

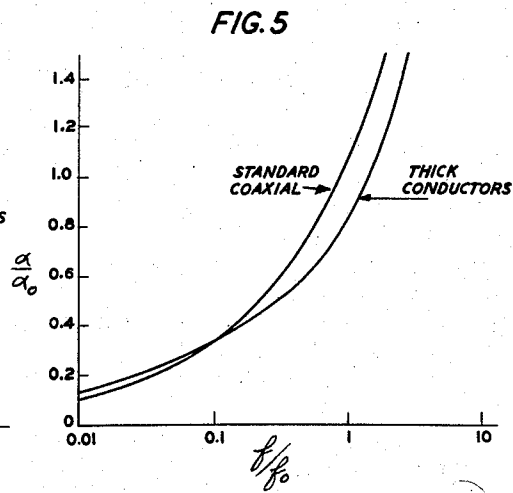
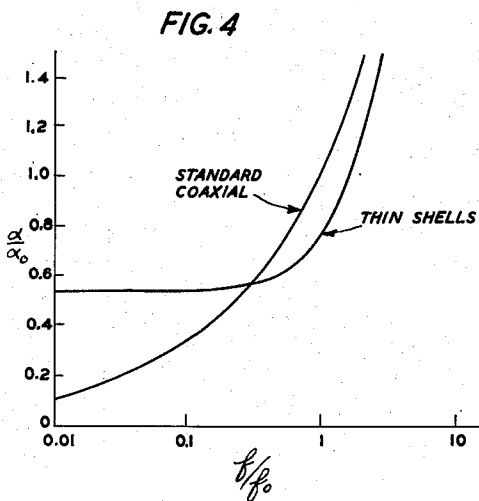
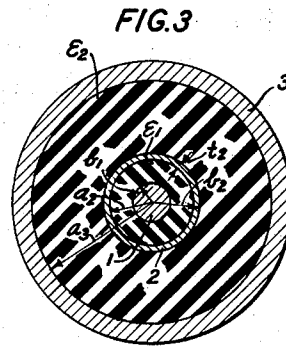
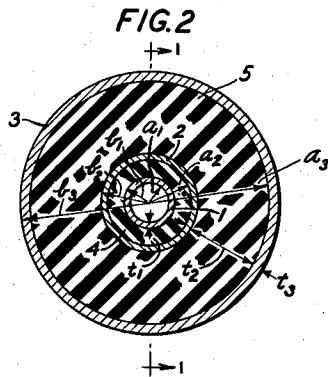
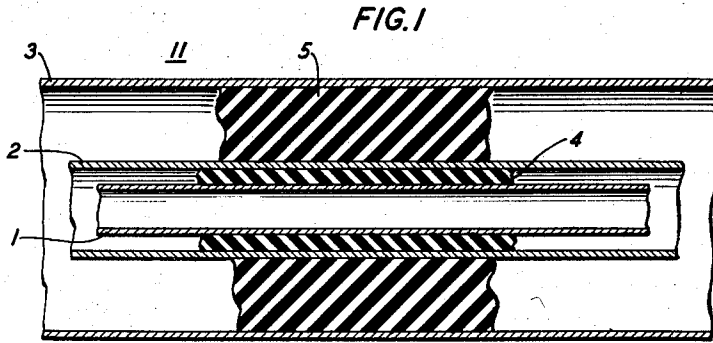
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2,924,795

LOW-LOSS TRANSMISSION LINE

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2,924,795

LOW-LOSS TRANSMISSION LINE

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15 Claims. (Cl. 333—96)

This invention relates to conductors of high-frequency waves, and, more particularly, to conductors of the concentric or coaxial type.

An object of this invention is to reduce skin effect in transmission lines operating at frequencies up to predetermined high-frequency limits.

A more specific object is to provide an efficient yet inexpensive means of improving the attenuation-bandwidth characteristics of insulation filled coaxial cables.

