AL6)
      C7=COS(AL7)
      A1=CC1-C1
      A2=CC2-C2
      A3=CC3-C3
      A4=CC4-C4
      A5=CC5-C5
      A6=CC6-C6
      A7=CC7-C7
      DK1=SIN(AL1)
      DK2=SIN(AL2)
      DK3=SIN(AL3)
      DK4=SIN(AL4)
      DK5=SIN(AL5)
      DK6=SIN(AL6)
      DK7=SIN(AL7)
      GG1=CD*SIN(PL1)
      GG2=CD*SIN(PL2)
      GG3=CD*SIN(PL3)
      GG4=CD*SIN(PL4)
      GG5=CD*SIN(PL5)
      GG6=CD*SIN(PL6)
      GG7=CD*SIN(PL7)
      BB1=DK1-GG1
      BB2=DK2-GG2
      BB3=DK3-GG3
      BB4=DK4-GG4
      BB5=DK5-GG5
      BB6=DK6-GG6
      BB7=DK7-GG7
      PK1=0.5*(1.0+R(1))
      PK2=0.5*(1.0+R(2))
      PK3=0.5*(1.0+R(3))
      PK4=0.5*(1.0+R(4))
      PK5=0.5*(1.0+R(5))
      PK6=0.5*(1.0+R(6))
      PK7=0.5*(1.0+R(7))
      QK1=0.5*V*(1.0-R(1))
      QK2=0.5*V*(1.0-R(2))
      QK3=0.5*V*(1.0-R(3))
      QK4=0.5*V*(1.0-R(4))
      QK5=0.5*V*(1.0-R(5))
      QK6=0.5*V*(1.0-R(6))
      QK7=0.5*V*(1.0-R(7))
      PAT1=(PK1*A1-QK1*BB1)/SD
      PAT2=(PK2*A2-QK2*BB2)/SD
      PAT3=(PK3*A3-QK3*BB3)/SD
      PAT4=(PK4*A4-QK4*BB4)/SD
      PAT5=(PK5*A5-QK5*BB5)/SD
      PAT6=(PK6*A6-QK6*BB6)/SD
      PAT7=(PK7*A7-QK7*BB7)/SD
      CUR1=XA(1)/(PK1*DK1-QK1*C1)
      CUR2=XA(2)/(PK2*DK2-QK2*C2)
      CUR3=XA(3)/(PK3*DK3-QK3*C3)
      CUR4=XA(4)/(PK4*DK4-QK4*C4)
```

```
      CUR5=XA(5)/(PK5*DK5-QK5*C5)
      CUR6=XA(6)/(PK6*DK6-QK6*C6)
      CUR7=XA(7)/(PK7*DK7-QK7*C7)
      S1=CUR1*PAT1
      S2=CUR2*PAT2
      S3=CUR3*PAT3
      S4=CUR4*PAT4
      S5=CUR5*PAT5
      S6=CUR6*PAT6
      S7=CUR7*PAT7
      PATT1=(PK1*(COS(AL1*SD)-C1)-QK1*(SIN(AL1)-SD*SIN(AL1*SD)))/CD
      PATT2=(PK2*(COS(AL2*SD)-C2)-QK2*(SIN(AL2)-SD*SIN(AL2*SD)))/CD
      PATT3=(PK3*(COS(AL3*SD)-C3)-QK3*(SIN(AL3)-SD*SIN(AL3*SD)))/CD
      PATT4=(PK4*(COS(AL4*SD)-C4)-QK4*(SIN(AL4)-SD*SIN(AL4*SD)))/CD
      PATT5=(PK5*(COS(AL5*SD)-C5)-QK5*(SIN(AL5)-SD*SIN(AL5*SD)))/CD
      PATT6=(PK6*(COS(AL6*SD)-C6)-QK6*(SIN(AL6)-SD*SIN(AL6*SD)))/CD
      PATT7=(PK7*(COS(AL7*SD)-C7)-QK7*(SIN(AL7)-SD*SIN(AL7*SD)))/CD
      SS1=CUR1*PATT1
      SS2=CUR2*PATT2
      SS3=CUR3*PATT3
      SS4=CUR4*PATT4
      SS5=CUR5*PATT5
      SS6=CUR6*PATT6
      SS7=CUR7*PATT7
      CS=S1+S2+S3+S4+S5+S6+S7
      CSS=SS1+SS2+SS3+SS4+SS5+SS6+SS7
      SSM=CSS/CUR1
      S=CS/CUR1
      SCK=S-V*SSM
      PATTC=CABS(SCK)
      PATTN=CABS(S)
C PATTN STANDS FOR PATTERN FACTOR OF THE ARRAY
C PATTC STANDS FOR PATTERN FACTOR OF THE CROSS ARRAY
      RINC=REAL(CZR)
      RIN=REAL(ZR)
      CCC=VI/VD(1)
      CPD=CABS(CCC)
      CTT=CCC-V*CCC
      CON=CABS(CTT)
      PW=CPD*CPD*RIN
C     PWC=CON*CON*RINC
      PWC=CPD*CPD*RINC
      BBB=CABS(CS)
      BBT=CS-V*CSS
      BBC=CABS(BBT)
      GC=120.0*BBB*BBB/PWC
      G=120.0*BBB*BBB/PW
      IF(GC.LE.0.0) GC=0.001
      IF(G.LE.0.0) G=0.001
      GDB=10.0*ALOG10(G)
      GDBC=10.0*ALOG10(GC)
      WRITE(9,601)FRR,ZR,ZIN,CZIN,CZR,GDB,GDBC
  601 FORMAT(/2X,F5.0,1X,8(E12.5,1X),2(F5.2,1X))
  122 CONTINUE
      STOP
      END
```

