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TRAVELING WAVE ANTENNA HAVING RADIATOR ELEMENTS WITH DOUBLY PERIODIC SPACING

Filed Sept. 17, 1964

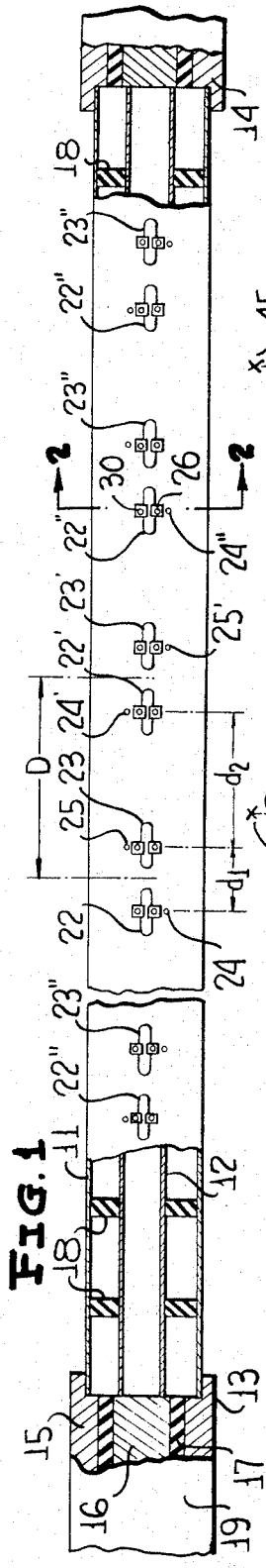


FIG. 1

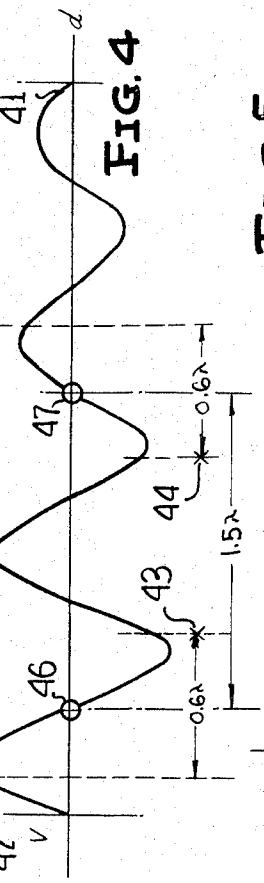


FIG. 4

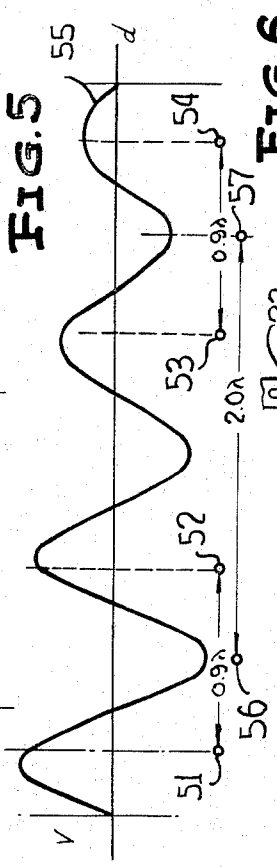


FIG. 5

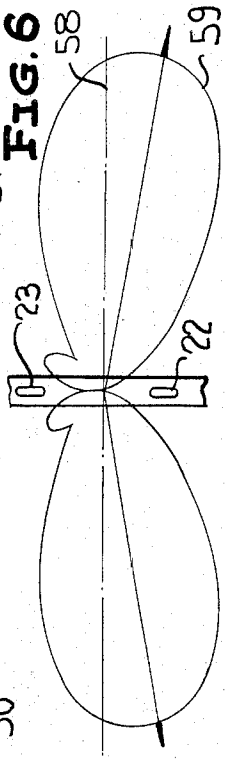


FIG. 6

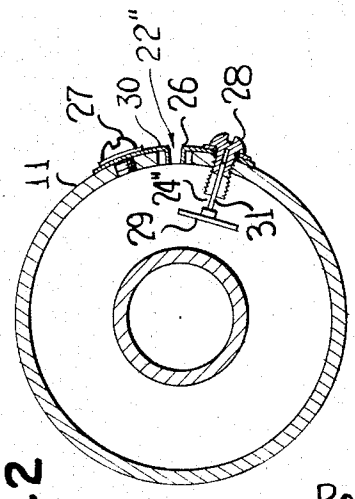


FIG. 2

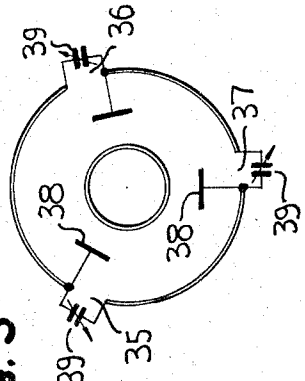


FIG. 3

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ABSTRACT OF THE DISCLOSURE

A traveling wave antenna in which electromagnetic waves are propagated along a transmission channel for absorption of energy therefrom and consequent radiation of the electromagnetic waves by a plurality of pairs of recurrent radiating elements coupled to the channel for cophasal excitation by each electromagnetic wave, the radiating elements of each pair separated longitudinally along the channel by a distance departing substantially from an integral number of half wavelengths of the wave, an element in any pair being separated from the adjacent element in the next pair by a distance differing substantially from the aforementioned distance, and the centers of recurrent pairs separated by an integral number of half wavelengths.

