

[54] MULTITURN LOOP ANTENNA

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[52] U.S. Cl. **343/708, 343/718, 343/744, 343/745, 343/787, 343/846, 343/895**

[51] Int. Cl. **H01d 11/12**

[58] Field of Search 343/705, 708, 895, 908, 343/867, 718, 744, 745, 787, 846

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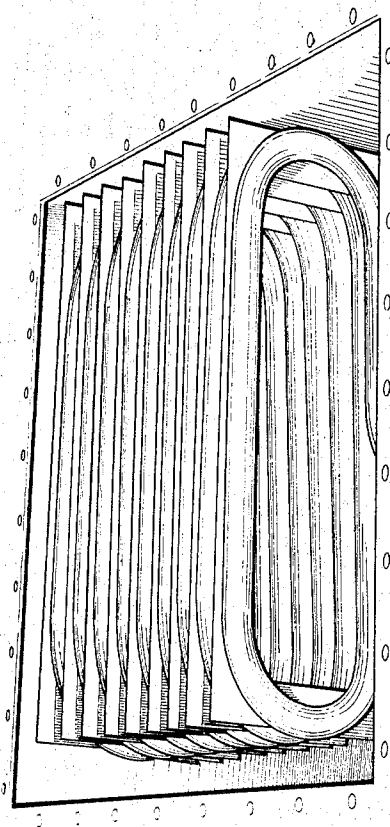
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[57] ABSTRACT

The invention is a multiturn loop antenna of an efficient design suitable for use singly, in arrays, or for inductively exciting radiating currents on a structure — the surface and the elements radiating to achieve optimum system performance. The antenna may be fed balanced or unbalanced and its input impedance may be either capacitive or inductive. In a preferred embodiment for omnidirectional coverage the antenna is positioned in a dielectric or ferrite filled cavity; the antenna may be a single element for linear polarization or a pair of multiturn loops at right angles to each other and with 90° phasing to effect circular polarization.

21 Claims, 35 Drawing Figures



PATENTED JUL 9 1974

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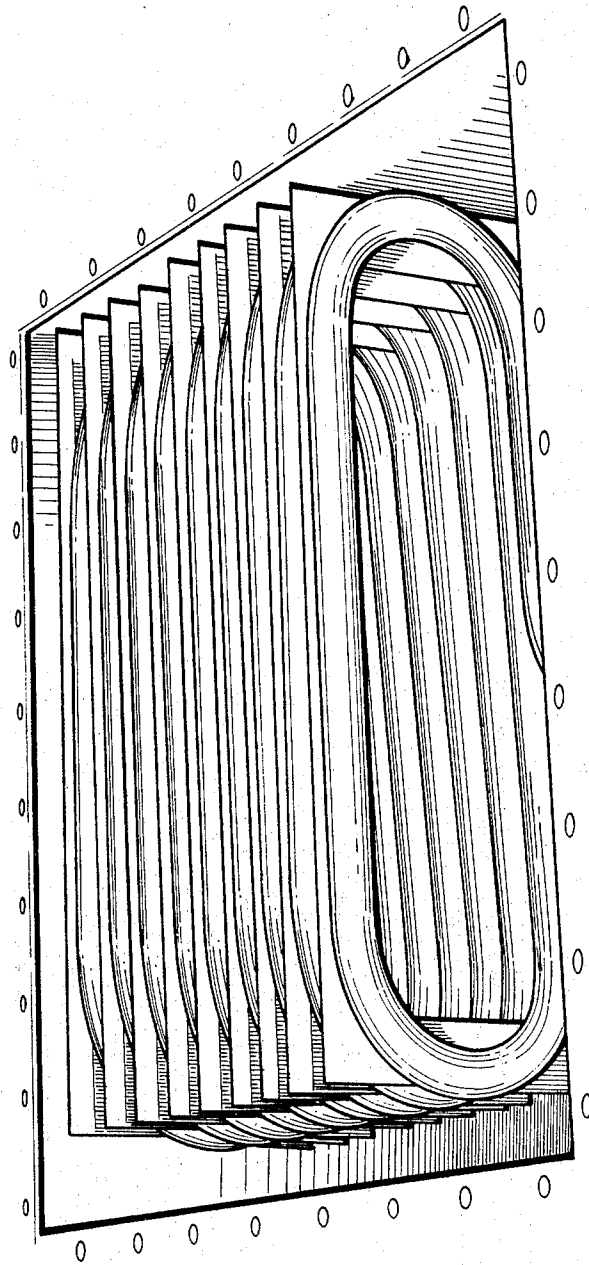


FIG. 1

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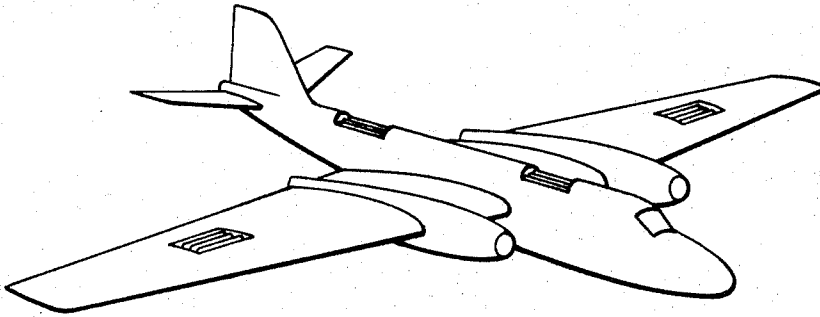


FIG. 2

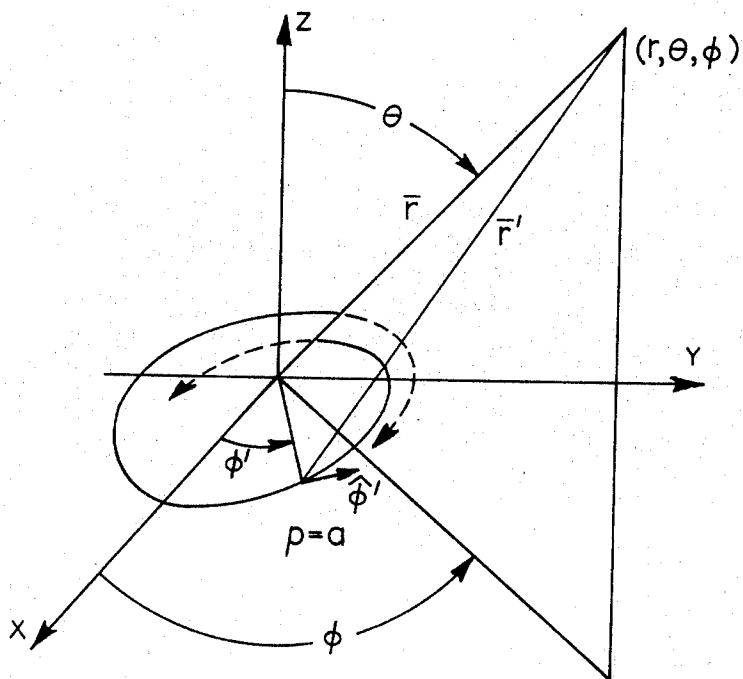


FIG. 3

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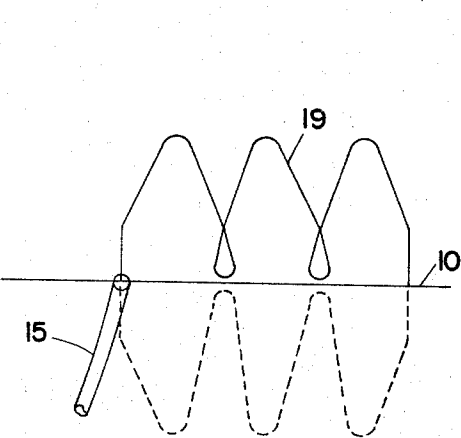


