# FOLDED DIPOLE ANTENNA

(AMIETE, Dec. 1989, 1978, 76, 74, 72, 64, May, 1991, 1992, 1977, 1970)

A very important variation of conventional half wave dipole is the *Folded dipole* shown in Fig. 9.1 in which two half wave dipoles — one continuous and the other split at the centre — have been folded and joined together in parallel at the ends. The split dipole is fed at the centre by a balanced transmission line. The two dipoles, therefore, have the same voltages at their ends. They are essentially two dipoles in parallel as far as radiation fields are concerned. The radiation pattern of a folded dipole and a conventional half wave dipole is same but the input impedance of the folded dipole is higher. It differs from the conventional dipole mainly in two respects e.g. *directivity and broadness* in band width. The directivity of folded dipole is bi-directional but because of the distribution of currents in the parts of the folded dipole the input impedance becomes higher.

FEED LINE
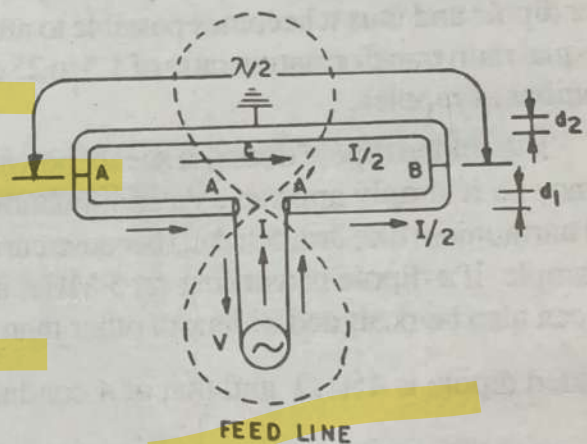
AB = Minimum current points
C = Maximum current point or Minimum voltage point $d_1 = d_2$.
Fig. 9.1 (a). Two wire folded dipole with
(i) Current distribution, (ii) Radiation pattern.

If the radii of the two conductors are equal, then equal currents flow in both the conductors, in the same direction *i.e. currents are equal in magnitude and phase in the two dipoles. Since the total power developed in folded dipole is equal to that developed in the conventional dipole, therefore, the input or terminal impedance of folded dipole is greater than that of the conventional dipole. It can be proved that the input impedance at the terminals of a folded dipole antenna is equal to the square of number of conductors comprising the antenna times the impedance at the terminals of a conventional dipole.*

For, if the total current fed at terminal AA′ is I (say) then the each dipoles will have curren provided their radii are equal. If this had been a straight dipole, the total current would have flowed in the only dipole — arm. Thus, with the same power applied, only half the current flows in first dipole and the input impedance is four times the straight dipole. Because the transmission line "sees" a higher imp i.e. as it is delivering the same power at only half the current. Hence input impedance (or radiation res $R_r = 2^2 \times 73 = 4 \times 73 = 292$ for a folded dipole of equal radii. This shows that a two wire folded can be fed with a conventional 300 Ω open wire transmission line without any matching device. Fur three wire (or tubing) is used for folded (i.e. Tripole) as shown in Fig. 9.2, then only one third of the radiating current would be supplied at the input terminals and hence the input impedance or terminal imped would be nine times the impedance of conventional dipole i.e. $3^2 \times 73 = 9 \times 73 = 657$ Ω. Thus a fol tripole is well suited for matching with a two wire open transmission line of 600 Ω.
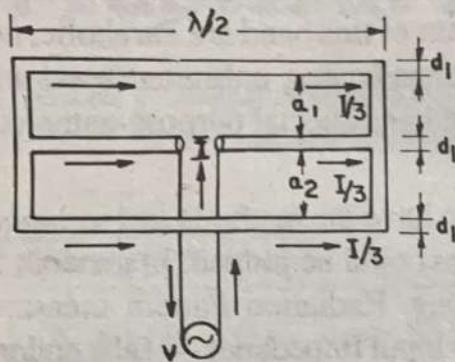

Fig. 9.2. Three wire folded dipole or Tripole with current distribution. The currents in the transmission line is equal and opposite and hence have cancelling effect.
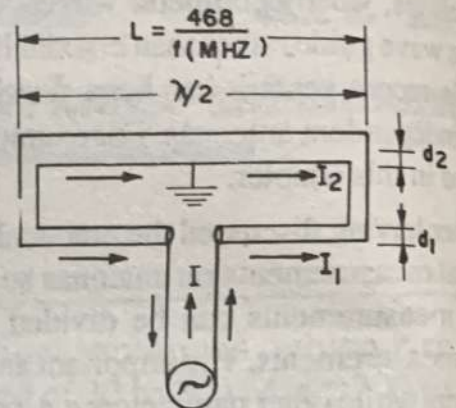

Fig. 9.3. Folded dipole with unequal radii, $I_1 < I_2$ and $I = I_1 + I_2$.