```
DOS FORTRAN IV 360N-FO-479 3-8          MAINPGM          DATE   19/04/87     TIME    13.41.41       PAGE

         C       SUBROUTINE FOR MAIN PROGRAM
 0001            SUBROUTINE TINTG2(FUZ,X1,X2,AIG,INDER,DIG)
 0002            COMPLEX FUZ,AIG,DIG,TEMP,SUM1,SUM2
 0003            DIMENSION R64(64),W64(64),W32(32)
 0004            DATA R64
                A/.9999824304,.9998728881,.9995987997,.9990981250,.9983166353
                B,.9972962594,.9957241047,.9938319632,.9914957212,.9886847575
                C,.9853714996,.9815311496,.9771415146,.9721626747,.9666378516
                D,.9604912687,.9537300064,.9463428584,.9383203978,.9296548574
                E,.9203400255,.9103711570,.8997448998,.8884592329,.8765134145
                F,.8639079382,.8506444948,.8367259382,.8221562544,.8069035320
                G,.7910849338,.7745956692,.7574839664,.7397560444,.7214230854
                H,.7024962065,.6829874311,.6629096600,.6422766425,.6211029467
                I,.5994039302,.5771957101,.5544951326,.5313197436,.5076877575
                J,.4836180269,.4591300120,.4342437493,.4089798212,.3833593242
                K,.3574038378,.3311353933,.3045764416,.2777498220,.2506787303
                L,.2233866864,.1958975027,.1682352516,.1404242332,.1124889435I
                M,.8445404002E-1,.5634431304E-1,.2818464895E-1,.0/
 0005            DATA W64
                1/.5053609519E-4,.1807395645E-3,.3777466463E-3,.6326073194E-3
                2,.9383698485E-3,.1299524083E-2,.1681142865E-2,.2108815246E-2
                3,.2568764944E-2,.3057753410E-2,.3572892784E-2,.4111503979E-2
                4,.4671050372E-2,.5249123455E-2,.5843449876E-2,.6451900053E-2
                5,.7072489995E-2,.7703375233E-2,.8342339754E-2,.8989275784E-2
                6,.9641177730E-2,.1029711696E-1,.1095573339E-1,.1161572332E-1
                7,.1227583056E-1,.1293483966E-1,.1359157101E-1,.1424487737E-1
                8,.1489364166E-1,.1553677556E-1,.1617321873E-1,.1680193857E-1
                9,.1742193016E-1,.1803221639E-1,.1863184826E-1,.1921990512E-1
                1,.1979549505E-1,.2035775505E-1,.2090585145E-1,.2143898001E-1
                2,.2195636631E-1,.2245726583E-1,.2294095423E-1,.2340677750E-1
                3,.2385405211E-1,.2428215502E-1,.2469052474E-1,.2507856965E-1
                4,.2544576997E-1,.2579162698E-1,.2511567338E-1,.2641747340E-1
                5,.2669662293E-1,.2695274967E-1,.2718551323E-1,.2739460526E-1
                6,.2757974957E-1,.2774373218E-1,.2787725148E-1,.2798921826E-1
                7,.2807645579E-1,.2813884992E-1,.2817631903E-1,.2818881418E-1/
 0006            DATA W32
                1/.3632214818E-3,.1265156556E-2,.2579049795E-2,.4217630442E-2
                2,.6115506822E-2,.8223007957E-2,.1049824691E-1,.1290380010E-1
                3,.1540675047E-1,.1797855157E-1,.2059423392E-1,.2323144664E-1
                4,.2586967933E-1,.2848675475E-1,.3107355111E-1,.3360387715E-1
                5,.3606443279E-1,.3843981025E-1,.4371551012E-1,.4287795003E-1
                6,.4491453165E-1,.4681355499E-1,.4856433041E-1,.5015713931E-1
                7,.5158325395E-1,.5283494679E-2,.5390549934E-1,.5478921050E-1
                8,.5548140543E-1,.5597843651E-1,.5627769983E-1,.5637762836E-1/
 0007            XM=.5*(X1+X2)
 0008            XD=.5*(X2-X1)
 0009            SUM1=(0.,0.)
 0010            SUM2=(0.,0.)
 0011            K=1
 0012            J=0
 0013            DO 10 I=1,63
 0014            DX=R64(I)*XD
 0015            TEMP=FUZ(XM-DX)+FUZ(XM+DX)
 0015            SUM1=SUM1+W64(I)*TEMP
 0017            IF(K.GT.0)GO TO 10
 0018            J=J+1
 0019            SUM2=SUM2+W32(J)*TEMP
 0020      10    K=-K
 0021            TEMP=FUZ(XM)
 0022            SUM1=SUM1+W64(64)*TEMP
 0023            SUM2=SUM2+W32(32)*TEMP
 0024            AIG=XD*SUM1
 0025            DIG=XD*(SUM1-SUM2)
 0026            INDER=0
 0027            RETURN
 0028            END
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                              SCALAR MAP