FIG. 4

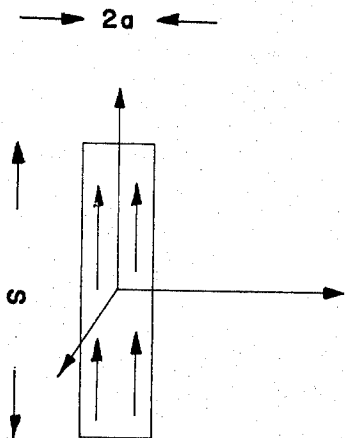


FIG. 5

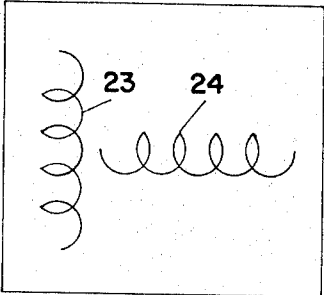


FIG. 6A

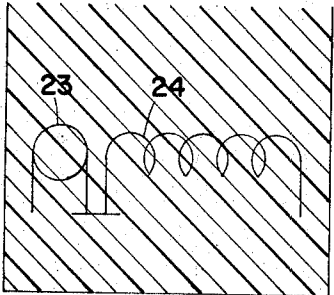


FIG. 6B

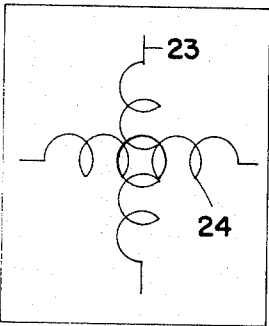


FIG. 6C

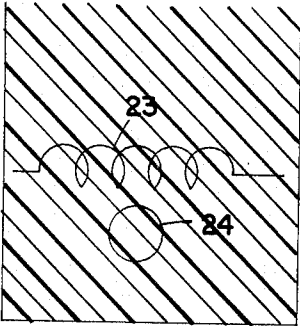
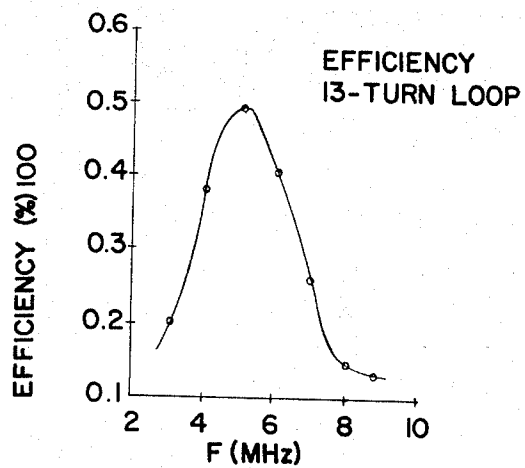
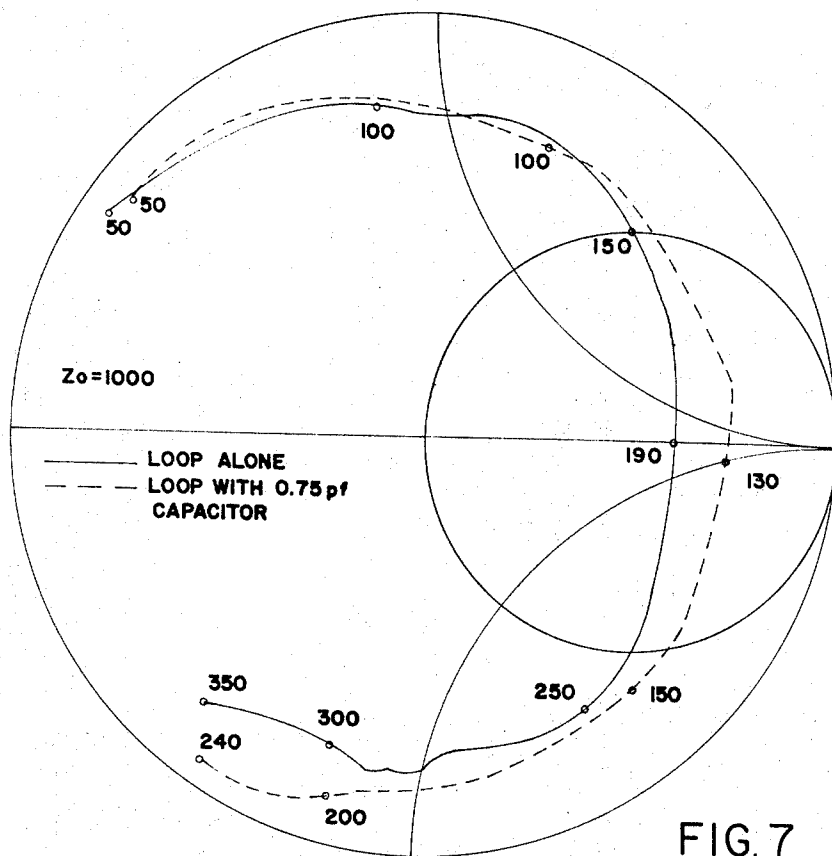


FIG. 6D

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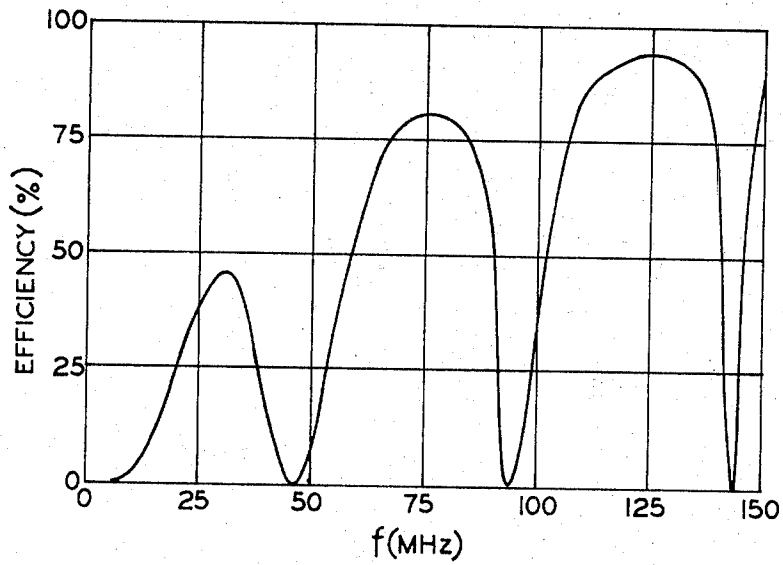


FIG. 8

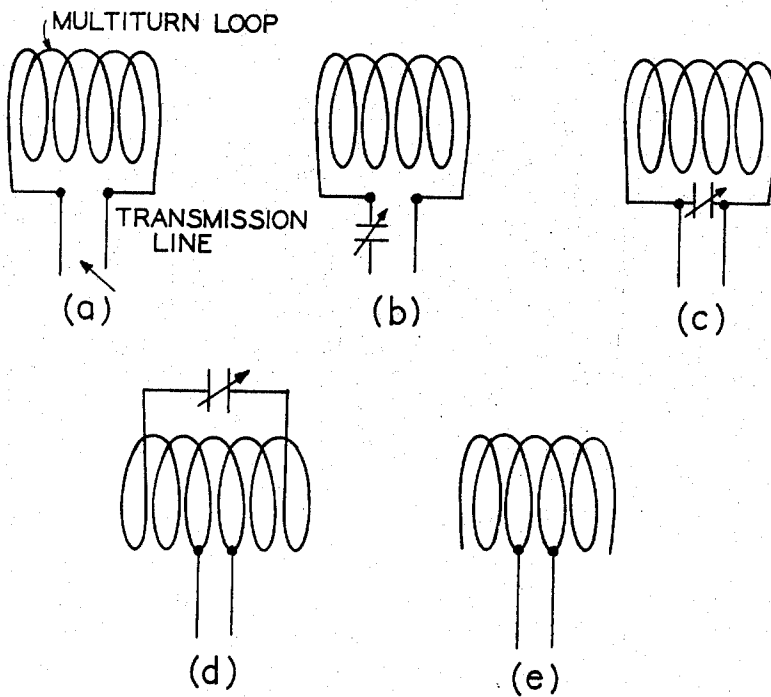


FIG. 9

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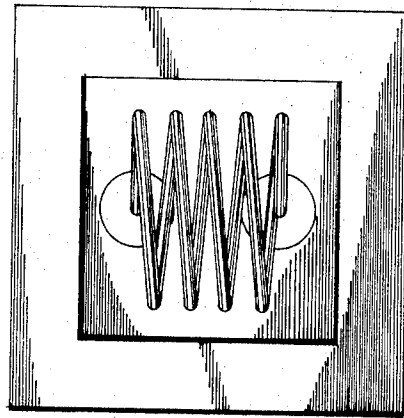


FIG. 10

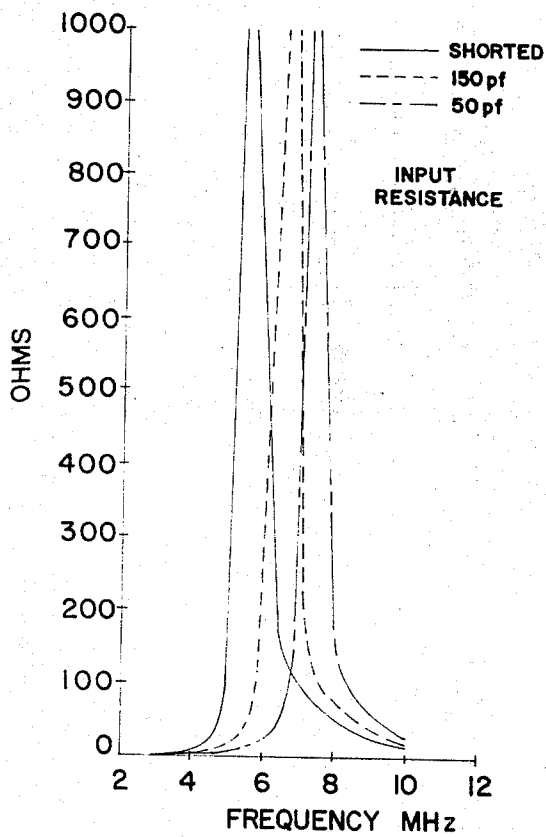


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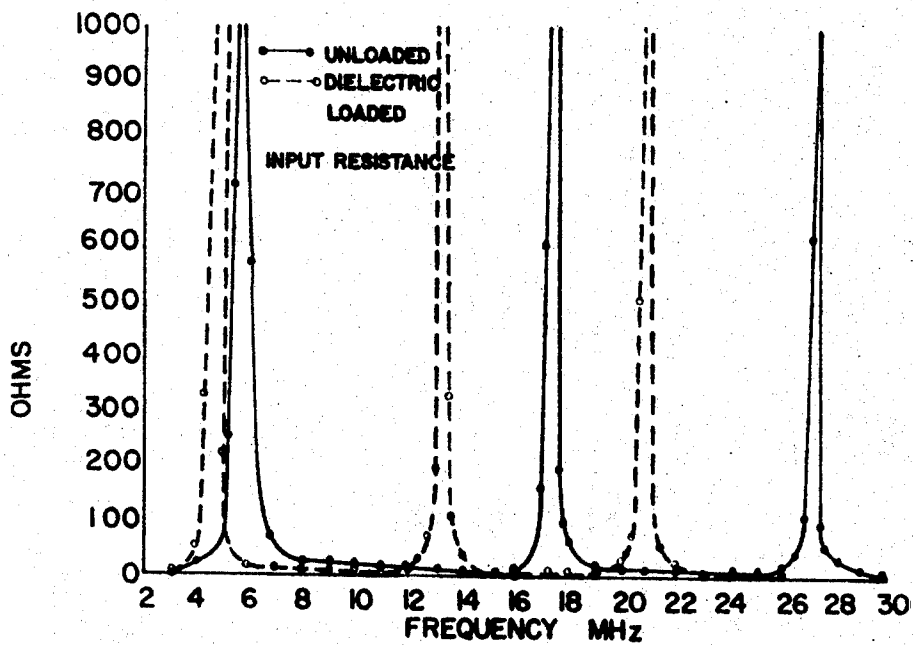