In other words, a folded dipole antenna has a *built in impedance transforming properties*. This ma it easy to match a transmission line that feeds the antenna. Alternatively, instead of changing the number dipole (i.e. element of the antenna), it is also possible to change the input impedance by *keeping the radii two dipoles unequal and so the currents are flowing* as shown in Fig. 9.3. By doing so larger current flow thicker dipole and thus it becomes possible to attain any input impedance that may be desired. With the d of unequal radii transformation ratio of 1.5 to 25 can be achieved and this ratio can still be boost up by incre the number of dipoles.

The folded dipole does not accept power at any even harmonics (i.e. 2nd, 4th etc.) of the fundam frequency so it simply appears as a continuation of the transmission line. However, it works with low S on odd harmonics (like 3rd, 5th etc.) because current distribution of $\lambda/2$ and $3\lambda/2$ antenna is almost sim For example, if a dipole is working on 5 MHz, it will also work on its 3rd harmonics 15 MHz. Further fo dipole can also be designed of length other than $\lambda/2$. It has been found that input impedance of 2 condu $\frac{3}{4}\lambda$ folded dipole is 450 Ω and that of 4 conductors folded dipole $\frac{3}{8}\lambda$ is 225 Ω.

9.2.1. Equation of Input Impedance. The equation for the input impedance or Terminal Impedance or radiation resistance of a folded dipole antenna can be deduced by considering the Fig. 9.1 and drawing its equivalent diagram as shown in Fig. 9.4.

Let V be the emf applied at the antenna terminals AA′. This is being divided equally in each dipole. Hence, voltage in each dipole is V/2 as shown and by nodal analysis

Fig. 9.4. Equivalent diagram of two folded $\frac{1}{2}$ wave dipole

$$\frac{V}{2} = I_1 Z_{11} + I_2 Z_{12}$$

where I_1, I_2 are currents flowing at the terminals of dipole no. 1 and 2 and Z_{11} and Z_{12} are self-imped of dipole no. 1 and mutual impedance between dipole 1 and 2 respectively. But $I_1 = I_2$.
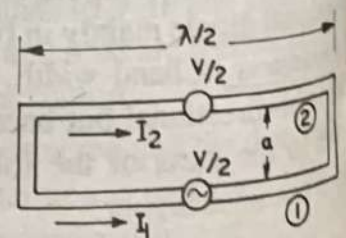
$$\frac{V}{2} = I_1 (Z_{11} + Z_{12})$$

... (9.2)

The two dipoles in the system are very close to each other. The spacing a between two dipoles is of the order of $\lambda/100$.

$$Z_{11} \simeq Z_{12}$$

... (9.3)

Applying condition 9.3 into eqn. 9.2, we have the Terminal Impedance or Input Impedance.

$$\frac{V}{2} = I_1 (2 Z_{11})$$

$$Z = \frac{V}{I_1} = 4 Z_{11} = 4 \times 73 \qquad | \quad \because Z_{11} = 73 \ \Omega \ \text{for a dipole} \quad \text{... (9.4)}$$

$$\boxed{Z = 292 \ \Omega}$$

or

Similarly, for a folded dipole of 3 wires, it can be proved that termination impedance (Fig. 9.2).

$$\frac{V}{3} = I_1 (3 Z_{11})$$
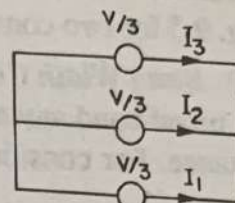
$$\frac{V}{I_1} = Z = 3^2 Z_{11} = 9 \times 73$$

or

$$\boxed{Z = 657 \ \Omega}$$

Fig. 9.4 (b).