The present invention relates generally to antennas and more particularly to a traveling wave antenna having radiator elements with doubly periodic spacing.

The need presently exists, particularly in the V.H.F. and U.H.F. transmitting frequencies, for an improved antenna having good reliability, high power-handling capability and high gain. Such an antenna should be capable of highly directional or omnidirectional, nominally broadside radiation beaming, the beams being of a prescribed vertical contour having no nulls in the primary service zone, and should further have electrical characteristics wherein low standing wave ratios are obtained over the operating frequencies of interest. For television transmission, it is desirable for such an antenna to be fed simultaneously with audio and video signals without the need of a diplexer.

It is accordingly, a principal object of the present invention to provide an antenna having the above-mentioned capabilities and characteristics.

Another object of the present invention is to provide a traveling wave antenna having doubly periodic spacing of radiators with respect to position along the antenna axis.

An additional object of the present invention is to provide a traveling wave antenna capable of producing a highly directional null-filled beam, which is substantially fixed in an arbitrarily predetermined direction with respect to the longitudinal axis of the antenna.

A further object of the present invention is to provide a new and improved traveling wave antenna comprising a transmission line having a low standing wave ratio.

Still another object of the present invention is the provision of an improved antenna having characteristics particularly appropriate for the transmission of television signals, and dispensing with the usual requirement of a diplexer for coupling audio and video signals thereto.

A still further object of this invention is to provide a traveling wave antenna having a omni-directional azimuth pattern.

The above and still further objects, features and advantages of the present invention will become apparent from a consideration of the following discussion, and of the

detailed description of a preferred embodiment which is provided for illustrative purposes only.

It is generally understood that a traveling wave line source antenna is a radiating structure along which a wave proceeds in a manner characterized by a complex propagation constant identifiable as $\gamma=\alpha+j\beta$, wherein the real part, α, represents the rate of decay of wave amplitude with respect to distance along the axis as a result of the gradual dissipation of energy by radiation, and the imaginary part, β, describes the intrinsic rate of phase retardation of the wave. Both components are functions of frequency, and may or may not involve position along the antenna axis. For an antenna to qualify as a traveling wave antenna, it is necessary for α to be greater than zero. For a strictly continuous straight uniform traveling wave structure, the direction angle θ of the phase front with respect to the forward traveling wave axis is given by

$$\cos\theta=\beta/\beta_0 \quad (1)$$

where β_0 is the phase propagation constant in the surrounding medium. Except for the possible distorting effect which the obliquity factor associated with a typical element of the array might exert upon the array pattern, this θ is also the direction of the radiation beam. The expression loses its immediate significance when β varies with position along the antenna axis, hence the discussion which follows will apply primarily to quasi-uniform structures in which the structural plan is at least repetitive.

It will be seen from Equation 1 that near-broadside radiation beaming can be achieved with continuous straight uniform structures only when $\beta<<\beta_0$, and precisely-broadside beaming only when $\beta=0$, that is, when $\cos\theta=0$. An example of such an antenna is the trivial degenerate case of a uniform straight tubular metallic waveguide with a continuous axial slot which is excited by a wave introduced at one end of the structure at the waveguide cut-off frequency.

It is possible, however, to control the nominal beam direction in certain cases by resorting to a twist configuration or to a helical wind, which, with a proper type of basic filament, modifies Equation 1 to the more general form,

$$\cos\theta=(\beta-k)/\beta_0 \quad (2)$$

where k is the angular rate of twist or turn per unit distance along the antenna axis. By prescribing the twist rate or wind rate, as the case may be, one actually prescribes an effective phase propagation constant for the aperture excitation which may be negative as well as positive, thereby introducing the possibility of a backyard beam traveling wave antenna. A familiar example of a continuous structure of this kind is the helical wire antenna. Such structures have the limitation that their radiation beams are omnidirectional in the sense that the radiation amplitude is basically independent of angular direction about the antenna axis at fixed co-latitudinal angles with respect to the axis.

Another way to gain control of the nominal beam direction in traveling wave antennas is through the use of periodicity of structure which implicitly permits a 180-degree arbitrariness of the excitation at any given radiator location with respect to the phase of the internal traveling wave which supports the excitation. Inasmuch as Equation 2 is fundamental for all quasi-uniform traveling wave structures, its application to straight, i.e., not physically identifiable as a twist structure, uniformly periodic ones is simply a matter of expressing correctly through k the conditions of coupling sense which determine whether or not the 180-degree arbitrariness is

utilized in the particular situation. The correction is at the rate of π radians per element spacing for a uniformly spaced set of elements in which alternate elements are coupled in the reverse sense. That is, $k=\pi/d$, where d is the spacing between successive elements. It is immediately obvious in this situation that the only control on beam direction for a given β is through the spacing of the elements. The condition for broadside radiation in any uniform periodic structure is that $k=\beta$. If the elements are alternately reversed in coupling relation with respect to the internal traveling wave, then they must be spaced exactly one-half wavelength apart as determined by β. If the reversed coupling degree of freedom is not used, the radiators must be spaced a full wavelength apart, because in that case $k=2\pi/d$. An anomaly apparently arises now, since a phase correction of 2π is electrically identical with a zero correction, and the condition for a lobe in some other direction thus exists simultaneously. The situation is resolved in practice by using elements which favor the desired lobe and then applying the corresponding k for that lobe in Equation 2. Thus, $k=2\pi/d$ is the proper correction for a broadside array without coupling reversals, while $k=0$ would be the correction term for the zero order, quasi-endfire lobe. Higher order lobes continue to become realizable as spacings increase and coupling reversals are utilized, but they are comparatively trivial. The correction terms are $k_n=n\pi/d$ for consideration of the nth order lobe, alternated couplings being implied for the odd-numbered orders.