The problem of skin effect is well known to the art and will be outlined here only because the present invention is concerned with this problem. A more complete explanation thereof can be found beginning at page 196 of "Fields and Waves in Modern Radio" by S. Ramo and J. R. Whinnery, John Wiley & Sons, Inc., 1946. Briefly stated though, skin effect is the increase with frequency of the resistance of a conductor caused by the tendency of alternating currents to crowd more and more toward the surface of the conductor as their frequency is increased. The surfaces toward which these currents tend to crowd are those surfaces nearest the fields producing them.

Expressed mathematically, the distribution of current density in a semi-infinite plane conductor where z is the direction of current flow and x is the direction normal to the surface of the conductor is as follows:

$$i_x = i_0 e^{-\frac{x}{\delta}} \frac{x}{\delta} \quad (1)$$

where i_0 is the density of current flowing in the z direction at the surface of the conductor; where the base e is the base of natural logarithms,

$$\delta = \frac{1}{\sqrt{\pi f \mu g}}$$

where f is frequency, μ is permeability of the conductor, and g is its conductivity. An example of a semi-infinite plane conductor to which Equation 1 is applicable is a plane conductor of infinite depth below the surface, infinite width, and infinite length. From Equation 1 it is apparent that at a distance δ below the surface of the conductor the current i_x has decreased to $1/e$ of its value at the surface. This distance δ is called the skin depth of penetration of current into the conductor at a given frequency.

In long distance wire communication two interrelated aspects of skin effect become important. The first of these is the increase in power losses in the transmission lines which necessitates signal amplification at frequent intervals along the lines. The second is the nonuniformity of the transmission characteristics of the lines for differing signal frequencies. A very well known way of dealing with the first aspect of the problem is to plate a thin layer of highly conductive metal such as silver upon a poorly conducting member such as a less expensive base metal. In this way the power losses may be reduced in the neighborhood of the highest utilized frequency by practicing the procedures described in U.S. Patent 2,052,317 issued to S. A. Schelkunoff, but at best

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there is only inadequate equalization of these losses at various low frequencies. Conversely, frequency equalization has been achieved at the expense of increased power losses in the way described in United States Patent 1,817,964 to J. R. Carson et al. Several structures disclosed in this patent resemble in appearance those to be described hereinafter in connection with the present invention, but it should be understood that none of the former structures succeeds in reducing power losses while at the same time achieving frequency equalization.

So far as is known the most efficient way of dealing with both aspects of the skin effect problem is disclosed in copending application to A. M. Clogston filed March 7, 1951, Serial No. 214,393 (now United States Patent 2,769,148, issued October 30, 1956). The structures disclosed in that application utilize as a means of reducing wave attenuation a number of thin, closely spaced conductors spaced apart and parallel to each other by insulating material. In order for high frequency waves to penetrate through out such a laminated conducting medium, which is a condition for most efficient operation, it is necessary that each individual conductor be at least as thin as the skin depth of penetration of waves in the conductor metal at the frequency of the waves. It is also necessary that the velocity of propagation of waves in the region surrounding the laminated medium be made substantially the same as the velocity within this medium. When these requirements are satisfied it is then possible to have a cable in which the power loss in the laminated conductor is appreciably less than the loss in a solid conductor of the same size and, from very low frequencies up to frequencies at which the skin depth becomes comparable to the thickness of the individual conductors, wave attenuation is almost uniform. By increasing the number of thin conductors used while at the same time making each conductor thinner, the attenuation-bandwidth characteristics of the laminated medium can be improved, within certain limits, in any required degree. Thus, one long distance cable built according to the above Clogston disclosure can theoretically transmit all the signals now being transmitted by a very large number of conventional long distance lines. In many applications, however, it is not economical to employ such a high-transmission-capacity cable. For this reason it would be desirable to have available a transmission line possessing attenuation-bandwidth characteristics substantially superior to those of conventional transmission lines previously known, but yet being only slightly more difficult to manufacture than these lines. The present invention, prompted by this need, is based upon the discovery that a two-conductor insulated transmission line can be improved substantially by the proper insertion into the electromagnetic field between these conductors of a single thin conductor together with the right kind of insulating material.