Generalizing, we have,

$$\frac{V}{n} = I_1 (n Z_{11}) ; \quad V = I_1 (n^2 Z_{11})$$

$$\frac{V}{I_1} = Z = n^2 Z_{11}$$

or

$$\boxed{Z = n^2 \times 73}$$

... (9.5)

where n is number of half wave dipoles.

Since the impedance transformation is possible by making unequal radii of the two dipoles, hence a formula, in this case, for input impedance can be shown, in general as

$$\boxed{Z = Z_{11} \left(1 + \frac{r_2}{r_1}\right)^2 = 73 \left(1 + \frac{r_2}{r_1}\right)^2}$$

... (9.6)

where r_2 and r_1 = radii of elements.

If $r_2 = 2 r_1$, then $Z = 73 \left(1 + \dfrac{2 r_1}{r_1}\right)^2 = 73 \times 9 = 657 \ \Omega$.

Since the impedance transformation not only depends upon the relative radii of the conductors but also on the relative spacing and hence a formula according to Uda and Mushiake is given by

$$\boxed{Z = Z_{11} \left(1 + \frac{\log \dfrac{a}{r_1}}{\log \dfrac{a}{r_2}}\right)^2 = Z_{11} \cdot Z_{ratio}}$$

... 9.7 (a)

where $\qquad a$ = distance between two conductors

r_1, r_2 = radii of the conductor.

or $\qquad Z_r = ITR = \left(1 + \dfrac{\log a/r_1}{\log a/r_2}\right)^2$ $\qquad$ ...9.10

$\equiv$ Impedance Transformation ratio or Impedance step up ratio.

This is the formula in which *spacing* between conductors and *radius* both are involved on which the I.T.R. depends and now any amount of impedance transformation ratio can be achieved. This method (using unequal radii of conductors) is specially suited when matching is done with low impedance antennas e.g. directive arrays using parasitic elements. Because the radiation resistance of these arrays are quite low. The required ratio of conductors radii to give a desired impedance ratio (Z_r) can be obtained graphically also as shown in Fig. 9.5 for two conductors.

9.2.2 Band Width Compensation. Besides high input impedance and broad band antenna band width, a folded dipole has a built in reactance. For consider, Compensation network shown in Fig. 9.6.
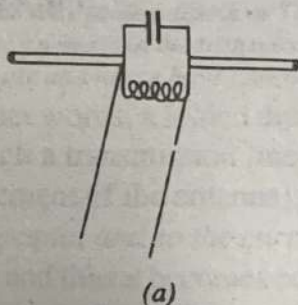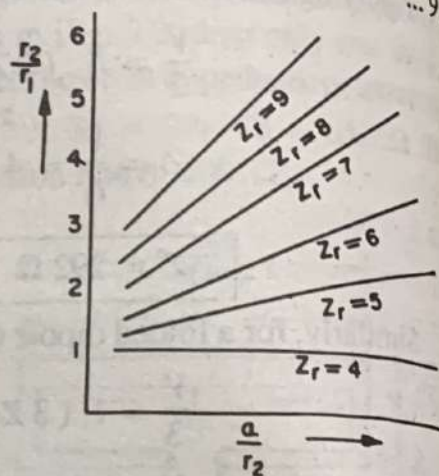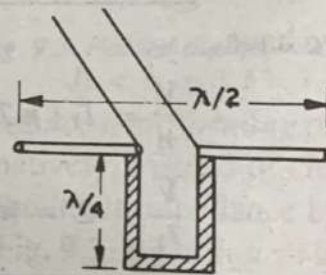
Fig. 9.5. Impedance transformation ratio for two conductors folded dipole as a function of radii and spacing.

(a) $\qquad\qquad\qquad\qquad$ (b)

Fig. 9.6. Compensating networks for increasing band width of half wave dipole using (a) LC cct. (b) $\lambda/4$ transmission line.

The parallel L.C. circuit Fig. 9.6 (a) is resonant at half wave dipole resonant frequency. At this resonant frequency the impedance is high resistive value without affecting the total impedance seen by the feeder. Further above and below the resonant frequency, the antenna reactance is inductive and capacitive while the reactance of LC circuit is capacitive and inductive respectively. Hence, there is tendency to compensate for reactance variations of the antenna over a small frequency band about the resonant frequency. Again, if the Q of the LC circuit is lowered say by introducing a small amount of resistance in it, then, it will provide almost a perfect compensation for the antennas reactance variations over a small frequency band. As shown in Fig. 9.6 (b), a short circuited quarter wave transmission line, connected in parallel with the antenna acts, in similar way like a parallel LC circuit compensations network for reactance variations. A folded dipole provides this type of compensation. Thus for compensation of both resistance and reactance variations over a larger band or frequencies, elaborate lumped constant or transmission line type of circuit is usually used.