FIG. 11

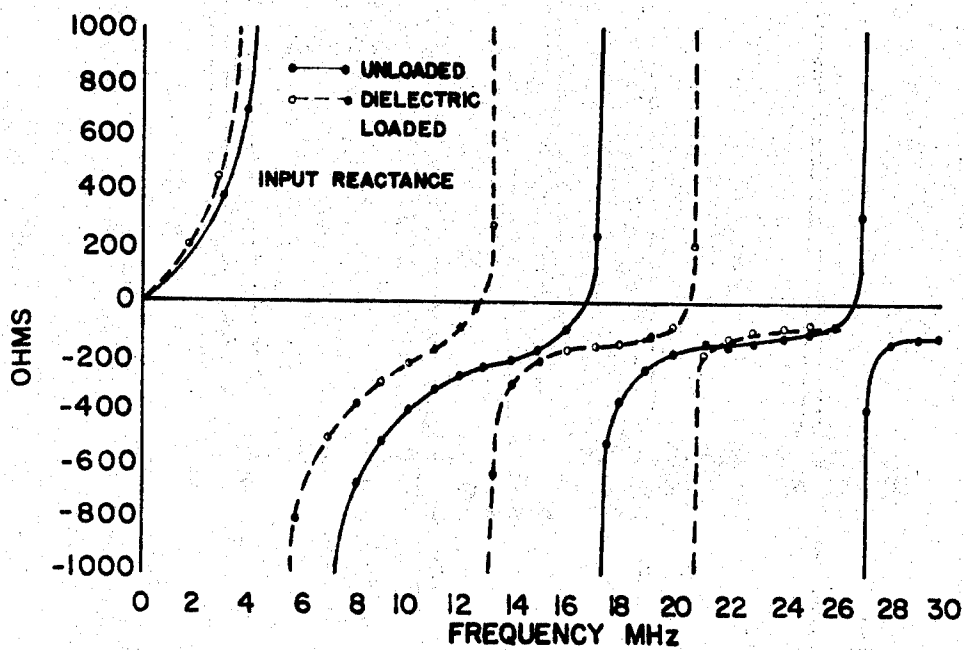


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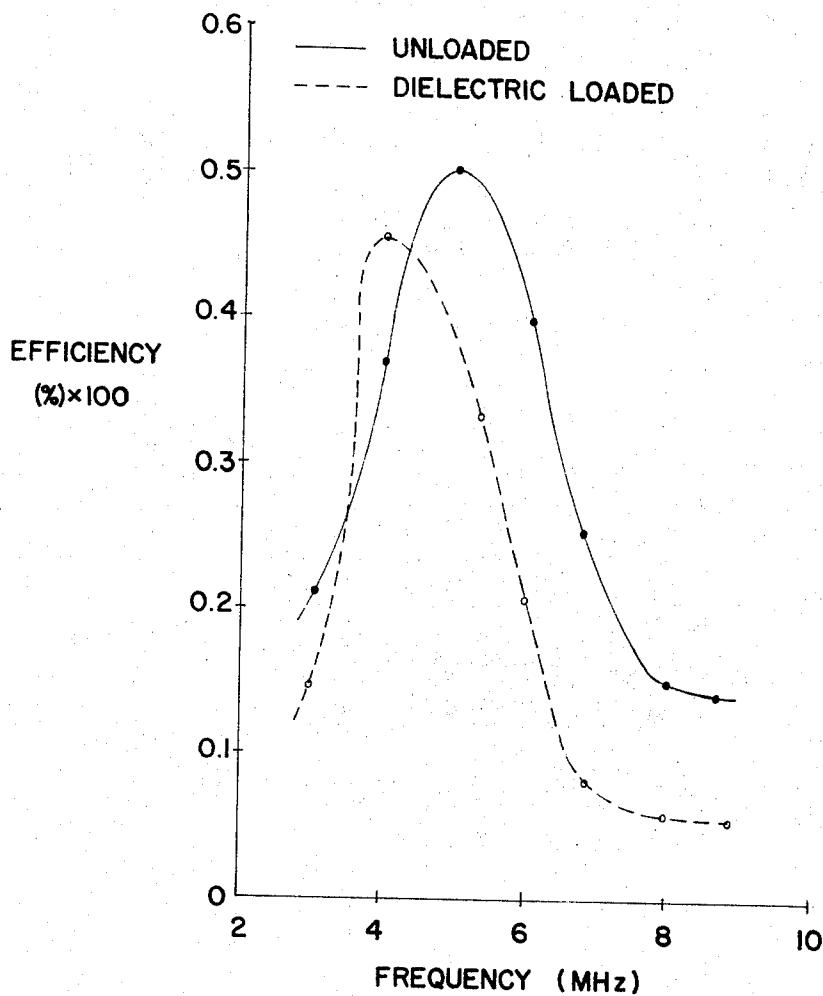


FIG. 13

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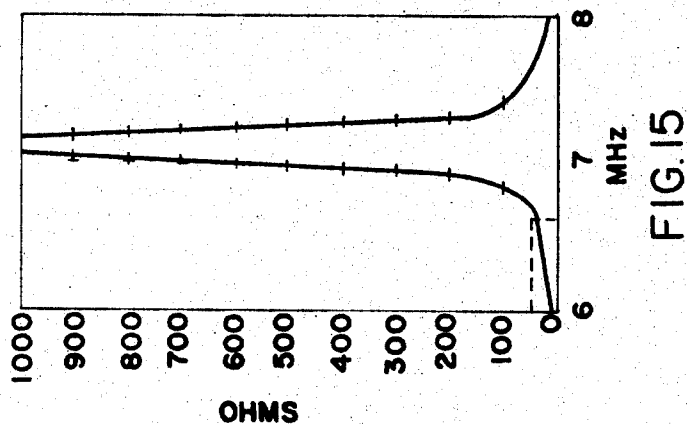


FIG. 15

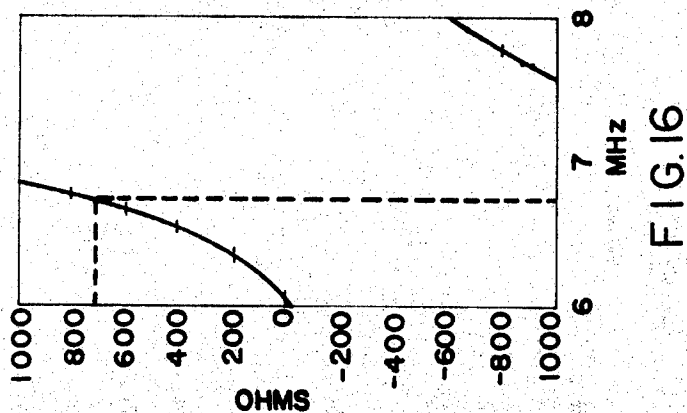


FIG. 16

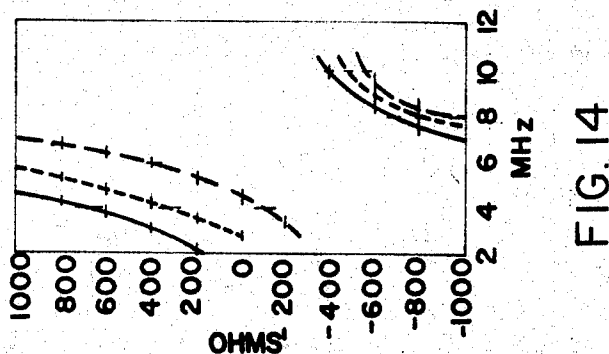


FIG. 14

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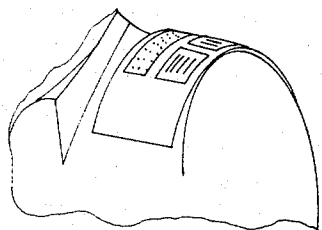