Unfortunately, the properties of such singly periodic systems are such that the attenuation constant α rapidly approaches zero as the spacing between the points at which the radiators are coupled to the transmission system approaches a whole multiple of electrical half-wavelengths as developed in the transmission system. The attenuation constant in such systems is thus exceedingly sensitive to variations in frequency for operation in the neighborhood of or at the broadside beaming condition. All elements receive substantially the same excitation, and the antenna characteristics resemble more nearly those of a resonant or standing wave type of structure than those of a true traveling wave structure. The input impedance, which would otherwise be rather insensitive to frequency variations, becomes very critical in broadside operation.

I have shown, for example in my U.S. Patent 2,947,988, that the strict dependence of beam direction with respect to the axis upon element spacing that seems to hold for essentially-aligned singly-periodic traveling wave antenna structures does not necessarily govern the more general category of twist configurations. The minimum condition that must be satisfied for this to be possible is that the distant electric field due to a short representative radiating segment (or basic source unit) of the structure have the property that

$$E_\theta(\theta,\phi)=-E_\theta(\theta,\phi+\pi) \qquad (2a)$$
$$E_\phi(\theta,\phi)=-E_\phi(\theta,\phi+\pi) \qquad (2b)$$

where the subscript denotes the coordinate contour along which the component in question is directed in a spherical coordinate system whose positive Z-axis coincides with the array axis, and the traveling wave starts at the origin. If both components exist, both must satisfy the relations, but it is neither essential nor particularly desirable that both components exist. An ordinary dipole centered on the antenna axis and oriented orthogonal to it is, perhaps, the most familiar such element which has both the θ- and ϕ-components, and which satisfies the requirements. The number of admissible source types is not great, but there are several.

One preferred practical arrangement is described in detail in my aforementioned U.S. patent. It comprises a pair of axially oriented slots in the wall of a circular conducting cylinder comprising the outer conductor of a large coaxial transmission line. The slots of a given pair are diametrically opposite each other and centered at the same cross section of the cylinder. They were excited in phase opposition by means of probe couplers to the internal traveling wave in such a way as to produce a figure-of-eight shaped amplitude radiation pattern in ϕ-traverses at any fixed angle θ. This unit source has only the E_ϕ radiation field component. A uniform twist traveling wave antenna was constructed using these basic units progressively advanced in angular orientation about the cylinder axis at regular 90-degree increments. In order to accomplish broadside beaming, it was necessary in consequence of Equation 2 to set the uniform longitudinal spacing between successive pair centers at 90-degree phase intervals along the structure. This arrangement satisfied the requirement of one turn of twist per electrical wavelength developed in the loaded transmission line. The result was a highly directional broadside beam having excellent circularity of coverage in azimuth and good stability of direction with respect to the axis over an appreciable band of frequencies.

The 90-degree twist rate of advance between successive elements was selected for several reasons, among them being the minimization of exterior radiation coupling between successive pairs and the especially favorable behavior of the internal system electrical characteristics. In principle, the beam direction is determined only by the twist rate. The axial spacing between elements may be any desired value compatible with the requirements on the attenuation rate, internal electrical characteristics, and pattern shape as related to the density of elements in the aperture. One must still exclude spacings approximating an integral number of half-wave lengths, because of the undesirable behavior of the attenuation function for that situation. The feature that is generally unavailable in twist structures, therefore, is arbitrariness of azimuth pattern, since all such structures are intrinsically omnidirectional. The only exception that might exist is in the case where one chooses to interpret the mathematical categorization so as to include those antennas having one element per half-turn, or per full turn, which twist rate throws the individual elements into axial alignment. It is purely a matter of viewpoint, since the twist in such antennas is not physically apparent. This is a borderline case inasmuch as such a structure cannot behave as a traveling wave antenna for broadside radiation because the attenuation rate is zero for the requisite integral half-wavelength element spacing. It qualifies as a traveling wave antenna only for beam directions substantially different from broadside, and may be considered a specially restricted twist traveling wave antenna for purposes of analysis if so desired.

I have dwelt briefly upon the characteristics of the more common types of traveling wave antenna in order to point out more clearly how my present invention is a significant improvement. I have found that it is possible to construct a broadside beaming traveling wave antenna having all the desirable characteristics of traveling wave antennas, but which is not limited to omnidirectional radiation pattern applications, by utilizing a doubly periodic distribution of radiating elements. The distribution is obtained by spacing the radiating elements alternately along the structure in a regular manner such that a wide spacing follows a narrow one, and the sum of any two consecutive spacings is a whole number of electrical half-wavelengths as developed internally in the propagating system. For directions other than broadside, the electrical length corresponding to one cyclic variation of the spacing arrangement will, in general, differ from a whole multiple of electrical half-wavelengths in order to be in agreement with Equation 2. To accomplish broadside beaming, one feasible distribution of the elements in terms of wavelengths developed in the system is $d_1=0.4\lambda$ and $d_2=0.6\lambda$ alternately applied in cadence with regular alternation of the coupling senses of the elements with respect to the phase of the traveling wave which excites them. It will

be seen that with $k=\pi/d$ this practice renders $k=\beta$ on the average, and hence produces broadside beaming of the radiation. The phase distribution of the resulting antenna excitation then alternates about the desired constant (or linearly progressive norm) by an amount determined by the difference between the actual spacing of the elements and what would be the corresponding uniform spacing for the same beam bearing. Corresponding deviations result in the amplitude distribution of the excitation, which varies about an exponential average curve.