As an illustration to built in reactance compensation network of folded dipole antenna Fig. 9.7 may be referred to. A folded dipole may be considered as two short circuited quarter wave transmission line, connected together at point B and is fed in series. The distributions of antenna current (I_A) and transmission line current (I_t) are as indicated in the figure, which shows that antenna current (I_A) is in the same direction and transmission line current (I_t) is in opposite direction when a voltage is applied at the
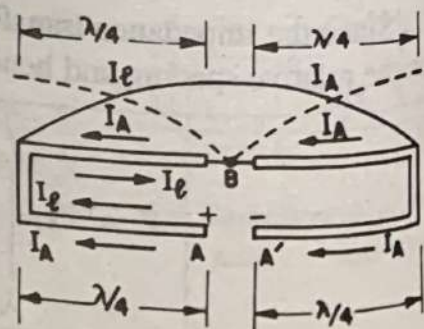
Fig. 9.7. Folded dipole with Antenna Current (I_A) and transmission line current (I_t).

input terminal AA', both the sets of currents flow simultaneously. The antenna current (I_A) being in the same direction contributes to radiation while, the transmission line current being in opposite direction does not contribute to radiation (as cancelled out) but, however, contribute to the input impedance. This may now be considered as the "antenna impedance" of the folded dipole in parallel with reactance of the series connected shorted transmission lines. Thus the reactance variation of the transmission line section tend to compensate for reactance variations of the antenna. This results in more constant impedance of the antenna, around the resonance (or centre) frequency.

It may however, be noted that the folded antenna is of no use at twice the centre frequency (*i.e.* even harmonics). Because the short circuited transmission line sections are each $\lambda/2$ long now, which places a short circuit across the antenna terminals. This renders antenna useless at this frequency. This fact is of importance in connection with television receiving antenna. Further, similarly Yagi-uda antenna is also a broad band antenna as the driven element is almost always a folded dipole.

9.2.3. Uses of Folded Dipole. In conjunction with parasitic elements folded dipole is used in wide band operation such as television. In this, in the Yagi antenna, the driven element is folded dipole and remaining are reflector and director. Reflector is 5% longer than $\lambda/2$ and directors are by 5% smaller. Grounding is made at point B, the mid-point of unbroken arm.

9.2.4. Advantages. It has :

(i) High input impedance.

(ii) Wide band in frequency.

(iii) Acts as built in reactance compensation network.

9.3. YAGI-UDA ANTENNA   (AMIETE, May 1980, 1978, 1971, Dec. 1992, 1972)

Yagi-uda or simply Yagi (as generally but less correctly called) antennas or Yagis are the most high gain antennas and are known after the names of Professor S. Uda and H. Yagi. The antenna was invented and described in Japanese by the former some time around 1928 and afterwards it was described by H. Yagi in English. Since the Yagi's description was in English so it was widely read and thus it became customary to refer this array as Yagi antenna, although he gave full credit to professor Uda. Accordingly a more appropriate name the *Yagi-Uda* antenna is adopted following the practice.