FIG. 17A

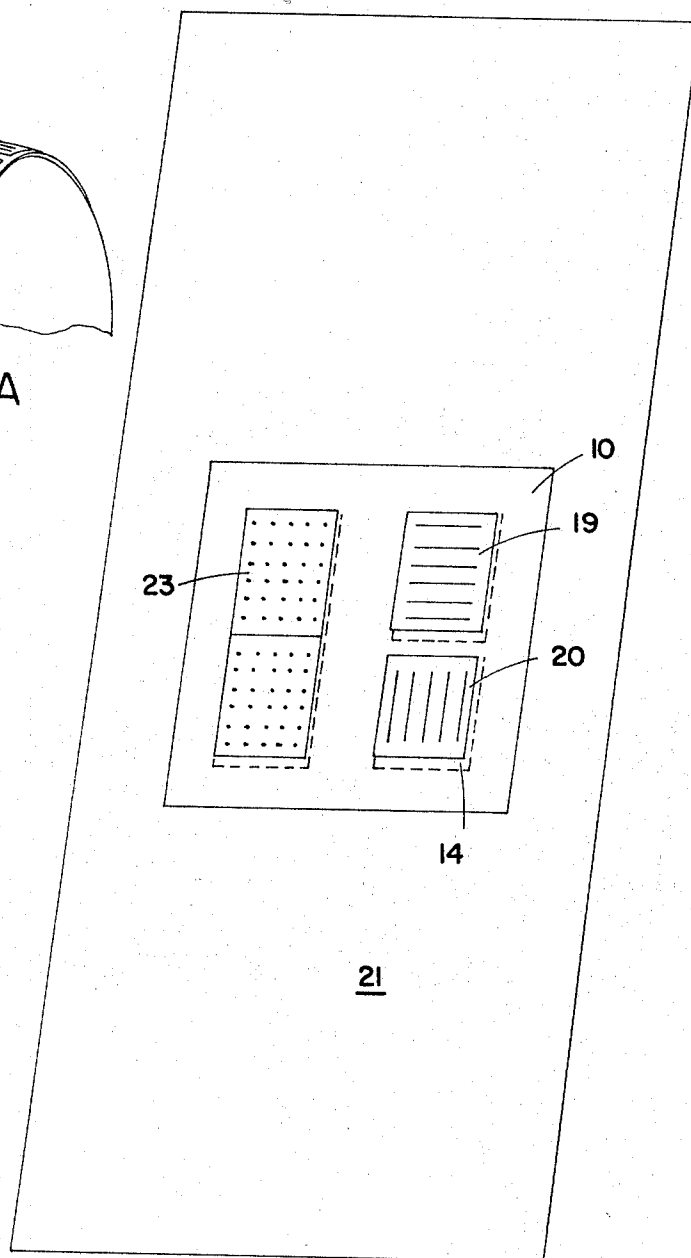


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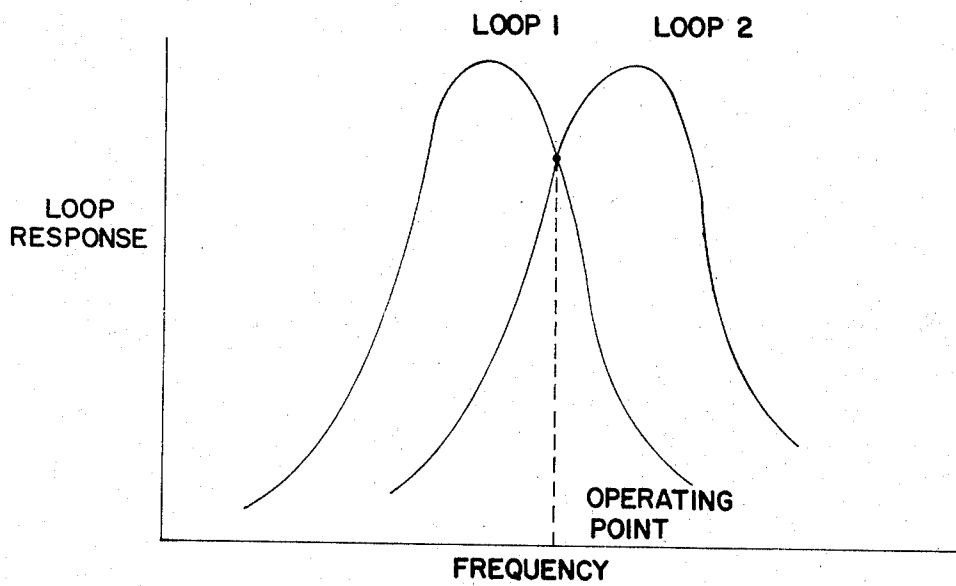


FIG. 18

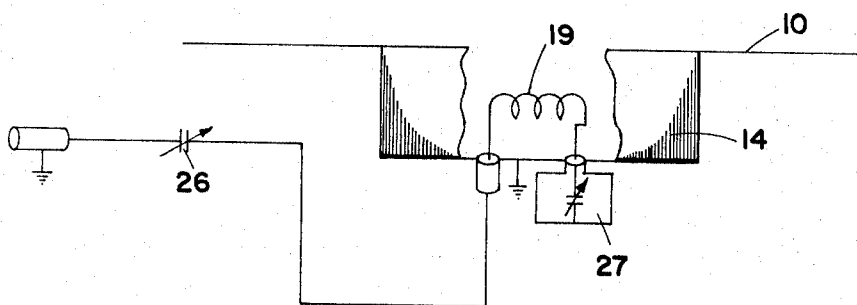


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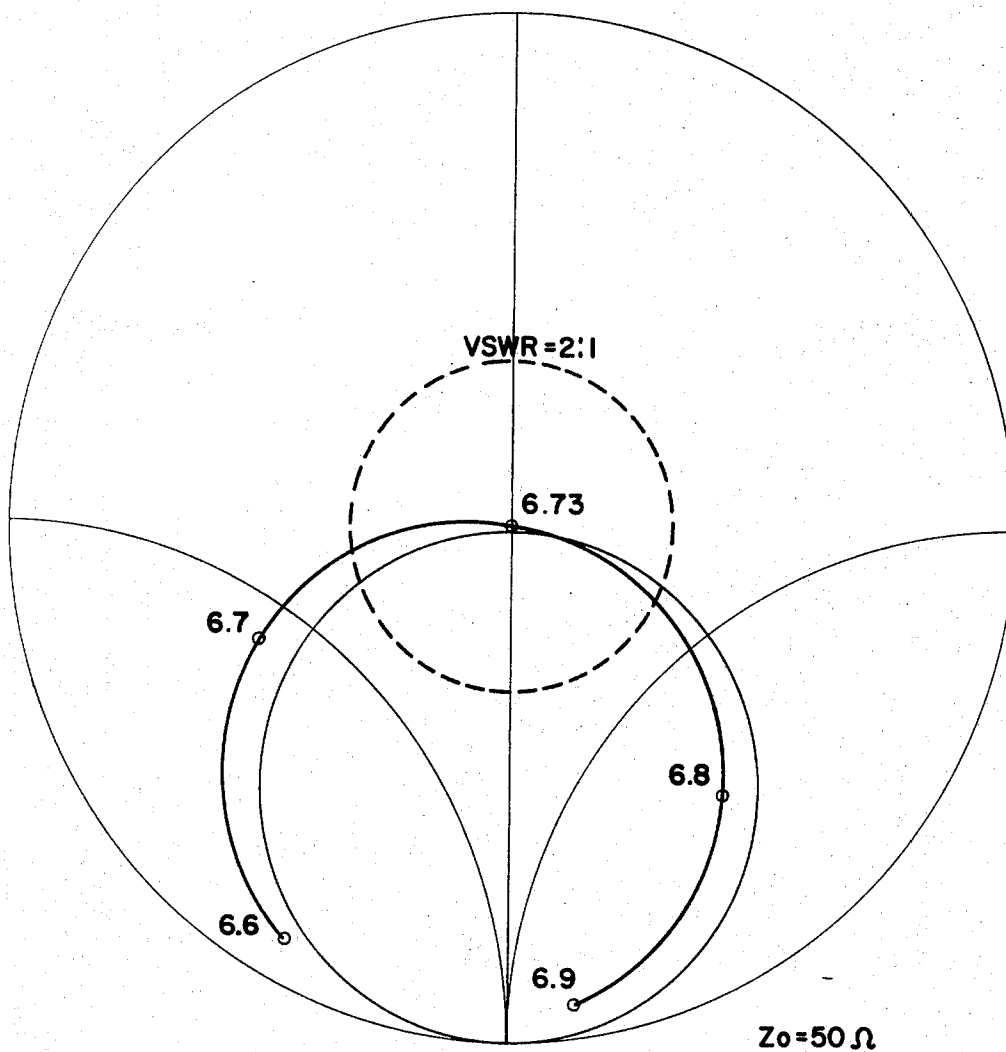