It may occur to the reader that the available correction term for beaming in any prescribed direction amounts to the specification of d for accomplishing that performance at a particular frequency. In order to specify the beam direction over a band of frequencies, it is necessary that $\beta=\beta(\omega)$ continue to satisfy Equation 2 over that band. If the beam direction is to remain fixed, for example, Equation 2 requires that

$$(\beta-k)\doteq A\beta_0$$

or

$$\beta\doteq A\beta_0+k \tag{3}$$

where A is the required constant value of cos θ. Since β_0 is linear in frequency, it is evident that β must also be a substantially linear function of frequency, but with a nominal slope now determined by the beam direction. If one is too precise in prescribing β over the band, it is found that α is substantially determined automatically to within an additive constant over the same band by virtue of certain relationships between attenuation and phase which are inviolate in realizable networks. An example of this relationship for a singly periodic traveling wave antenna has been derived. If Y_0 is the characteristic admittance of the unperturbed transmission system across which radiating elements having an admittance given by

$$Y=G+jB$$

are periodically connected at spacings d, then the relation between α and β per iterated mid-shunt section, in terms or normalized admittance, $y=Y/Y_0=g+jb$, is

$$\cos\beta=\frac{1}{\cos h\alpha}\left[\cos\bar{\beta}d-\frac{b}{2}\sin\bar{\beta}d\right] \tag{4}$$

where $\bar{\beta}$ is the phase propagation constant in the unperturbed section of transmission line. Close examination of the behavior of this function shows that α must approach zero as $\bar{\beta}d$ approaches $n\pi$. Explicit expressions for sin $h\alpha$ and cos β in terms of y and βd for such structures show that α becomes zero in the limit for all finite values of y as $\bar{\beta}d$ approaches integral values of π; and that β per iterated section simultaneously approaches $\bar{\beta}d$, taking on integral values of π in the limit. Traveling wave antennas are used mostly in high gain applications which require electrically long apertures and, hence, values of α so low that cos $h\alpha$ is almost equal to unity. Equation 4 shows that under these circumstances β depends almost entirely upon the load susceptance b for a given design value of $\bar{\beta}d$. Inasmuch as the beam direction and β are directly related by Equation 3, it follows that a specification on the frequency dependence of the beam direction corresponds to a constraint on the functionality of b. To render the first frequency derivative of beam direction zero in a given direction, for example, requires b to have a certain negative slope with respect to frequency, for a given load spacing. These comments are important in the present invention, since it is also true for doubly periodic structures that the desired performance can only be achieved by connecting suitable radiation elements in the proper manner.

The case of a doubly periodic structure may also be analyzed on the same basis. If d_1 and d_2 are the two alternated spacings between successive identical elements,

then the relationship between α and β per mid-series repeated two-section is

$$\cos\beta=\frac{1}{\cos h\alpha}\left[\cos\bar{\beta}(d_1+d_2)-b\sin\bar{\beta}(d_1+d_2)+\frac{(g^2-b^2)}{2}\sin\bar{\beta}d_1\sin\bar{\beta}d_2\right] \tag{5}$$

This expression is not readily assessed by inspection, but it is evident that when $\bar{\beta}(d_1+d_2)$, with $d_1\neq d_2$, is exactly a whole multiple of π, then a residual third term in the bracket of Equation 5 exists through which control can be accomplished to a certain extent by properly choosing the characteristics of the coupled antenna element. It is this doubly-periodic feature of my invention which basically constitutes the improvement over existing traveling wave antennas.

Since no requirement has been made for twist or wind in the doubly periodic structure, the azimuthal pattern is entirely independent of the conditions affecting the beam direction θ. It can therefore be made either directional or omnidirectional in the azimuthal plane at will without affecting the beam pattern in the axial plane.

It will be understood that antennas constructed in accordance with the present invention are applicable to any situation in which it is desired to beam the radiation in a particular direction with respect to the longitudinal axis of the array. However, such an antenna is particularly advantageous in applications requiring nominally broadside beaming, and in the following detailed description an exemplary embodiment for such use will be set forth. In order to accomplish both broadside beaming independent of azimuth pattern and internal wave attenuation resembling that occurring in a long dissipative transmission line, it is essential that pairs of radiating elements be spaced by a factor differing from integral half-wavelengths of the electromagnetic wave in the transmission channel structure to which they are coupled. Thus, in accordance with the present invention an antenna, or antenna array, comprises a doubly periodic distribution of radiating elements along the longitudinal axis of the structure. This distribution is such that a plurality of pairs of recurrent radiators are coupled to the wave transmission channel as nearly in phase as possible, compatible with the requirement that the radiators be spaced alternately along the structure in a recurring or regular pattern of successively narrow and wide intervals, d_1 and d_2, each interval differing from an integral number of half wavelengths. If the attenuation factor α is to be relatively insensitive to tolerance on spacing, these spacing intervals should depart from integral half-wavelengths by a factor of at least one-tenth wavelength. The principal requirement with respect to providing broadside beaming is that the sum, D, of two consecutive spacing intervals be an integral number of half-wavelengths. In other words, $D=d_1+d_2=N\lambda/2$, but d_1, $d_2\neq n\lambda/2$, where N and n are integers, $N>n$, for broadside beaming. Put another way, the centers of recurrent pairs are separated by a distance of approximately an integral number of half-wavelengths, the radiation beam of the array approaching a broadside beam as the aforesaid distance approaches an integral number of half-wavelengths.