It consists of a driven element, a reflector and one or more directors *i.e.* Yagi-uda antenna is an array of a *driven element* (or active element where the power from the T_x is fed or which feeds received power to the R_x) and one or more *parasitic elements* (*i.e.* passive elements which are not connected directly to the transmission line but electrically coupled). The driven elements is a resonant half-wave dipole usually of metallic rod at the frequency of operation. The parasitic elements of continuous metallic rods are arranged parallel to the driven element and at the same line of sight level. They are arranged collinearly and close together as shown in Fig. 9.8 with one reflector and one director. The optical equivalent is also shown.
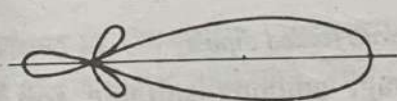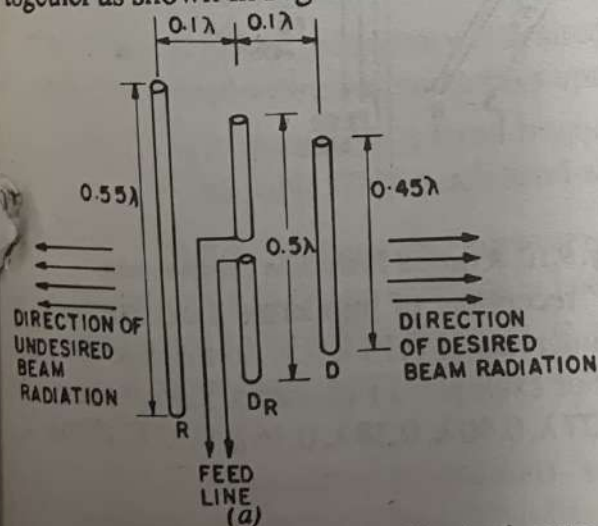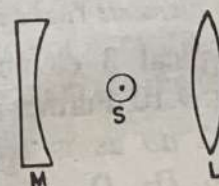

Fig. 9.8 (a) Yagi-uda antenna, (b) Its radiation pattern, (c) Its optical equivalent
R = Reflector (Parasitic element) ; D_R = Driven element ; D = Driector (Parasitic element).

The parasitic elements receive their excitation from the voltages induced in them by the current in the driven element. The phase and currents flowing due to the induced voltage depend on the spacing between the elements and upon the reactance of the elements (i.e., length). The reactance may be varied by dimension the length of the parasitic element. The spacing between driven and parasitic elements that are usually in practice, are of the order of $\lambda/10$ i.e. 0.10λ to 0.15λ. The parasitic element in front of driven, element is known as *director* and its number may be more than one, whereas the element in back of it is known *reflector*. Generally both directors and reflectors are used in the same antenna. The reflector is 5% more director is 5% less than the driven element which is $\lambda/2$ at resonant frequency. In practice, for 3-element of Yagi antenna the following formulae gives lengths which work satisfactorily.

$$\text{Reflector length} = \frac{500}{f\,(\text{MHz})} \text{ feet}$$

...9.8

$$\text{Driven element length} = \frac{475}{f\,(\text{MHz})} \text{ feet}$$

...9.8

$$\text{Director length} = \frac{455}{f\,(\text{MHz})} \text{ feet}$$

...9.8

Eqn. 9.8 provides average length of Yagi antenna determined experimentally for elements of length diameter ratio of 200 to 400 and spacing from 0.10λ to 0.20λ. The parasitic elements can be clamped metallic support rod because at the middle of each parasitic element, the voltage is minimum i.e. there is a voltage node. Even driven element may also be clamped if it is shunt feed. The clamping over the support rod makes a rigid mechanical structure.

Further use of parasitic elements in conjunction with driven element causes the dipole impedance fall well below 73 Ω. It may be as low as 25 Ω and hence it becomes necessary to use either shunt feed folded dipole so that input impedance could be raised to a suitable value, to match the feed cable. While folded dipole the continuous rod may also be clamped to the support as shown in Fig. 9.9.
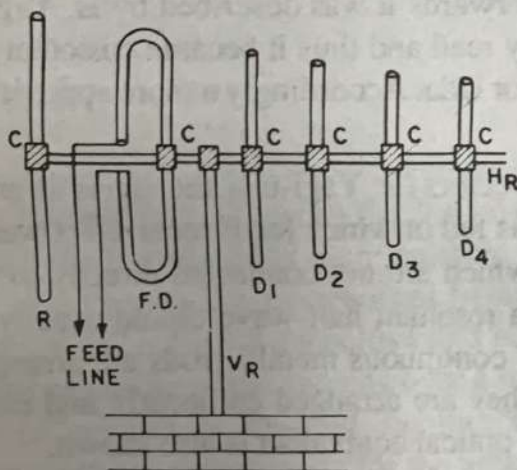
R = Reflector
FD = Folded dipole
$D_1 D_2 D_3 D_4$ = Directors
VR = Vertical rod to support horizontal rods
HR = Horizontal rod to support elements
C = clamps