In accordance with the present invention in one specific embodiment thereof, a hollow cylindrical conductor having a wall thickness less than the skin depth at a predetermined frequency is suitably placed between and axially aligned with the inner and outer conductors of a coaxial cable. Insulating material having a low dielectric constant is used to separate this thin conductor from the inner conductor and different insulating material having a somewhat higher dielectric constant is used to separate it from the outer conductor. The exact value of the dielectric constant of this latter material and the thickness of the thin conductor together with certain other physical dimensions and relationships in this embodiment are critical in reducing cable attenuation. These optimum dimensions and relationships will appear in the following more complete description of the invention and it suffices here to say that they exist.

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When a structure is made satisfying these requirements its attenuation-bandwidth characteristics will be substantially better than those for a conventional non-airfilled coaxial cable of the same outer diameter.

The advantages of the present invention will be gathered from and a better understanding of its general nature will best be gained by the careful study of the following detailed description given in connection with the accompanying drawings, in which:

Fig. 1 is a sectional view of a preferred embodiment of the invention taken along the line 1-1 of Fig. 2;

Fig. 2 is a cross-sectional view of a preferred embodiment of the invention;

Fig. 3 is a cross-sectional view of another preferred embodiment;

Fig. 4 is a plot of the attenuation vs. frequency characteristic of the cable of Figs. 1 and 2 as compared with a standard coaxial cable; and

Fig. 5 is a plot of the attenuation vs. frequency characteristic of the cable of Fig. 3 as compared with a standard coaxial cable.

In explanation of the invention it is necessary to give a brief, mathematical treatment of a transmission system consisting of three coaxial cylindrical conductors separated by two insulating members. An analysis of a system of n coaxial conductors may be found in "Transmission Characteristics of the Submarine Cable," by John R. Carson and J. J. Gilbert, in The Journal of the Franklin Institute, vol. 192, pp. 705-735, December 1921. The instant analysis may be termed a special case of a system of n conductors, with a more thorough study of the effect of the proportions and materials of the cable on the transmission characteristics thereof. The principles of the present invention may be more fully understood in connection with a detailed treatment of the three conductor cable. In the following treatment, the MKS Rational System of units is used throughout except where specified otherwise.

Turning now to the drawings, there is shown in Figs. 1 and 2 a three conductor coaxial cable 11 having an inner cylindrical conducting shell 1, an intermediate cylindrical conducting shell 2 coaxial with shell 1, and an outer cylindrical conducting shell 3, coaxial with shells 1 and 2. Shells 1, 2, and 3 are made of suitable conducting material, such as copper, and shell 1 may be advantageously formed by copper plating an iron, steel, or other suitable material core, not shown. In addition, shell 3 may be surrounded by a reasonably high impedance shield, not shown, without detrimentally affecting the operation or the characteristics of the cable. Shell 2 is spaced from and maintained in coaxial relationship with shell 1 by insulating material 4 having a dielectric constant ϵ_1 and shell 3 is spaced from shell 2 by insulating material 5 having a dielectric constant ϵ_2 . Materials which are suitable for insulator 4 are to some extent determinative of what constitutes a suitable material for insulator 5, as will be explained hereinafter, hence a discussion of these materials is postponed until later.

In Figs. 1 and 2, the inner radius of the i th conducting member ($i=1, 2, 3$) is denoted by a_i , its outer radius by b_i , and its thickness by $t_i=b_i-a_i$. For purposes of this analysis, we consider electromagnetic fields which propagate along the cable according to the factor

$$\exp(-\Gamma z + j\omega T) \quad (2)$$

where z is distance along the axial direction, Γ is the propagation constant, ω is 2π times the cyclic frequency, and T is time. Furthermore, let I_1 be the current in amperes flowing in the positive z direction in conductor 1, and let I_3 be the current in amperes flowing in the positive z direction in conductor 3, in which case the current in the positive z direction in conductor 2 is $I_2=-(I_1+I_3)$, neglecting axial displacement currents in the insulators. It may be shown, as in the aforementioned Carson and Gilbert paper, that the axial electric