To provide an azimuth pattern that is substantially unidirectional, a column of radiating elements having the above-described doubly periodic distribution is aligned longitudinally along the transmission channel of the antenna structure. The most elementary practical embodiment would be a coaxial transmission line or hollow waveguide in which the radiating elements are slots in the outer conductor or guide, respectively, appropriately coupled to the wave transmission channel for the required excitation. If an omni-directional azimuthal pattern is desired, three slots may be provided separated at angles

of 120° about the structure axis at regular cross-sections, i.e, cross-sections perpendicular to the longitudinal axis, in accordance with the doubly periodic pattern. To obtain other azimuthal patterns, the radiators are suitably placed at other angular positions about the axis of the structure.

It will, of course, be understood that while reference is made to the use of slots as suitable radiating elements, it is not to be inferred that the radiators are limited to slots. Any radiating element which is suitable to the particular practical application involved will give the required performance if properly connected so as to satisfy constraints placed on the coupled impedance or admittance function as discussed above with reference to Equation 4. By way of example, the duality of theory will indicate that if a negative slope is required for the susceptance function of a shunt load, as in the case of slots suitably coupled along the transmission channel of a coaxial line, then a negative slope would be required of the reactance function of a series load, and so forth. Thus, a loop coupled series resonant radial monopole element, for example, will accomplish in a series load what a probe coupled anti-resonant slot element will accomplish in a shunt load for a given specification of rate of change of beam direction with respect to frequency at the design frequency.

If an antenna in accordance with the present invention is to be employed for television transmission, the customary diplexer for audio and video signals may be dispensed with by coupling each signal at an opposite end of the antenna. The antenna in such a case should be several wavelengths long, for example nine or ten wavelengths of the applied signal, so that negligible interaction occurs at the ends. That is, with the element spacing of the doubly periodic pattern there is sufficient attenuation of signal along the traveling wave antenna such that there is, from a practical consideration, no signal mixing at the ends. In addition, the long antenna obviates the possibility of reflections so the driving point impedance variation may be minimized.

With the above considerations in view, both the expressed and unexpressed objects, features and advantages attendant to such considerations will become further apparent as reference is now made to the accompanying drawings in which:

FIGURE 1 is a fragmentary side view, partially in section, of one embodiment of the present invention;

FIGURE 2 is a view taken through the lines **2—2** of FIGURE 1;

FIGURE 3 is a cross-sectional view of a modification of the embodiment of FIGURES 1 and 2;

FIGURE 4 illustrates the voltage distribution function of an electromagnetic wave traveling along one embodiment of antenna in accordance with the present invention;

FIGURE 5 illustrates a voltage distribution function for another embodiment; and

FIGURE 6 is an illustration of the antenna pattern of a single radiator pair in an axial plane.

FIGURE 1 illustrates a traveling wave antenna having a pair of coaxial, circular conductive tubes **11** and **12**. Disposed at opposite ends of tubes **11** and **12** are coaxial connectors **13** and **14**, each of which has an exterior metal sleeve **15** and an interior metal sleeve **16**. Sleeves **15** of each of connectors **13** and **14** make electrical contact with exterior tube **11** while interior sleeve **16** contacts tube **12** in accordance with standard coaxial construction. A cylindrical insulator **17** is interposed between sleeves **15** and **16** in each connector. Tubes **11** and **12** are separated by annular dielectric rings **18**, fabricated of a suitable electrically insulating material having a dielectric constant as close as possible to that of air (≈ 1), which exists in the space between the tubes.

The configuration thus far described is, of course, a coaxial transmission line or channel, to which suitable signals may be coupled at either end via the connectors. The line may be modified for performance as a traveling wave antenna structure by providing a plurality of longitudinally aligned slots in recurring pairs, as **22**, **23**. In accordance with the requirements for the traveling wave type of behavior, adjacent slots are spaced at intervals of other than a half-wavelength as measured in terms of phase velocity in the propagating system. Slots **22** and **23** are, for example, spaced at a distance d_1, while slots **23** and **22'** are spaced by an interval d_2, the sum of d_1 and d_2 being designated in FIGURE 1 as D. D is obviously also the distance between centers of recurring pairs as the slots are maintained at regular short and long intervals, d_1 and d_2 respectively, in longitudinal alignment down the line. As previously described, both d_1 and $d_2 \neq n\lambda/2$ where n is an integer and λ is the wavelength of the electromagnetic wave traveling down the line. Preferably both d_1 and d_2 depart from an integral number of half-wavelengths by at least one-tenth wavelength. The separation of pair centers, $D=d_1+d_2$, is maintained at approximately $N\lambda/2$, N being an integer greater than n, and as D aproaches an integral number of half-wavelengths, i.e. $D \rightarrow N\lambda/2$, the beam radiated by the overall structure approaches a true broadside beam.

By way of example, the electrical distance between slots may be set in the range $0.35\lambda \leq d_1 \leq 0.45\lambda$, with d_2 being the complementary part of an electrical wavelength as developed in the system. Obviously, numerous other possible spacing combinations exist, such as $d_1=0.6\lambda$, $d_2=0.9\lambda$, or $d_1=0.9\lambda$, $d_2=1.1\lambda$, which will permit the antenna to exhibit traveling wave characteristics with broadside beaming.