DOS FORTRAN IV 360N-FO-479 3-8          FUNC            DATE   19/04/87     TIME    13.41.54       PAGE

 0001            COMPLEX FUNCTION FUNC(Z)
 0002            COMPLEX V,A1,A2,A3,F1,F2,F3,F4,RE1,RE2,F5,F6,F7,CJZ,CNJ
 0003            COMMON V,ANTL1,ANTL2,D12,W,ANTR,ANL,RE1,RE2
 0004            PQR=ANTL1-Z
 0005            RR1=D12*D12+PQR*PQR
 0006            R1=SQRT(RR1)
 0007            PW=Z+ANTL1
 0008            RR2=D12*D12+PW*PW
 0009            R2=SQRT(RR2)
 0010            RR3=D12*D12+Z*Z
 0011            R3=SQRT(RR3)
 0012            A1=-V*W*R1
 0013            A2=-V*W*R2
 0014            A3=-V*W*R3
 0015            F1=CEXP(A1)/R1
 0016            F2=CEXP(A2)/R2
 0017            F3=CEXP(A3)/R3
 0018            AKL1=W*ANTL1
 0019            P=COS(AKL1)
 0020            Q=SIN(AKL1)
 0021            F4=(F1+F2-2.0*P*F3)*(1.0+RE1)*W
 0022            F5=PQR*(1.0/RR1+V*W/R1)*F1+PW*(1.0/RR2+V*W/R2)*F2-2.0*W*Q*F3
 0023            F6=V*(1.0-RE1)*F5
 0024            F7=F4+F6
 0025            AR=W*(ANTL2-Z)
 0026            B=SIN(AR)
 0027            C=COS(AR)
 0028            CJZ=-V*(1.0-RE2)*C+(1.0+RE2)*B
 0029            CNJ=CONJG(CJZ)
 0030            FUNC=F7*CNJ
 0031            RETURN
 0032            END
```

DOS FORTRAN IV 360N-FO-479 3-8 FUNC1 DATE 19/04/87 TIME 13.42.06

```
      COMPLEX FUNCTION FUNC1(Z)
      COMPLEX V,A1,A2,A3,F1,F2,F3
      COMMON V,ANTL1,ANTL2,D12,W,ANTR,ANL,RE1,RE2
      ZL1=ANL-Z
      RR1=ANTR*ANTR+ZL1*ZL1
      R1=SQRT(RR1)
      ZL2=ANL+Z
      RR2=ANTR*ANTR+ZL2*ZL2
      R2=SQRT(RR2)
      RRZ=ANTR*ANTR+Z*Z
      RZ=SQRT(RRZ)
      A1=-V*W*R1
      A2=-V*W*R2
      A3=-V*W*RZ
      F1=CEXP(A1)/R1
      F2=CEXP(A2)/R2
      F3=CEXP(A3)/RZ
      P=COS(W*ANL)
      FUNC1=F1+F2-2.0*P*F3
      RETURN
      END
```

DOS FORTRAN IV 360N-FO-479 3-8 FUNC2 DATE 19/04/87 TIME 13.42.15