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field components at the surfaces of the insulators are related as follows:

$$E_z(a_2) - E_z(b_1) = -[\Gamma^2 + \omega^2 \mu_1 \epsilon_1] \frac{I_1}{Y_1} \quad (3)$$

$$E_z(a_3) - E_z(b_2) = [\Gamma^2 + \omega^2 \mu_2 \epsilon_2] \frac{I_3}{Y_2} \quad (4)$$

where

$$Y_1 = \frac{2\pi j \omega \epsilon_1}{\ln(a_2/b_1)} \quad (5)$$

$$Y_2 = \frac{2\pi j \omega \epsilon_2}{\ln(a_3/b_2)} \quad (6)$$

and μ_1 is the permeability of inner insulator 4 in henries per meter, μ_2 is the permeability of outer insulator 5, ϵ_1 is the dielectric constant of the inner insulator 4 in farads per meter, ϵ_2 is the dielectric constant of the outer insulator 5 and $E_z(\rho)$ is the component of electric field in volts per meter parallel to the axis at a distance ρ therefrom. In the present analysis, it is assumed that the conductivities of the dielectrics are zero, but both dielectric and magnetic dissipation, if present, can be accounted for by assigning complex values to the dielectric constants and permeabilities.

When a hollow cylindrical conductor of inner radius a and outer radius b carries a given total current which returns on one or more conductors coaxial with it, the longitudinal electric field intensities at its surface are related to the current by certain factors called surface and transfer impedances. The impedances of hollow cylindrical shells have been discussed in the paper "Electromagnetic Theory of Coaxial Transmission Lines and Cylindrical Shields," by S. A. Schelkunoff (Bell System Technical Journal, vol. 13, pp. 532-579, October 1934). If, of the total current $I_a + I_b$ flowing in the hollow conductor, the part I_a returns inside and the part I_b outside, the longitudinal electric fields at the inner and outer surfaces are given by:

$$E_z(a) = Z_{aa}I_a + Z_{ab}I_b \quad (7)$$

$$E_z(b) = Z_{ba}I_a + Z_{bb}I_b \quad (8)$$

where Z_{aa} and Z_{bb} are surface impedances in ohms and $Z_{ab}=Z_{ba}$ is the transfer impedance. While exact formulae for the impedances may be obtained from the aforementioned Schelkunoff article, they are somewhat unwieldy and require access to tables of Bessel functions in computing the values of the impedances. Useful approximations sufficient for the present analysis are contained in a paper entitled "Über den Wechselstromwiderstand von Geraden Drähten" by S. Ekelöf appearing in Elektrische Nachrichtentechnik, vol. 10, No. 3, pp. 115-122, 1933. In that paper, Ekelöf gives equations which refer to a conducting cylinder whose inner and outer radii are both large compared to the skin depth, and whose thickness $t=b-a$ is small compared to the mean radius $\frac{1}{2}(a+b)$. In the present notation these formulae read:

$$\frac{Z_{aa}}{Z_{ab}} = \cosh \sigma t \left\{ 1 + \left[\frac{t}{2b} + \frac{3t^2}{8b^2} \right] \left[1 - \frac{\tanh \sigma t}{t} \right] + \frac{3t^3}{16b^3} \left[\frac{5}{3} - \frac{2 \tanh t}{\sigma t} + \frac{1}{\sigma^2 t^2} \left[1 - \frac{\tanh \sigma t}{\sigma t} \right] \right] + \dots \right\} \quad (9)$$

$$\frac{Z_{bb}}{Z_{ba}} = \cosh \sigma t \left\{ 1 - \left[\frac{t}{2a} - \frac{3t^2}{8a^2} \right] \left[1 - \frac{\tanh \sigma t}{\sigma t} \right] - \frac{3t^3}{16a^3} \left[\frac{5}{3} - \frac{2 \tanh \sigma t}{\sigma t} + \frac{1}{\sigma^2 t^2} \left[1 - \frac{\tanh \sigma t}{\sigma t} \right] \right] + \dots \right\} \quad (10)$$

$$\frac{1}{\pi g(a+b)tZ_{ab}} = \frac{2b}{(a+b)} \frac{\sinh \sigma t}{\sigma t} \left\{ 1 - \frac{t}{a} + \frac{3t^2}{8a^2} \left[1 + \frac{\coth \sigma t}{\sigma t} - \frac{1}{\sigma^2 t^2} \right] - \frac{3t^3}{16a^3} \left[\frac{5}{3} + 3 \left\{ \frac{\coth \sigma t}{\sigma t} - \frac{1}{\sigma^2 t^2} \right\} \right] + \dots \right\} \quad (11)$$