As previously discussed, for broadside beaming the radiator couplings must be arranged so that $k=\beta$, which would be accomplished in this embodiment by an alternate reversal of the couplings of consecutive slots. For maximum smoothness of excitation of the aperture, the densest possible radiator population compatible with element physical size is preferable, and the spacing between elements should be governed accordingly. I have found that pronounced side lobes will appear in the resultant patterns for the wider electrical spacings unless the system is constructed with β significantly larger than β_0 so that these spacings are of the order of 0.5 wavelength in the surrounding medium. In this event the patterns would be substantially the same as those for the shorter electrical spacings on an ordinary transmission system.

For operation of the antenna of FIGURE 1 as a television transmitting antenna, the U.H.F. or V.H.F. audio signal may be introduced via connector **13** while the video signal containing the picture information may be introduced at connector **14**. Such signal coupling will avoid the necessity of a diplexer. To insure high overall attenuation and hence obviate the possibility of reflection and of signal mixing at the ends, the antenna should be several wavelengths long in terms of the median frequency of the signals propagated. The exact length of the antenna is determined by the requirements of the radiation pattern directivity (or gain), the radiation pattern shape and direction, the degree to which the traveling wave must be attenuated to achieve a high antenna efficiency, and to some extent by the frequency band over which the antenna must function.

To provide a clear understanding of the manner in which the slots **22**, **23**, and so forth, are excited, concurrent reference should be made to FIGURES 1 and 2. Each slot, as **22''**, is somewhat shorter than its natural resonant length, and preferably electrically shorter than a half-wave-length to render them moderately selective upon tuning. Metal tabs **26** and **30** are provided at the midpoint of the longitudinal sides of each slot, each tab extending into the slot a distance approximately equal to the thickness of tube **11**. In addition, each tab includes an elongated hole to permit lateral movement of the tabs across the slot relative to the circumference of tube **11**. The tabs thereby provide an adjustable capacitance for slot tuning. Screws **27** and **28** extend into tapped holes

in the tube to maintain the tabs in fixed position for a desired center frequency.

Slot excitation is accomplished through capacitive coupling to the feed line or transmission channel along which the electromagnetic waves travel. To this end, each slot is provided with an adjustable capacitive probe, as **24''**. The probe comprises a metallic disc **29**, suitably connected by way of conductive shaft **31** to a bore located interiorly of screw **28** to permit sliding electrical contact. The proximity of the probe disc **29** to inner tube **12** will determine the degree of coupling between transmission channel and slot. The probe positions are alternately reversed in a regular pattern along consecutive slots to provide the necessary phase relationship of slot excitation by the traveling wave. That is, the probes, for example those probes designated **24, 25, 24', 25', 24''**, and so forth, are located at alternately opposite sides of the slots to which they correspond to place these radiating elements are nearly in phase as possible, consistent with the restrictions imposed on radiating element spacing. In other words, for a wave traveling from slot **22** to slot **23**, the phase at slot **23** relative to that at slot **22** as a result of probe reversal is $(-\beta d+\pi)$ radians. If βd were equal to π, the two elements would be precisely in phase, but since, as previously discussed, this spacing is disallowed for broadside beaming attenuated wave antennas, the radiating elements are excited as nearly in phase as restrictive conditions will permit.

This slot excitation will be better understood by reference to FIGURE 4 which represents, in a qualitative form, the voltage distribution function

$$E(z)=Re\{E(0)\epsilon^{-j\bar{\gamma}z}\epsilon^{jwt}\}$$

of an electromagnetic wave **41** traveling along a smooth dissipative transmission line in the absence of reflections, at some given instant t_1 in time. The phase distribution (not shown) is simply a linear retardation along the antenna axis at the rate β, which is the imaginary component of the complex propagation function $\bar{\gamma}=\bar{\alpha}+j\bar{\beta}$, in a smooth dissipative transmission line. Thus, for example, the phase at point **43**, representing a probe coupling position, lags that at point **42**, a probe position 0.6λ from point **43**, by $0.6\times 2\pi$ radians. Wave **41** is a damped sinusoid, attenuation being in accordance with the value of $\bar{\alpha}$ which is largely dependent upon extraction of energy by the radiators and thus upon radiating element location and configuration. Probes **42** and **43** are separated, for illustrative purposes, by 0.6 wavelengths of the median frequency of the energy applied to the antenna, probes **44** and **45** being similarly separated. Center points **46** and **47** between these probe pairs are separated by a distance corresponding to D in FIGURE 1, for example 1.5 wavelengths.

From an observation of the first quarter cycle of wave **41**, it will be noted that probe **42** is fed, at this particular instant of time, by a wave amplitude just below the peak positive value which the wave had at that probe at a prior instant in time. Probe **43**, located on the opposite side of its respective slot from probe **42** is fed by a somewhat attenuated wave segment approaching its peak negative value. It will be seen that the probe positioning results in as nearly an in phase excitation of the slots corresponding to probes **42** and **43** as is permissible with the discussed slot spacing restrictions. The same analysis may be applied with respect to probes **44** and **45**, and so forth.