Fig. 9.9. 6 Elements Yagi antenna with folded dipole.
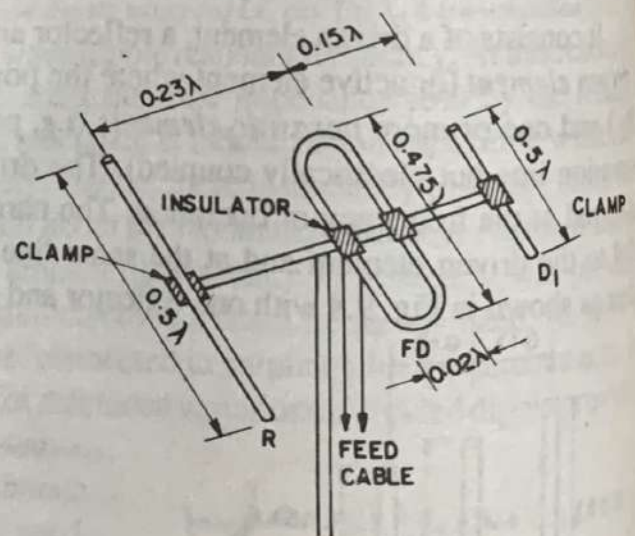
Fig. 9.10. A typical Television Yagi Antenna.

A typical 3 elements yagi antenna suitable for TV reception of moderate field strength shown in Fig. 9.10. Further addition of directors can be done at intervals of 0.15λ i.e. to increase the g even upto 12 db as is required in for fringe area reception. For example, 11 elements Yagi antenna lengths of $D_2, D_3, D_4, D_5, D_6, D_7, D_8, D_9$ are respectively $0.427\lambda, 0.40\lambda, 0.38\lambda, 0.36\lambda, 0.32\lambda, 0.304$ and 0.29λ.

9.3.1. Action. The spacing between elements and the lengths of the parasitic elements determine the phases of the currents. Parasitic antenna in the vicinity of radiating antenna is used either to reflect or to direct the radiated energy so that a compact directional antenna system could be obtained.

A parasitic element of equal or greater length than $\lambda/2$, will be inductive while elements of length less than $\lambda/2$, will be capacitive. Hence, phases of the currents in the former case (i.e. length > $\lambda/2$), will lag the induce voltage whereas in latter case (i.e. length < $\lambda/2$) will lead the induced voltage. Properly spaced dipoles shorter than $\lambda/2$ acts as director and add the fields of driven element in the direction away from the driven element. If more than one directors are employed, then each director will excite the next. On the other hand an element of length equal or greater than $\lambda/2$ acts as reflector and add up the fields of driven element in the direction from reflector towards driven element, if properly spaced.

Additional gain is achieved by using additional directors in the beam direction. The distance between two elements may range from 0.1λ to 0.3λ, close spacing of elements are used in parasitic arrays to get a good excitation. Addition of additional director (than one in 3 elements array) must be adjusted to achieve maximum gain. The greater the distance between driven and director elements, the greater the capacitive reactance needed to provide correct phasing of parasitic current. Therefore, the length of rod is tapered off to achieve the capacitive reactance instead.

The driven element radiates from front to rear (i.e. from reflector to directors).

Part of this radiation induces current in the parasitic element(s) which in turn re-radiate virtually all the radiation. By suitable dimensioning the lengths of parasitic elements and spacing between two elements, the radiated energy is added up in front and tend to cancel the backward radiation. If the distance between driven and parasitic element is decreased, then it will load the driven element, irrespective of its length. Thus input impedance at the input terminals of driven element reduces. This is why a folded dipole is invariably used as driven element so that reduction in input impedance is compensated i.e. raised.

9.3.2. General Characteristic : -

1. If three elements array (i.e. one reflector, one driven and one director) is used, then such type of Yagi-uda antenna is generally referred to as **beam antenna.**

2. It has unidirectional beam of moderate directivity with light weight, low cost and simplicity in feed system design.

3. With spacing of 0.1λ to 0.15λ, a frequency band width of the order of 2% is obtained.

4. It provides *gain* of the order of 8 *db* or **front to** *back ratio* of about 20 *db*.

5. It is also known as *super directive or super gain antenna* (as sometimes called) due to its high gain and beam-width per unit area of the array. An antenna or array which provides directive gain, appreciable greater than that obtainable from *uniform distribution* is known as *super directive or super gain antenna.*

6. If greater directivity is desired, further elements may be used. For example, five or six elements are used with ease, and arrays upto 40 can be constructed.