FIG.20

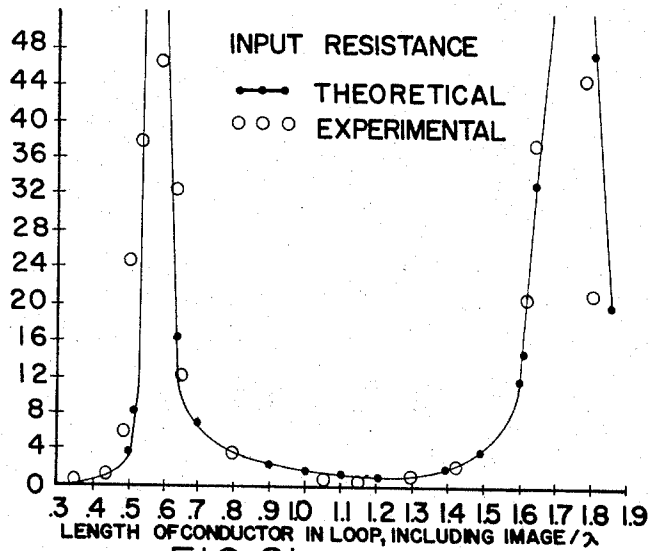


FIG. 21

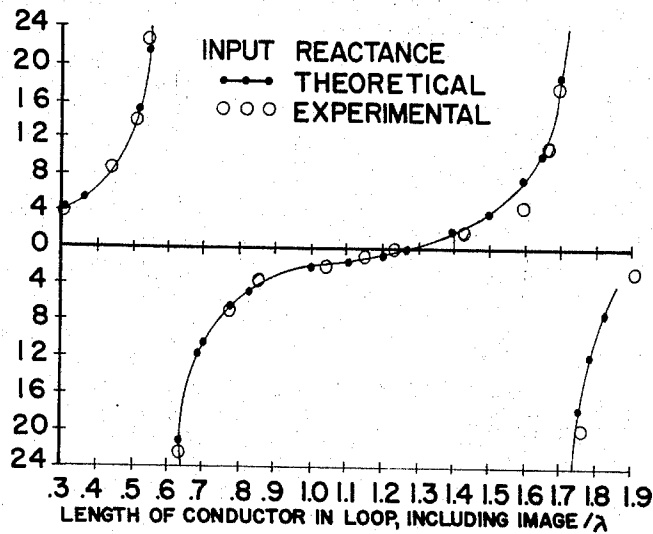


FIG. 22

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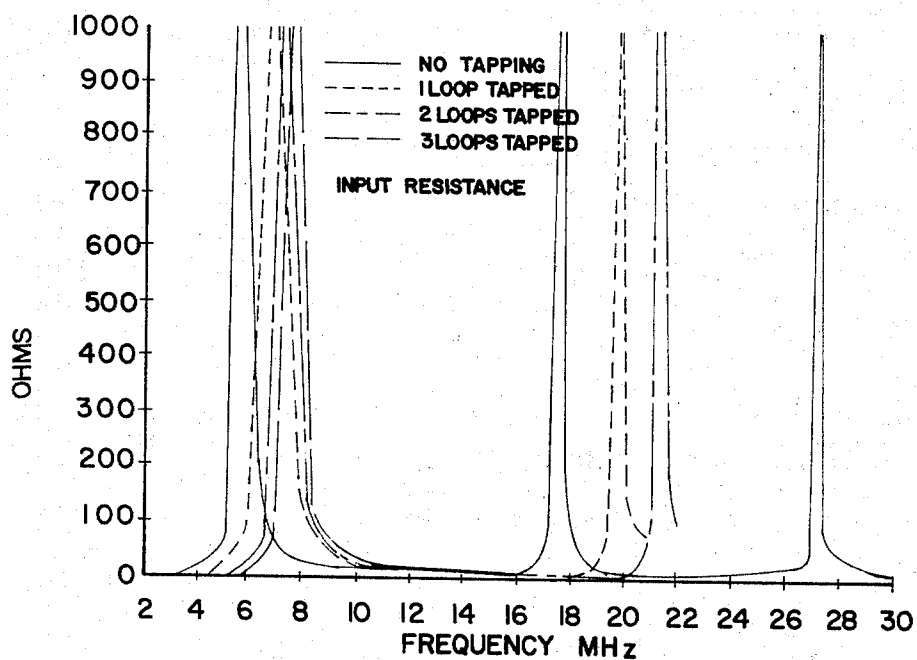


FIG. 24

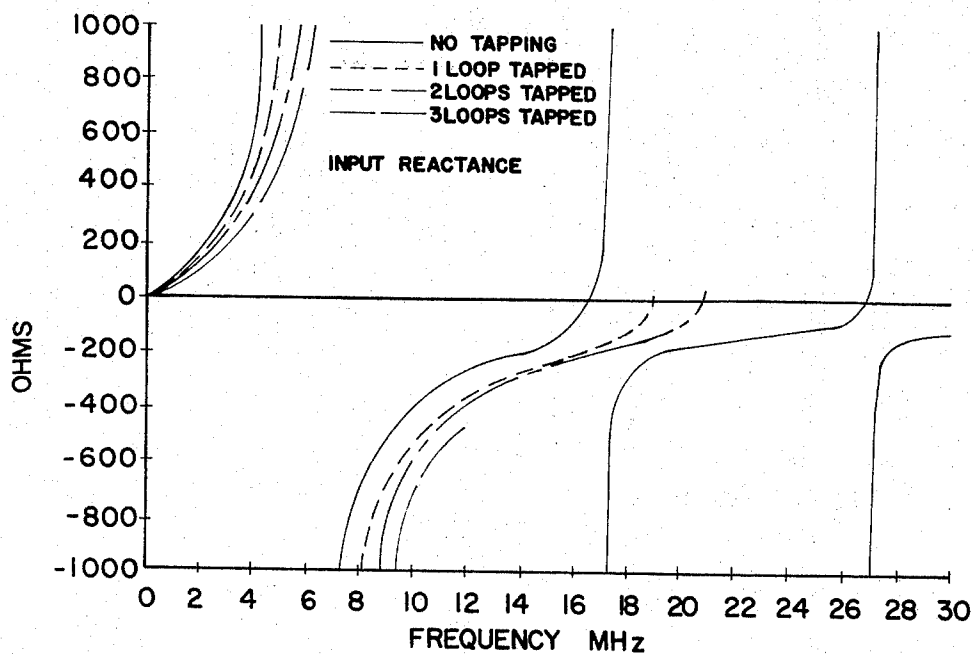


FIG. 25

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MULTITURN LOOP ANTENNA

BACKGROUND

Because of space limitations, many antennas at very high frequency (VHF) and below must be relatively small electrically. This generally implies an undesirable impedance characteristic and low efficiency. The success of such an antenna system may depend upon the efficient induction of tangential H-fields on any nearby support structure; e.g., the excitation of currents on an aircraft structure in order to make use of a larger radiating surface. Currents on a conducting surface are associated with an external tangential magnetic field that is maximum at the conducting surface. Such a magnetic field can generally be established more effectively by a loop element than by a stub or dipole element. However, the conventional small single-turn loop antenna is often too inefficient to be used as the basic radiating element in a feasible antenna system at VHF and below.

SUMMARY OF THE INVENTION

The invention relates to a small diameter loop antenna with an efficiency greatly enhanced over that of a conventional loop by appropriately increasing the number of turns to achieve a natural resonance of the antenna. That is, the antenna is in the form of a multiturn loop that is particularly well suited for use in the very high frequency (VHF) bands and below. This antenna may be used singly or in an array. It is more efficient than a conventional small loop having the same loop area and it is especially well suited for flush mounting. By exciting sizable radiating currents over the supporting structure its efficiency is increased.

An antenna system comprising a pair of multiturn loops at right angles provides omnidirectional radiation with improved efficiency. The array may be positioned in an isolated manner, above a ground plane or some metallic structure, or in a cavity which may be airfilled or loaded with a ferrite or a dielectric that shifts the natural resonances downward in frequency. The antenna is small and since it may be flush mounted is well suited for use in compact receivers and transceivers.

A detailed and mathematical analysis is made for balanced and unbalanced tuning for adaptation at frequencies where the input reactance is either inductive or capacitive.

OBJECTS

Accordingly it is a principal object of the invention to provide an improved loop antenna particularly suited to HF and VHF operation but also well suited to operation at frequencies below HF.

Another object of the invention is to provide an antenna system which permits omnidirectional radiation.

Another object of the invention is to provide an antenna system which is small and flush mounted.

A further object of the invention is to provide an antenna system which permits operation with good efficiency at high power levels or at low signal levels.

Still a further object of the invention is to provide an antenna system which permits reliable communication operation.

For a complete understanding of the invention, together with other objects and advantages thereof, refer-

ence may be made to the accompanying drawings, in which:

DESCRIPTION OF THE DRAWINGS

FIG. 1 is a pictorial representation of the preferred embodiment of the multiturn loop antenna mounted in a cavity in a structure such as an aircraft surface;

FIG. 2 is a diagrammatical representation of the preferred embodiment mounted on a typical aircraft;

FIG. 3 is a graphical illustration of the geometry of the coordinate system used in the theoretical analysis of the basic multiturn loop antenna;

FIG. 4 is a schematic illustration of a multiturn loop antenna over a ground plane of the present invention;

FIG. 5 is a graphical illustration of a wire segment of length "s" and radius "a" of the turns of FIG. 4 with uniform current density on its surface;

FIGS. 6A to 6D illustrate schematically an array of orthogonal multiturn antennas in accordance with the invention;

FIG. 7 is a graphical representation of the input impedance of the 13 turn one-twelfth scale model loop antenna with parallel capacitive tuning;

FIG. 8 is a graphical representation of the efficiency of a five turn loop having a first resonance at about 30 MHz;

FIGS. 9(a through e) are schematic illustrations of a sample of acceptable electromagnetic feed arrangements for the balanced multiturn loop antenna;

FIG. 10 is a pictorial illustration of a practical antenna in a cavity built in accordance with the present invention;

FIG. 11 is a graphical illustration of the input resistance with dielectric loading;

FIG. 12 is a graphical illustration of the input reactance with dielectric loading;

FIG. 13 is a graphical illustration of the efficiency with dielectric loading;

FIG. 14 is a graphical representation of the input reactance with a tuning capacitor between loop and ground;

FIG. 15 is a graphical representation of the input resistance with capacitor termination on ground side of antenna;

FIG. 16 is a graphical illustration of the input reactance with capacitor on ground side of antenna;

FIG. 17 is a schematic side illustration of a pair of multiturn loop antennas;

FIG. 17A is another side view of FIG. 17;

FIG. 18 is an illustration of stagger tuning at the multiturn loop antennas to achieve 90° phase shift for circular polarization;

FIG. 19 is a graphical representation of the input resistance with a tuning capacitor between loop and ground;

FIG. 20 is a graphical illustration of the measured impedance for the antenna of FIG. 1 fed unbalanced and according to the design as depicted in FIGS. 15 and 16;

FIG. 21 is a graphical illustration of measured and computed input resistance of the multiturn loop antenna;

FIG. 22 is a graphical illustration of the measured and computer input reactance of the multiturn loop antenna;

FIG. 23 is a graphical representation of the efficiency of the preferred embodiment illustrated in FIG. 1 with an unbalanced feed as a function of frequency.