A typical voltage pattern obtained by combining the slightly out of phase contributions of the two slots **22** and **23** (FIGURE 1) in a plane containing the antenna axis is shown in FIGURE 6. The difference in their effective phases causes the maximum radiation vector, in the direction of major lobe **59** in this particular case, to be squinted slightly backward from broadside with respect to the internal direction of wave propagation. The difference in amplitudes of excitation of the two slots prevents the occurrence of complete diffraction nulls. The end-on nulls are characteristic of axial slot radiators, and not due to diffraction phenomena. The pattern thus derived is not highly directive because of the limited extent of this sub-array, and the amplitude in the broadside direction is practically the same as that on the beam nose. The beam maximum of a long nominally-broadside traveling wave array of such pairs therefore occurs with great precision in that particular direction in which all the individual pair contributions add in phase. I have found the oblique characterstic of the component pair pattern to be very helpful in obtaining close approximations to such unsymmetrical beam shapes in the axial plane as the so-called cosecant-θ contour, which is highly desirable in television broadcast applications. The amplitudes of the patterns of successive pairs of slots along the array are progressively reduced according to the amount of attenuation introduced, and it is this feature which automatically precludes the occurrence of any diffraction nulls in the vicinity of the main beam.

To stabilize the beam in a given direction relative to the antenna axis with respect to frequency variation about the frequency of interest, it is essential that the radiators reflect into the transmission system impedance or admittance load functions whose reactive components have negative derivatives with respect to frequency over the operating range. The degree to which the beam can be stabilized in direction depends upon the selectivity (Q) of the radiator, the attenuation rate of the traveling wave, the required bandwidth, and the direction of the beam. The individual radiators must either have a moderately high Q, or else be closely coupled, in order to accomplish significant frequency compensation in a low traveling wave antenna; the former condition limiting the bandwidth, and the latter the effective length, or gain. In the preferred embodiment the slot radiators are electrically shorter than a half-wavelength to make them moderately selective upon tuning. The metal tabs **26** and **30** provided on both longitudinal sides at the midpoint of each of the slots permit greater control over tuning of the radiators to resonance at the design frequency. The probe type of exciter is required in the preferred embodiment to provide the necessary type of loading in the system.

With the traveling wave antenna disclosed in FIGURES 1 and 2, the azimuthal beam extends only in a direction substantially broadside of the slots (FIGURE 6). To obtain other types of azimuth patterns, radiating elements may be suitably placed at angular positions about the antenna axis while retaining the specified doubly periodic configuration. Referring now to FIGURE 3, a modification of the previously described antenna may be made to accomplish an omni-directional azimuth pattern. In this embodiment, three slot radiators, **35, 36** and **37**, are located at 120° spacing about the circumference of the outer tube, in place of the single slot at that point along the longitudinal axis depicted in FIGURE 2. Similarly, three such slot radiators will be so spaced about the longitudinal axis at each cross section at which a single slot appears in FIGURE 1. In this manner, there will be three longitudinal columns of aligned recurrent radiators in the doubly periodic pattern, each column spaced 120° from the others. Associated with each of the slots, for example **35, 36, 37**, is a separate capacitive probe of the type illustrated in FIGURE 2. The slots at each cross-section will be excited by probes positioned at an identical side of each slot, as shown, the probe positions being reversed at the next adjacent slots, as previously indicated with respect to FIGURES 1 and 2. Capacitive tuning of the slots is accomplished by tab structure as shown in FIGURE 2, the adjustable capacitance being shown schematically at **39**, for simplicity, for each of the slots in FIGURE 3.

If it is desired to control the azimuth pattern in a different manner, the radiators may be located at other than equally-divided circumferential positions on outer tube **11**. Attachments, such as metallic flares, may also be used

as necessary to assist in pattern control without voiding the scheme. In general, the center of each of the slots in a particular cross-section is located at the same axial point on the transmission line.

Referring now to FIGURE 5, there is shown the voltage distribution function, as noted with respect to FIGURE 4, for an antenna embodiment having radiator probes 51–54, each located on the same side of its respective slot. Probes 51 and 52 are separated, for example, by a distance of 0.9λ, this spacing being alternately repeated, as for 53, 54, as before. For illustrative purposes, the spacing between centers 56, 57, of adjacent radiator slot pairs is shown as 2.0λ. Examination of wave form 55, having the same characteristics as wave form 41 of FIGURE 4, reveals that a positive maximum occurs in proximity to each of probes 51 and 52, and, similarly, in proximity to each of probes 53 and 54, resulting in an approximately in-phase excitation.

It will be understood that the transmission channel for any particular application is not limited to a coaxial arrangement, this having been shown merely for simplicity and clarity, but may take any suitable form which is capable of supporting traveling waves.

While I have described certain preferred embodiments of my invention, it will be apparent that variations of the details of construction which are specifically illustrated and described may be resorted to without departing from the true spirit and scope of the invention as defined by the appended claims.

I claim:

1. A traveling wave antenna comprising a transmission channel for electromagnetic wave propagation, means for coupling a source of said electromagnetic waves to said channel, said waves having a wavelength λ within said channel, a plurality of pairs of recurrent radiators coupled to said channel in approximately in-phase excitation, radiators of each pair being aligned and separated longitudinally along said channel by a distance departing substantially from $n\lambda/2$, where n is an integer, an element in each pair being separated from the adjacent element in the next pair by a distance differing substantially from the first-mentioned distance, and the centers of said recurrent pairs being separated by $N\lambda/2$, where N is approximately an integer and is greater than n, said antenna radiating a beam approaching a broadside beam as N approaches an integer.