```
      COMPLEX FUNCTION FUNC2(Z)
      COMPLEX V,A1,A2,A3,F1,F2,F3,H3,H4,H6,H7,H8,H9,H10
      COMMON V,ANTL1,ANTL2,D12,W,ANTR,ANL,RE1,RE2
      ZL1=ANL-Z
      RR1=ANTR*ANTR+ZL1*ZL1
      R1=SQRT(RR1)
      ZL2=ANL+Z
      RR2=ANTR*ANTR+ZL2*ZL2
      R2=SQRT(RR2)
      RR3=ANTR*ANTR+Z*Z
      R3=SQRT(RR3)
      A1=-V*W*R1
      A2=-V*W*R2
      A3=-V*W*R3
      F1=CEXP(A1)
      F2=CEXP(A2)
      F3=CEXP(A3)/R3
      P=ANL*W
      Q=SIN(P)
      H1=2.0*W*Q
      T=R1*R1*R1
      H2=1.0/T
      H3=V*W/RR1
      H4=H2+H3
      U=R2*R2*R2
      H5=1.0/U
      H6=V*W/RR2
      H7=H5+H6
      H8=ZL1*H4*F1
      H9=ZL2*H7*F2
      H10=H1*F3
      FUNC2=H8+H9-H10
      RETURN
      END
```

HIGHEST SEVERITY LEVEL OF ERRORS FOR THIS MODULE WAS 0

DOS FORTRAN IV 360N-FO-479 3-8 MAINPGM DATE 19/04/87 TIME 13.42.23

```
      C       SUBROUTINE TO OBTAIN INVERSE OF A MATRIX
              SUBROUTINE INVERT(ZA,N,ZI)
              DIMENSION ZA(7,14),ZI(7,7)
              COMPLEX ZA,ZI,P
              M=N+N
              M2=N+1
              DO 10 LI=1,N
              DO 10 LJ=M2,M
      10      ZA(LI,LJ)=(0.0,0.0)
              DO 80 K=1,N
              I2=K+N
      80      ZA(K,I2)=(1.0,0.0)
              DO 100 LJ=1,N
              J2=LJ+1
              P=ZA(LJ,LJ)
              DO 40 I=1,M
      40      ZA(LJ,I)=ZA(LJ,I)/P
              DO 100 LK=1,N
              DO 100 LI=J2,M
              IF(LK-LJ)20,100,20
      20      ZA(LK,LI)=ZA(LK,LI)-ZA(LJ,LI)*ZA(LK,LJ)
      100     CONTINUE
              DO 110 I=1,N
              DO 110 J=M2,M
              L3=J-N
      110     ZI(I,L3)=ZA(I,J)
              RETURN
              END
```

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```
      C       SUBROUTINE TO MULTIPLY A MATRIX BY A VECTOR
              SUBROUTINE MATV(ZA,VO,XA,N)
              DIMENSION ZA(7,7),XA(7),VO(7)
              COMPLEX ZA,SUM,XA
              N=7
              DO 9 I=1,N
              SUM=(0.0,0.0)
              DO 11 K=1,N
      11      SUM=SUM+ZA(I,K)*VO(K)
      9       XA(I)=SUM
              RETURN
              END
```