FIG. 24 is a graphical illustration of the input resistance of a multiturn antenna showing effect of tapping off various turns;

FIG. 25 is a graphical representation of the input reactance showing effect of tapping off various turns; and

FIG. 26 is a graphical illustration of the multiturn loop antenna in a preferred embodiment with capacitor matching (C_s) and capacitor tuning (C_t).

BRIEF DESCRIPTION OF THE DRAWINGS

The multiturn loop antenna of the present invention consists, in one fundamental configuration, of about a quarter wavelength or odd multiple of quarter wavelength of conductor coiled into two or more turns and mounted in or over a ground plane or metallic structure as illustrated in FIGS. 1 and 4, respectively. The antenna in this arrangement is fed with an unbalanced line and together with its image forms about a half wavelength of conductor at its lowest antiresonant frequency. Another fundamental configuration consists of the antenna with about a half wavelength, or odd multiples thereof, of conductor and fed with a balanced line as illustrated in FIG. 9. A basic characteristic which distinguishes the multiturn loop of the present invention is its operation in the region of a natural antiresonance which substantially increases the efficiency over that of conventional small loops.

By choosing the number of turns and tuning so that the loop element operates below the first resistance peak, the reactive part of the input impedance of the multiturn loop element may be made positive. In this case, the balanced loop element can be matched to a balanced power amplifier by a purely capacitive matching network. This system eliminates matching coils, a major source of loss in conventional antenna systems.

The effect of parallel capacitive balanced tuning is illustrated in FIG. 7. FIG. 7 is a Smith chart plot of the terminal impedance of one-twelfth scale model of the multiturn loop as shown in FIG. 1 with and without a 0.75 pf. capacitor connected across its terminals. As seen in FIG. 7 the loop alone resonates at about 190 MHz while the loop and capacitor combination resonates at about 129 MHz. Thus the resonance frequency of the loop can be greatly altered by a parallel tuning capacitor.

The relative bandwidth of an antenna is the interval of frequencies over which the antenna reactance is less than the antenna resistance, divided by the resonance frequency. From FIG. 7 the bandwidth of the tuned and untuned one-twelfth scale model multiturn loop is about 26.3 percent for the untuned loop and about 11.5 percent when the loop is tuned to resonate at 129 MHz. The bandwidth of the actual model is slightly greater than that measured on the scale model, since the loss resistance is scaled down, relative to the radiation resistance and reactance.

The efficiency curve of the antenna using all its turns is similar in shape to the curve of FIG. 8. The first efficiency "hump" appears at approximately the lowest operating frequency, other "humps" occur at 3, 5, 7, 9, . . . times this frequency, and the low efficiency points are at frequencies in between. However, by

switching out turns the "humps" are moved so that most efficient radiation can occur at chosen frequencies. For example, assuming a 10 to 1 operating band, the first two regions of high efficiency corresponding to $L \approx \lambda/2$ and $3\lambda/2$, would be used, and each would be tuned over a 3 to 1 band by switching out turns on the loop antenna. Therefore, efficiency calculations for various numbers of turns switched out of the loop antenna are made to obtain information on the efficiency throughout the operating frequency band.

In the construction of a working embodiment the efficiency performance for the particular loop geometry is analyzed as noted above. The loop conductor size and configuration, i.e., tube or wire bundle, is then chosen for an optimum combination of loop efficiency, weight, size, and complexity. Several alternative ways of feeding the balanced multiturn loop are within the teachings of the preferred embodiment, certain of these are shown in FIGS. 9a through 9e.

It has been determined that radiating currents are excited on a ground plane by the antenna of the present invention with significant effects. In essence, the excitation of currents on the ground plane doubles the effective radiation, i.e., it is equivalent to an image antenna. The effective antenna is shown schematically in FIG. 4 and analyzed below.

In a practical working embodiment the antenna might not be positioned directly above the ground plane but could be integrally formed into the ground plane as shown in FIG. 26. The height of the loops $a^1 \dots a^n$ is somewhat smaller than the depth of cavity 14; and the diameter of loops $a^1 \dots a^n$ is somewhat smaller than the width of the cavity 14. The feed terminal in this embodiment comprises a coaxial cable 15 more clearly shown in the equivalent schematic illustration of FIG. 4 with the center conductor of the coaxial cable connected to one end of the loop. The other end of the loop together with the outer conductor of the coaxial cable is connected to the ground plane 10. The other end of the center conductor is coupled to a transmitter, receiver, or transceiver through a simple tuning capacitor for proper impedance matching at the operating frequencies. The cavity is a small metal box or metal-coated plastic box which is used as a protective housing and counterpoise and not as a resonant cavity. The structure may be flush mounted into a small counterpoise which can be made light-weight, and may be designed to house other components of the system. By operating the antenna slightly below its natural resonance, the inductive reactance is tuned out by a single capacitor thereby providing for a match to a transmission line into the transceiver. The efficiency (see analysis below) of such a multiturn loop antenna is far greater than that of a monopole of comparable height and may approach that of a quarter wave monopole (whip). The relative efficiency of the multiturn loop compared to a monopole increases as the size (i.e., maximum linear dimension) of the monopole decreases below a quarter wavelength.

The antenna shown in FIG. 10 is mounted in an air-filled cavity 14. The cavity 14 is so small in terms of the wavelength that there are no cavity resonances and its electrical effect on the antenna is relatively small.

To permit reduction of the antenna size, the cavity may be loaded with a dielectric or ferrite. That is, for a given antenna dielectric or ferrite loading shifts the natural resonances downward in frequency as graphi-

cally illustrated in FIGS. 11 and 12. The dielectric used in this embodiment was paraffin ($\epsilon_r \approx 2.23$) since it could be heated and easily poured into the cavity. Certainly other higher dielectric constant materials could be used, but the use of paraffin here serves to demonstrate the effect of dielectric loading. It is interesting to note that the paraffin loading shifts the natural resonant frequencies by nearly the anticipated factor of $1/\sqrt{\epsilon_r}$. Furthermore, preliminary measurements indicate that dielectric loading need not have a particularly adverse effect on the efficiency of the multiturn loop as evidenced by the curves of FIG. 13.

In general whether the loop is mounted in a cavity or on a supporting body such as transmitter, receiver or transceiver, if the permittivity or the permeability (or both) of the medium in which the loop is placed is increased, the effect will be to slow down the traveling wave currents on the loop and thus make the loop appear electrically larger in size. The effect will be to decrease the frequency at which the resistance peaks occur. Thus dielectric loading provides a means of effectively reducing loop size for a given level of performance. Loading the loop with a permeable material such as ferrite would have a similar effect. The loading may take the form of a core for the loop or the loop may be partially or completely embedded in the material.

In a first practical application of the multiturn loop antenna of the present invention, the antenna of FIG. 1 comprising the 13 loops in a cavity was fitted directly into a conducting surface. In this way a flush mounted antenna is provided.

Another practical embodiment of the multiturn loop antenna is that shown in FIGS. 17 and 17A. The multiturn loop is constructed as a module readily attached to the shoulder position of a vest-type garment 21. Owing to the fact that the electromagnetic fields are tightly contained in close proximity to the antenna, relatively little ground plane or counterpoise 10 need be provided around the antenna 19 and 20. Therefore, it is to be expected that the body effects are minimal or that they can be easily compensated for by proper system design.

This configuration of two orthogonal multiturn loops give omnidirectional coverage in the upper hemisphere and thus have no deep nulls to hinder communications no matter what the position of the person wearing the antenna. The omnidirectional characteristic of a pair of right angle antennas is described below. The antenna is connected to a transceiver 23. The antenna-speaker-microphone package could be attached to the vest garment with heavy duty zippers designed to provide electrical contact between the antenna and the small counterpoise 10 around the antenna and also provide a physically secure low profile means of attachment.

With particular reference now to FIGS. 6, 6A, 6B and 6C, there is shown a first multiturn loop and a second multiturn loop antenna positioned at right angles to each other.

Under ideal conditions where it is not disturbed by any nearby objects the multiturn loop has essentially the pattern of a small loop or magnetic dipole. When the loop is mounted over a large ground plane the resulting pattern is essentially that of the loop and its image. For a small ground plane (including cavity mounted loops) the total pattern can be obtained from a superposition of the pattern of the loop and that of the currents that the loop excites on the ground plane

structure. In general the pattern will have a minimum value along the axis of the loop. This minimum ranges from a perfect null for the ideal loop case to something on the order of 10 db or less in the presence of support structure. However, the pattern minimum along the loop axis is effectively eliminated by using a pair of multiturn loops mounted orthogonally (see FIG. 6) and with 90° phasing. Ideally this can be pictured as a two-element array of magnetic dipoles at right angles to one another. With 90° phasing, the resultant pattern is circularly polarized on the axis of the array and vertically polarized with omnidirectional pattern in the plane of the array for an array in the horizontal plane.

The 90° phasing may be achieved by a quarter wavelength of line, by a 90° hybrid coupler, by lumped circuitry, or by proper selection of the operating points of the two multiturn loop antennas. The latter is illustrated in FIG. 18 where one loop is operated somewhat below its resonance such that its input impedance is $R + jX$ and the other just above resonance such that its input impedance is $R - jX$. If, for example, $X = R$ and the loops are connected in parallel, the currents in the two loops will be equal in amplitude but differ in phase by 90° which is the desired condition. A series arrangement might also be used.

Thus, with no additional circuitry, the two loops mounted and operated in such array eliminate the usual minima associated with a single loop or dipole antenna. The physical position of the two loops should not be critical except for the orthogonal relationship. That is, the loops might be mounted in separate cavities as shown in FIG. 17, in the same cavity but physically separated as shown in FIG. 6, positioned one above the other as shown in FIG. 6B, orthogonally interwound as shown in FIG. 6C.