2. An antenna comprising a transmission channel capable of supporting traveling waves, means coupling said channel to a source of energy for producing electromagnetic waves having a wavelength λ in said channel, a plurality of pairs of radiating elements aligned in a doubly periodic configuration longitudinally along said channel, said pairs of elements coupled to said channel for substantially cophasal excitation by electromagnetic waves traveling therein, said doubly periodic configuration comprising repetitively consecutive long and short interval longitudinal spacing between successive pairs of said radiating elements wherein said short interval spacing differs from an integral multiple of half wavelengths λ of said waves, and wherein said long interval spacing is the complement of said short interval spacing such that the sum of two consecutive spacings is approximately an integral multiple of said half wavelengths λ.

3. The combination according to claim 2 including means for resonant tuning of said radiating elements.

4. The combination according to claim 2 including means for varying the degree of coupling between said transmission channel and individual ones of said radiating elements.

5. The combination according to claim 2 wherein said transmission channel is a coaxial line and wherein said radiating elements comprise axially aligned longitudinal slots less than a half wavelength λ long in the outer conductor of said line.

6. The combination according to claim 5 including means for adjusting the resonant frequency of each said slots.

7. The combination according to claim 6 wherein said means for adjusting resonant frequency comprises a pair of conductive tabs along the longitudinal edges of each of said slots, said tabs being adjustable to vary the distance therebetween to vary the capacitance across said slots.

8. The combination according to claim 5 including means for varying the degree of coupling between said coaxial line and said slots.

9. The combination according to claim 8 wherein said means for varying the degree of coupling comprises a separate capacitive probe associated with each of said slots, each of said probes being radially aligned and conductively connected to said outer conductor and separately adjustable to vary the capacitive coupling between the respective slot and the inner conductor of said coaxial line.

10. The combination according to claim 9 wherein each of said probes is positioned at approximately the longitudinal midpoint of each of said slots and along successively opposite longitudinal sides thereof such that adjacent pairs of said slots have associated probes located in an alternating reversed pattern along the longitudinal sides thereof.

11. An energy radiator comprising a transmission channel having an axis of symmetry and having at least one path for supporting traveling wave propagation, means for coupling wave energy into one end of said channel, a plurality of pairs of radiating elements coupled to said channel for substantially in-phase excitation in a repetitive doubly periodic array for radiating energy from said wave energy propagating along said channel, said repetitive doubly periodic array having a configuration wherein said radiating element pairs are substantially aligned along said channel axis, said elements in each of said pairs being separated by a distance $d \neq n\lambda/2$ where λ is the wavelength of said wave energy in said channel and n is a positive integer, and corresponding elements of adjacent pairs being separated by a distance $D \doteq N\lambda/2$ where N is a positive integer greater than n.

12. The combination according to claim 11 wherein said plurality of pairs of radiating elements extend along the entire length of said channel, said channel being several wavelengths of said wave energy long, such that said pairs of radiating elements are sufficiently numerous to attenuate substantially all of said propagating wave energy between said one end of said channel and the end remote from said one end whereby there is substantially an absence of reflected wave energy at said remote end.

13. The combination according to claim 12 wherein said channel includes a second traveling wave path, means for coupling wave energy of a wavelength different from said first-mentioned wave energy into said remote end of said channel for propagation along said second path, and wherein said attenuation characteristic of said channel in conjunction with said radiating elements is sufficiently high to retard intermodulation of said first and last mentioned wave energy and to provide isolation between said first and last mentioned coupling means.

14. The combination according to claim 12 wherein is included means for varying the degree of coupling between each of said radiating elements and said channel.

15. The combination according to claim 12 wherein is included means for adjustably tuning the resonant frequency of each of said radiating elements.

16. The combination according to claim 12 wherein said repetitive doubly periodic array comprises a single alignment of radiating elements along said channel axis, and wherein means are provided for varying the degree of coupling between said elements and said channel to produce broadside beaming in the azimuth pattern of said energy radiator relative to said channel axis.

17. The combination according to claim 12 wherein said repetitive doubly periodic array comprises a plurality

of alignments of radiating elements along said channel axis, said alignments being arranged to provide a controlled azimuth pattern for said energy radiator relative to said channel axis.

18. The combination according to claim 17 wherein radiating elements in each of said alignments are positioned at common planes perpendicular to said channel axis, said planes being separated by said distances *d* and D in accordance with said doubly periodic configuration, whereby said radiating elements are both longitudinally and radially alined with respect to said channel axis.

19. A traveling wave antenna comprising transmission means for electromagnetic wave propagation along a longitudinal transmission path, means coupling said transmission means to a source of said waves, a plurality of radiating elements coupled to said transmission means for cophasal excitation by said waves, said radiating elements located in a doubly periodic pattern along said transmission means, wherein said radiating elements are longitudinally spaced in alignment along said transmission path at successive long and short intervals relative to each other,

in a repeating pattern, each interval differing from an integral number of half wavelengths of said waves.

20. The combination according to claim 19 wherein each of said spacing intervals differs from an integral number of half wavelengths by at least one-tenth wavelength of said waves.

21. The combination according to claim 13 in which the separation of radiating elements adjacent said ends of said channel is adjustable to vary the pattern of energy radiated from said channel in accordance with said wavelengths of said first and last-mentioned wave energy.

References Cited

UNITED STATES PATENTS

2,756,421	7/1956	Harvey et al.	343—770
3,100,300	8/1963	Sletten	343—771
3,308,467	3/1967	Morrison	343—771

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