5.	[illegible]	[illegible]	[illegible]E-07	0.51995E+01	0.12216E-07	[illegible]E+01	0.24507E-01	-0.33917E+04	1.73	4.74
10.	0.1949[illegible]+01	[illegible]	0.157[illegible]	0.10414E+02	0.78534E-06	0.5[illegible]E+01	0.97247E-01	-0.16873E+04	1.77	4.78
15.	0.43857E+0[illegible]	-0.26419E+0[illegible]	0.1833[illegible]-04	0.15711E+02	0.91526E-05	0.7[illegible]E+01	0.21943E+00	-0.11209E+04	1.77	4.78
20.	0.76276E+00	-0.15732E+0[illegible]	[illegible]-03	0.21118E+02	0.53065E-04	0.10[illegible]E+02	0.39139E+00	-0.83658E+03	1.78	4.79
25.	0.12259E+[illegible]	[illegible]	[illegible]-03	0.2067[illegible]+02	0.21097E-03	0.13339E+02	0.61443E+00	-0.60501E+03	1.78	4.79
30.	0.17799E+01	[illegible]997E+0[illegible]	[illegible]	[illegible]	0.66318E-03	0.1622[illegible]+02	0.[illegible]E+00	-0.54985E+03	1.79	4.80
35.	0.2439[illegible]+01	-0.35363E+03	[illegible]-02	0.3846[illegible]+02	0.17793E-02	0.1[illegible]+02	0.12197E+01	-0.46692E+03	1.79	4.80
40.	0.3211[illegible]+01	[illegible]827[illegible]+03	[illegible]	0.44607E+02	0.42048E-02	[illegible]	0.16053E+01	[illegible]14E+03	1.80	4.81
45.	0.41012[illegible]01	-0.70953E+03	[illegible]-01	0.51552E+02	0.94122E-02	0.25751E+02	0.20506E+01	-0.35479E+03	1.81	4.82
50.	0.511[illegible]	-0.62971[illegible]	[illegible]-01	0.58827[illegible]	0.19530E-01	0.2741[illegible]+02	0.25569E+01	-0.31465E+03	1.82	4.83
55.	0.62[illegible]E+01	-0.56359E+0[illegible]	[illegible]	0.65736E+02	0.38693E-01	0.333[illegible]E+02	0.31276E+01	-0.28176E+03	1.82	4.84
60.	0.75[illegible]E+01	-0.50762E+03	[illegible]+00	0.75461E+02	0.74079E-01	0.377[illegible]E+02	0.37660E+01	-0.25361E+03	1.84	4.85
65.	0.8[illegible]	-0.45906E+0[illegible]	[illegible]+00	0.93237E+02	0.13643E+00	0.43619E+02	0.44758E+01	-0.22982E+03	1.65	4.86
70.	0.10621E+02	-0.41791[illegible]	[illegible]+00	0.96354E+02	0.25464E+00	0.4819[illegible]+02	0.52[illegible]E+01	-0.20[illegible]E+03	1.66	4.87
75.	0.12247E+02	-0.38120E+03	[illegible]E+00	0.10936E+03	0.46491E+00	0.54681E+02	0.61235E+01	-0.19060E+03	1.68	4.89
80.	0.14137E+02	-0.34859E+03	[illegible]+01	0.12485E+03	0.54937E+00	0.62424E+02	0.70637E+01	-0.17429E+03	1.69	4.90
85.	0.1619[illegible]E+02	-0.31939E+0[illegible]	[illegible]+01	0.14388E+03	0.15668E+01	0.71938E+02	0.8[illegible]E+01	-0.1597[illegible]E+03	1.91	4.92
90.	0.18433E+02	-0.2930[illegible]	[illegible]E+01	0.16810E+03	0.29497E+01	0.84050E+02	0.92163E+01	-0.14[illegible]E+03	1.93	4.94
95.	0.20[illegible]E+02	-0.2692[illegible]	[illegible]	0.20029E+03	0.57463E+01	0.10315E+03	0.10421E+02	-0.134[illegible]E+03	1.95	4.96
100.	0.23426E+02	-0.2475[illegible]	[illegible]+03	0.24527E+03	0.11806E+02	0.12264E+03	0.11713E+02	-0.12379E+03	1.97	4.98
105.	0.26170E+02	-0.2276[illegible]	[illegible]	0.31142E+03	0.26273E+02	0.15571E+03	0.13085E+02	-0.11392E+03	2.00	5.01
110.	0.29057E+02	-0.20983E+03	[illegible]E+03	0.40596E+03	0.65409E+02	0.20448E+03	0.14528E+02	-0.10492E+03	2.02	5.03
115.	0.32050E+02	-0.19344E+03	[illegible]E+03	0.49871E+03	0.17838E+03	0.24936E+03	0.16025E+02	-0.96718E+02	2.05	5.06
120.	0.35095E+02	-0.17855E+03	0.73[illegible]E+03	0.25301E+03	0.35255E+03	0.12651E+03	0.17548E+02	-0.89274E+02	2.08	5.09
125.	0.38112E+02	-0.16509[illegible]+03	[illegible]+03	-0.1394[illegible]E+03	0.23772E+03	-0.69701E+02	0.19056E+02	-0.82[illegible]E+02	2.11	5.12
130.	0.409[illegible]E+02	-0.15301[illegible]	[illegible]+02	-0.20457E+03	0.17698E+03	-0.10[illegible]E+03	0.204[illegible]E+02	-0.7[illegible]E+02	2.13	5.14
135.	0.41365E+02	-0.14203[illegible]	[illegible]+03	-0.12105E+03	0.12251E+03	-0.909[illegible]E+02	0.21732E+02	-0.71117E+02	2.16	5.17

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