As described above relative to FIG. 9, a balanced tuning arrangement for the multiturn loop antenna is effective and preferred in certain applications. However, it has now been determined that a much improved antenna is that with unbalanced tuning.

With particular reference to FIGS. 1 and 26, an unbalanced feeding arrangement was used for the multiturn loop antenna 19 mounted in a cavity 14. The center conductor of the coaxial feed is connected to one end of the loop antenna while the shield is connected to the cavity (ground plane) 10. The other end of the loop is also connected electrically to the cavity 10 such as through capacitor 27 in FIG. 26. This type of feed eliminates the balun transformer and measurements can be made with standard equipment which has coaxial fittings; mechanical and electrical stability are also improved. Further, since the antenna appears longer by the image effect the resonance points occur at lower frequencies. In this way with the multiturn loop antenna shown in FIG. 1, for example there are three anti-resonance points in the HF frequency band. This allows greater possibilities for matching the antenna to a specified impedance level.

It has been found desirable to operate the multiturn loop antenna at frequencies just below its natural anti-resonances. Although it would appear that the antenna is limited to various frequency ranges in the regions of antiresonances, by tapping (shorting out) turns it is possible to shift the natural antiresonance points. This, in turn, shifts the points where capacitive matching is effective with an upward shift in antiresonances for

one, two, or three turns tapped. Smaller shifts in operating points are obtained by tapping fractional turns.

Although the tapping technique is very effective, it requires an elaborate mechanical or diode-switching arrangement. It was found that the same effect can be produced by using a capacitive termination as 27 in FIG. 26. In this manner, the antenna electrical length is decreased and the antiresonance points shifted accordingly. It should be noted that inductive terminations produce the opposite effect of increasing the electrical length of the antenna. However, since the use of inductors affords a loss in efficiency, only capacitive terminations are considered. Their effect is illustrated in FIGS. 14 and 19.

Therefore, with two variable capacitors, one in series with the feeding system and one acting as a termination as illustrated in FIG. 26, the antenna allows continuous coverage over a wide frequency range. The termination simply shifts the impedance curve to the desired frequency while the series input capacitance nulls out the resulting inductive reactance.

TEST OF CONSTRUCTED EMBODIMENT WITH UNBALANCED FEEDING

To demonstrate the use of this capacitive termination technique, with the antenna operable in the vicinity of 7 MHz, a small capacitive termination (1.8 pf.) is selected to produce a large frequency shift. With the antiresonance point occurring just above 7 MHz, the real part of the input impedance is 50 ohms at 6.66 MHz as indicated in FIG. 15, while the input reactance is 720 ohms as indicated in FIG. 16. Theoretically a series capacitance of 33 pf on the input side tunes out the 720 ohm inductive reactance. Measurements show that the impedance does become pure real at about 6.66 MHz (actually 6.73 MHz) as shown in FIG. 20. The slight error can be contributed to lumped element tolerances and measurement inaccuracies.

FIGS. 21 and 22 show the calculated input impedance as a function of l/λ through two antiresonances. The first antiresonance occurs at the frequency where $l = 0.58\lambda$ while the second one occurs at the frequency where $l = 1.74\lambda$.

In order to verify the predicted characteristics, an experimental model was constructed. The measured input impedance is also plotted as a function of l/λ in FIGS. 21 and 22, and the agreement with the predicted curves is clearly shown.

THEORETICAL AND ANALYTICAL ANALYSIS

In analyzing the multiturn loop antenna of the present invention, pulse basis functions are used. Each turn in the antenna is geometrically approximated by many short, straight segments. Since the segments are not all parallel to each other, the integral operator for this problem is more complicated. To be considered are both the axial and radial components of the electric field from each segment. That is, for the segment of FIG. 5,

$$L_{op}(J(z)) = \hat{z} E_z^s + \hat{\rho} E_\rho^s$$

where L_{op} is defined by Eq. 9,
where

$$E_z^s = \frac{I\sqrt{\mu/\epsilon}}{4\pi jk} \int_{-s/2}^{s/2} [(1+jkr)(2r^2-3a^2) + k^2a^2r^2] \frac{\exp(-jkr)}{r^5} dt$$

Eq. (1a)

and

$$E_\rho^s = \frac{I\rho\sqrt{\mu/\epsilon}}{4\pi jk} \int_{-s/2}^{s/2} (z-t)(3+3jkr-k^2r^2) \frac{\exp(-jkr)}{r^5} dt$$

Eq. (1b)

$$r = \sqrt{\rho^2 + a^2 + (z-t)^2}$$

(1c)

The integration for E_ρ^s may be carried out in closed form giving

$$E_\rho^s = (\rho I \sqrt{\mu/\epsilon} / 4\pi jk) (1+jkr) \exp(-jkr) / r^3 \ln r_1 r_2$$

(2)

where

$$r_1 = \sqrt{\rho^2 + a^2 + (z+s/2)^2}$$

(2a)

$$r_2 = \sqrt{\rho^2 + a^2 + (z-s/2)^2}$$

(2b)

These expressions are accurate if $\rho = 0$ or if r_1^2 and r_2^2 are large in comparison with the quantity $a\rho$. For the segmentation problem of interest here, these conditions will hold true.

To calculate the elements in the impedance matrix, we must in effect calculate the tangential component of the electric field radiated by segment j when the observation point is at the center of segment i . Details of this calculation are found below. The number of equations N is, of course, equal to the total number of segments. Thus, the i^{th} equation may be written as

$$\sum_{j=1}^N Z_{ij} I_j = E_i^i$$

Eq. (3)

However, in this problem there exists a two-fold symmetry between the multiturn loop antenna and its image. Thus the number of unknowns can be reduced by a factor of two and the i^{th} equation may be written as

$$\sum_{j=1}^{N/2} (Z_{ij} + Z_{i, N/2+j}) I_j = E_i^i$$

Eq. (4)

where $I_j = I_{N/2+j}$.

The quantity E^i represents the field due to the source. It is advantageous to use the magnetic frill current to represent the aperture where the coaxial cable joins the ground plane. Thus, an accurate modeling of an actual source aids in computing accurate impedance data that can be confirmed experimentally. For example, FIGS. 21 and 22 show both calculated and experimental values for the input impedance of a two and one-half turn version of the multiturn loop antenna. The agreement between theory and experiment for the reactance as well as the resistance is seen to be excellent. In the calculations, 126 segments were used.

In brief, it has been shown that the effects of adding more turns to the multiturn loop antenna while keeping the physical length of conductor constant and the same turn to turn spacing are threefold. That is to say, the

bandwidth decreases, the first antiresonant frequency is lowered somewhat, and the radiation resistance at nonantiresonant frequencies is reduced.

The implications of the gain and efficiency effects of any nearby structure are also important when considering the loop element. These same loops have efficiencies of 50 percent or possibly more when placed on a metallic surface, such as an airframe, in the correct manner. To obtain these efficiencies, it is necessary that the loops be constructed with thick wire (i.e., on the order of at least 1/100 of the wavelength of the operating frequency).

When the permittivity of the medium in which the multiturn loop is placed is increased, the effect is to decrease the velocity of propagation of the currents on the loop. Hence, dielectric loading provides an effective means of increasing the electrical length of the antenna, and it can be employed in situations where size reduction is important. This also applies to ferrite loading.

If it were possible to surround the entire antenna with a total dielectric material space ϵ , the frequency shift would be given by

$$f_0/f_d = \sqrt{\mu_0\epsilon}/\sqrt{\mu_0\epsilon_0}, \quad (5)$$

where $\epsilon = \epsilon_0\epsilon_r$. Or the shifted frequency, f_d , is given by

$$f_d = f_0/\sqrt{\epsilon_r}, \quad (6)$$

where f_0 is the original frequency and ϵ_r is the relative permittivity constant.

Experimental measurements on the dielectric loaded model of FIG. 1 tend to confirm the theory. With paraffin, $\epsilon_r = 2.23$, FIGS. 11 and 12 illustrate the shifts in resonant frequencies. At the higher frequencies the near fields of the multiturn loop are more tightly bound to the structure, and hence, in terms of electrical dimensions, the antenna is almost entirely surrounded by dielectric space, ϵ . Thus, the first antiresonance point is shifted down a smaller percentage than the second and third antiresonance points.

Efficiency measurements in the vicinity of the first antiresonance show that the 13 turn loop with an air-filled cavity has a peak efficiency of approximately 50 percent. FIG. 13 shows that with dielectric loading the antenna efficiency is reduced slightly and the peak value occurs in the vicinity of the shifted antiresonance frequency.

Although the multiturn loop antenna of the present invention has numerous potential applications, it has defied a rigorous theoretical analysis. A prior analysis considered a balanced feeding arrangement and did not allow for cavity mounting. Experimentally it has been demonstrated that the cavity walls have little effect in influencing impedance characteristics. Unbalanced feeding has the most pronounced in the sense that the antenna appears to be about twice as long as the physical length of conductor due to image effect and antiresonance points occur at lower frequencies.

The experimental curves presented provide significant information for design. For instance, if the fact that the multiturn loop element has anti-resonance points at frequencies where the antenna length (includ-

ing image) corresponds approximately to $\mu/2$, $3\mu/2$, $5\mu/2$, etc., while allowing a loading factor of approximately 0.8 for the cavity effect and interaction between turns, such antennas may be designed for any specified frequency band. Simply using these two pieces of information, antennas have been successfully designed for operation at 40 and 150 MHz. Once the antiresonance point has been placed in the vicinity of the desired frequency, the fine adjustment is obtained by a variable capacitor termination as described above.

The theoretical-numerical method is utilized for determining the current distribution and input impedance associated with the multiturn loop element. The total length of wire (including image) is divided into N (denoted FN in the computer program) segments; the actual turns are approximated by M -sided polygons. If each wire segment is very short in comparison with the wavelength, it is reasonable to assume that the current density is uniformly distributed over the surface of each segment. That is, the true current distribution is approximated by a "staircase" function.