7. It is essentially a fixed frequency device i.e. frequency sensitive and a band width of about 3% is obtainable. This much band width is sufficient for television reception.

9.3.3. Voltage and current relations in Parasitic Antennas. One or more passive elements coupled magnetically to driven element (PRT) is known as parasitic antenna. The presence of parasitic element effects the directional pattern. The effect on the directional pattern produced depends upon the magnitude and phase of the induced current in the parasitic elements i.e. on the *spacing* of the antenna and tuning of the parasitic antenna.

The quantitative relations between voltages and currents of an antenna system involving parasitic antennas can be given by considering the general equation.

$$V_1 = I_1 Z_{11} + I_2 Z_{12} + I_3 Z_{13} + \ldots + I_n Z_{1n} \qquad \ldots 9.9\,(a)$$
$$V_2 = I_1 Z_{21} + I_2 Z_{22} + I_3 Z_{23} + \ldots + I_n Z_{2n} \qquad \ldots 9.9\,(b)$$
$$V_3 = I_1 Z_{31} + I_2 Z_{32} + I_3 Z_{33} + \ldots + I_n Z_{3n} \qquad \ldots 9.9\,(c)$$

$$\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots$$

$$V_n = I_1 Z_{n1} + I_2 Z_{n2} + I_3 Z_{n3} + \ldots + I_n Z_{nn} \qquad \ldots 9.9\,(d)$$

$V_1, V_2, V_3 \ldots V_n$ = Voltage applied to antenna no. 1, 2, 3, ... n.

$I_1, I_2, I_3 \ldots I_n$ = Current flowing in antenna no 1, 2, 3, ... n.

$Z_{11}, Z_{22}, Z_{33} \ldots Z_{nn}$ = Self–impedances of antenna no. 1, 2, 3, ... n.

$Z_{12}, Z_{21}, Z_{13} \ldots Z_{31}$ = Mutual impedances between antennas of subscript i.e. Z_{1n} means

mutual impedance between antenna no. 1 and n.

If the individual antennas are not excited, then corresponding applied voltages e.g. V_1, V_2, V_3, etc. are zero. Thus in an antenna system involving parasitic antennas, the voltages are zero in case of transmitting while in receiving case these applied voltage are the voltages induced in each parasitic antenna by the electromagnetic waves.

Now considering the simplest case with one driven element (subscript 1) and one parasitic antenna (subscript 2) as shown in Fig. 9.11.
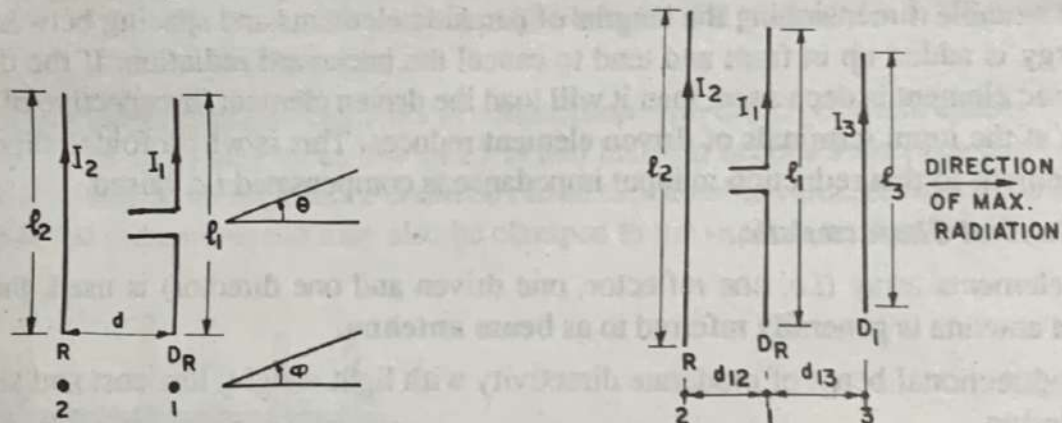
(a) Driven element with one parasitic.   (b) Driven element with two parasitic.

Fig. 9.11. φ = Horizontal plane.

θ = Vertical plane.

From general eqn. 9.9, we can write

$V_1 = I_1 Z_{11} + I_2 Z_{12}$; $V_2 = I_1 Z_{21} + I_2 Z_{22}$ $\because Z_{12} = Z_{21}$; $Z_{13} = Z_{31}$ etc. and $V_2 = 0$ being parasitic

$$\therefore \quad V_1 = I_1 Z_{11} + I_2 Z_{12} \qquad \ldots 9.10\,(a)$$

$$0 = I_1 Z_{12} + I_2 Z_{22} \qquad \ldots 9.10\,(b)$$

or $I_1 Z_{12} = -I_2 Z_{22}$; $I_2 = -I_1 \left(\dfrac{Z_{12}}{Z_{22}}\right)$ $\ldots (9.11)$

$$I_2 = -I_1 \left| \frac{(R_{12} + j\,X_{12})}{(R_{22} + j\,X_{22})} \right| \begin{array}{l} \angle \tan^{-1} \dfrac{X_{12}}{R_{12}} \\[2mm] \angle \tan^{-1} \dfrac{X_{22}}{R_{22}} \end{array}$$

Putting eqn. 9.11 in 9.10 (a), we have

$$V_1 = I_1 Z_{11} - I_1 \left(\frac{Z_{12}}{Z_{22}}\right) \cdot Z_{12} = I_1 \left(Z_{11} - \frac{Z_{12}^{\,2}}{Z_{22}}\right)$$

or

$$I_1 = \frac{V_1}{Z_{11} - \dfrac{Z_{12}^2}{Z_{22}}}$$

... (9.12)

and

$$I_2 = -I_1 \frac{Z_{12}}{Z_{22}} = -\frac{V_1}{\left(Z_{11} - \dfrac{Z_{12}^2}{Z_{22}}\right)} \left(\frac{Z_{12}}{Z_{22}}\right) = \frac{-V_1(Z_{12})}{Z_{11} Z_{22} - Z_{12}^2}$$

$$I_2 = \frac{V_1}{\left(Z_{12} - \dfrac{Z_{11} Z_{22}}{Z_{12}}\right)}$$

... (9.13)

From eqns. 9.12 and 9.13, the input impedances of driven and parasitic elements are given by

$$Z_1 = \frac{V_1}{I_1} = Z_{11} - \frac{Z_{12}^2}{Z_{22}}$$

... (9.14)

and

$$Z_2 = \frac{V_2}{I_2} = Z_{12} - \frac{Z_{11} Z_{22}}{Z_{12}}$$

... (9.15)

Eqns. 9.14 and 9.15 indicate that presence of parasitic elements modifies input impedance of driven elements as the mutual impedance term Z_{12} exists in the eqn. 9.14. Also the input impedance of parasitic element, besides other factor, is also dependent on self-impedance (Z_{11}) of driven element. Further the field distribution of the antenna system is obtained by assuming a constant current in the driven element and calculating the magnitude and phase of current in the parasitic element. The field pattern E_o (φ) in the horizontal plane is given by

$$E_\theta (\varphi) = k (I_1 + I_2/\beta d \cos \varphi)$$

... (9.16)

where k is a constant and I_1 and I_2 are eqns. 9.12 and 9.13. A compact directional pattern is obtained due to the fact that parasitic antenna close to driven antenna is used either to reflect (by parasitic reflector) or to direct (by parasitic director) the radiated energy in the desired direction.

A still another typical values of Yagi–Uda antenna are $l_1 = 0.5\,\lambda$, $l_2 = 0.58\,\lambda$ and $l_3 = 0.45\,\lambda$ and $d_{12} = d_{13} = 0.1\,\lambda$. For reflector, usually a spacing $\lambda/4$ is kept. However, for director a spacing $\lambda/4$ is less effective and hence a spacing of $0.15\,\lambda$ is commonly used. In practice, the effective impedance at the centre is achieved by cutting the length of parasite more or less (about 5%) than a $\lambda/2$ length.

9.3.4. *Adjustments.* The most convenient way to adjust the parasitic antennas, for transmitting case, is to excite the driven element. Place a receiver at a convenient distance in the desired direction. Now vary the tuning of the parasitic antenna by cut and try until the best results are achieved.

For the array for receiving case, place a transmitter of small power at some convenient distance in the undesired direction. Then adjust the parasitic antenna by cut and try until a minimum response is indicated in the receiver associated with receiving antenna.