In a rigorous solution, the tangential electric field intensity vanishes everywhere on the surface of each perfectly conducting wire segment. If the wire radius is small, and the segments are short in terms of the wavelength, it is found that accurate results can be obtained by forcing the total tangential electric field to vanish at just one point at the geometric center of each segment.

Since the complex value of the uniform current on each segment is unknown, it is necessary to generate a system of N equations in terms of N uniform currents. Generation of the necessary N equations can be accomplished by requiring that

$$E = E^s + E^i \quad (7)$$

where E is the total electric field at the surface or inside the wire conductor and represents the sum of the scattered field E^s and the incident field E^i .

Assuming that the electric field does not penetrate appreciably to the axis of the segment, E is taken to be zero.

Hence,

$$E^i = -E^s \quad (8)$$

is the condition enforced at the center of each segment. E^i is the incident field from the primary source, while E^s is the scattered electric field caused by the induced current in the conductor. Thus, the problem is to determine the induced current value which satisfies Eq. (8).

The first step in solving the given problem is to accurately determine the value of the incident electric field, E^i , produced by an unbalanced coaxial feed. An unbalanced coaxial feed can be compactly represented by a magnetic frill current source. The near and far-zone electric and magnetic fields from an annular ring of circumferentially directed magnetic current are calculated. For the near fields, numerical integration is used to obtain the electric vector potential, and then numerical differentiation methods are used to determine the near-zone field values. With suitable approximations,

closed-form expressions for the fields are then found both in the far near-zone and on the axis of the ring.

Using the fields of the magnetic frill in Eq. (8), the required system of N simultaneous linear equations in N unknowns I_n is developed in the following manner:

Defining the linear (integral) operator

$$L_{op}(J) = (-E^s)_{tan} \quad (9)$$

where E^s is the scattered electric field and J is the current density on a particular segment, the equation for J can be written as

$$L_{op}(J) = (E^i)_{tan} \quad (10)$$

where $(E^i)_{tan}$ denotes the tangential component of the bracketed quantity on the segment.

If a unique solution to Eq. (10) exists for all (E^i) , then L_{op}^{-1} exists such that

$$J = L_{op}^{-1} (E^i)_{tan} \quad (11)$$

A suitable inner product for the problem is

$$\langle J, E \rangle = \iint_S J \cdot E ds, \quad \text{Eq. (12)}$$

$$\begin{bmatrix} \langle \delta(s-s_1), LJ_1 \rangle & \langle \delta(s-s_1), LJ_2 \rangle \\ \langle \delta(s-s_2), LJ_1 \rangle & \langle \delta(s-s_2), LJ_2 \rangle \\ \vdots & \vdots \\ \langle \delta(s-s_N), LJ_1 \rangle & \langle \delta(s-s_N), LJ_2 \rangle \end{bmatrix}$$

which is quantity called reaction and represents a closed surface integral over the scatterer (loop) surface S . It satisfies all the following requirements:

$$\langle J, E^i \rangle = \langle E^i, J \rangle \quad (13)$$

$$\langle \alpha J_1 + \beta J_2, E^i \rangle = \alpha \langle J_1, E^i \rangle + \beta \langle J_2, E^i \rangle \quad (14)$$

$$\langle J^*, J \rangle > 0 \text{ if } J \neq 0 \quad (15a)$$

$$\langle J^*, J \rangle = 0 \text{ if } J = 0 \quad (15b)$$

where α and β are scalars and $*$ denotes a complex conjugate. J_1 and J_2 denote two different current densities.

Now applying the method of moments to Eq. (10), J is expanded in a series of functions in the domain of L_{op} such that

$$J = \sum_{n=1}^N I_n J_n, \quad \text{Eq. (16)}$$

where I_n are complex constants. The J_n are denoted sub-domain (segmentation) basis functions, where

$$J_i = 1 \text{ for segment } i \quad (17)$$

= 0 otherwise.

Substituting Eq. (16) into Eq. (10) and using the linearity properties of L ,

$$L_{op} \left(\sum_{n=1}^N I_n J_n \right) = (E^i)_{tan} \quad \text{Eq. (18)}$$

or

$$\sum_{n=1}^N I_n L_{op}(J_n) = (E^i)_{tan} \quad \text{Eq. (19)}$$

Now defining a set of weighting functions, $W_1, W_2, \dots, W_m, \dots, W_N$, which are tangential vectors on the respective segments, the inner product

$$\sum_{n=1}^N I_n \langle W_m, L_{op} J_n \rangle = \langle W_m, E^i \rangle \quad \text{Eq. (20)}$$

is valid for $m = 1, 2, \dots, N$. The subscript "tan" has been dropped since the inner products involve tangential components only. There are several choices for the weighting functions, W_m . One common one is the Dirac delta function; this procedure is called point-matching. Equation (19) is satisfied only at discrete points — namely the center of each segment. Letting $W_m = \delta(s - s_m)$, where s is the measure of the distance from some reference point on the antenna and s_m represents the distance from the same reference point to the center of segment m , the following matrix equation results:

$$\begin{bmatrix} \langle \delta(s-s_1), LJ_1 \rangle & \langle \delta(s-s_1), LJ_2 \rangle \\ \langle \delta(s-s_2), LJ_1 \rangle & \langle \delta(s-s_2), LJ_2 \rangle \\ \vdots & \vdots \\ \langle \delta(s-s_N), LJ_1 \rangle & \langle \delta(s-s_N), LJ_2 \rangle \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ \vdots \\ I_N \end{bmatrix} = \begin{bmatrix} \langle \delta(s-s_1), E_1^i \rangle \\ \langle \delta(s-s_2), E_2^i \rangle \\ \vdots \\ \langle \delta(s-s_N), E_N^i \rangle \end{bmatrix} \quad \text{Eq. (21)}$$

or

$$[Z] (I) = (V) \quad (22)$$

where the elements of $[Z]$ are called generalized impedances, those of (I) generalized currents, and those of (V) generalized voltages. Solving for (I) , one obtains

$$(I) = [Z]^{-1} (V) \quad (23)$$

and J is given by Eq. (16). I_n represents the current density induced on segment n , and E_n^i is the tangential component of the incident electric field at the center of segment n .

What is claimed is:

1. An antenna comprising an electrically conducting surface, and a multiturn element having an overall length substantially at $\frac{1}{4}$ wavelength or odd multiples thereof; a capacitance means terminating one end of said multiturn element to said surface, and an unbalanced feed further comprising a matching capacitance.

2. An antenna system as set forth in claim 1 wherein said capacitor further comprises a capacitor in series with said multiturn element and said surface to finely adjust the apparent electrical length of said loops.

3. An antenna system as set forth in claim 1 wherein said capacitor further comprises a capacitor in parallel

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with said multiturn element and said surface to finely adjust the apparent electrical length of said loops.

4. An antenna system as set forth in claim 1 further comprises dielectric loading means of said multiturn element to lower the frequencies at which the resistance peaks of said elements occur.

5. The antenna of claim 1 wherein said multiturn loop is embedded in a ferrite material.

6. An antenna as set forth in claim 1 wherein said loop further comprises a discontinuity which varies the length of said loop as a function of said operating frequency.

7. An antenna of claim 1 wherein said multiturn loop is embedded in a dielectric material.

8. The antenna of claim 1 wherein said multiturn element comprises a first multiturn loop and a second multiturn loop to provide an omnidirectional radiation pattern.

9. The antenna of claim 1 wherein said first and second loops are positioned at right angles relative to one another.

10. An omnidirectional antenna of claim 8 wherein each of said multiturn loops is tuned to the same frequency.

11. An omnidirectional antenna of claim 8 wherein said first multiturn loop is tuned slightly lower and second multiturn loop tuned slightly higher than the operating frequency of said antenna.

12. An omnidirectional antenna of claim 9 wherein said right angle position comprises positioning said multiturn loops in a common plane.

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13. An omnidirectional antenna of claim 9 wherein said right angle position comprises positioning said multiturn loops one above the other.

14. An omnidirectional antenna of claim 9 wherein said multiturn loops are intertwined one within the other.

15. An omnidirectional antenna of claim 9 wherein said conductive surface is a cavity and wherein said first and second right angle multiturn loops are positioned in said cavity.

16. An omnidirectional antenna of claim 15 wherein said cavity is a single cavity and said loops are positioned in said single cavity.

17. An omnidirectional antenna of claim 15 wherein said cavity is a pair of cavities and said first multiturn loop is positioned in said first cavity and said second multiturn loop is positioned in said second cavity.

18. An omnidirectional antenna of claim 13 wherein said cavity further comprises material loading means for said loops to vary the resonant frequency of said loops.

19. An omnidirectional antenna of claim 16 wherein said material loading means is a dielectric.

20. An omnidirectional antenna of claim 16 wherein said material loading means is a ferrite.

21. An antenna system as set forth in claim 16 further comprises ferrite loading means of said loops to lower the frequencies at which the resistance peaks of said elements occur.

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UNITED STATES PATENT OFFICE
CERTIFICATE OF CORRECTION

Patent No. 3,823,403 Dated July 9, 1974
Inventor(s) Carlton H. Walter et al.

It is certified that error appears in the above-identified patent
and that said Letters Patent are hereby corrected as shown below:

Item [73] "Columbus, Ga." should read

-- Columbus, Ohio --.

Signed and sealed this 6th day of May 1975.

(SEAL)
Attest:

RUTH C. MASON
Attesting Officer

C. MARSHALL DANN
Commissioner of Patents
and Trademarks