where g is conductivity in mhos per meter and

$$\sigma = \sqrt{j\omega\mu g} = \frac{1+j}{\delta}$$

In the three conductor cable shown in Figs. 1 and 2, the impedances of the i th conductor are denoted by

$$Z_{aa}^{(i)}, Z_{bb}^{(i)}, \text{ and } Z_{ab}^{(i)}$$

Then, from Equations 7 and 8,

$$E_z(b_1) = Z_{bb}^{(1)} I_1 \quad (12)$$

$$E_z(a_2) = Z_{aa}^{(2)} I_1 - Z_{ab}^{(2)} I_3 \quad (13)$$

$$E_z(b_2) = Z_{bb}^{(2)} I_1 - Z_{bb}^{(2)} I_3 \quad (14)$$

$$E_z(a_3) = Z_{aa}^{(3)} I_3 \quad (15)$$

Substituting these relations into Equations 3 and 4 we get

$$[(\Gamma^2 + \omega^2 \mu_1 \epsilon_1) - Y_1(Z_{bb}^{(1)} + Z_{aa}^{(2)})] I_1 - Y_1 Z_{ab}^{(2)} I_3 = 0 \quad (16)$$

$$-Y_2 Z_{ab}^{(2)} I_1 + [(\Gamma^2 + \omega^2 \mu_2 \epsilon_2) - Y_2(Z_{bb}^{(2)} + Z_{aa}^{(3)})] I_3 = 0 \quad (17)$$

which are a pair of homogeneous linear equations in terms of the currents I_1 and I_3 . These equations may be satisfied by values of I_1 and I_3 which are not both zero if and only if the determinant of the coefficients vanishes. Setting the determinant equal to zero gives a quadratic equation in Γ^2 :

$$\begin{aligned} &(\Gamma^2 + \omega^2 \mu_1 \epsilon_1)(\Gamma^2 + \omega^2 \mu_2 \epsilon_2) \\ &- [Y_2(Z_{bb}^{(2)} + Z_{aa}^{(3)})(\Gamma^2 + \omega^2 \mu_1 \epsilon_1) + Y_1(Z_{bb}^{(1)} + Z_{aa}^{(2)})(\Gamma^2 + \omega^2 \mu_2 \epsilon_2)] \\ &+ Y_1 Y_2 [Z_{bb}^{(1)} Z_{bb}^{(2)} + Z_{bb}^{(1)} Z_{aa}^{(3)} + Z_{aa}^{(2)} Z_{bb}^{(2)} + Z_{aa}^{(2)} Z_{aa}^{(3)} - (Z_{ab}^{(2)})^2] = 0 \end{aligned} \quad (18)$$

Equation 18 is a perfectly general relation which must be satisfied by the propagation constants of the transmission line modes on any system of three coaxial cylindrical conductors separated by two insulators. It has two roots, which may be denoted by Γ_1 and Γ_2 . (The roots $-\Gamma_1$ and $-\Gamma_2$ which correspond to backward traveling waves also satisfy the equation.) The propagation constants may be separated into real and imaginary parts:

$$\Gamma_1 = \alpha_1 + j\beta_1 \text{ and } \Gamma_2 = \alpha_2 + j\beta_2 \quad (19)$$

which represent attenuation and phase constants respectively.

One purpose of the present invention is to obtain, with a three-conductor transmission system of predetermined outside diameter, a lower attenuation constant at a specified frequency than may be obtained with a conventional two-conductor coaxial system of the same outside diameter at the same frequency. Inasmuch as there are two transmission-line modes in the three conductor system, as shown by Equations 18 and 19, it is necessary that that mode which has the smaller attenuation constant be utilized, which, for convenience, may be designated as mode 1. The corresponding attenuation constant α_1 may be minimized by proper choice of materials and proportions in the three-conductor cable.

As an example, assume that, in the cable of Figs. 1 and 2, the outside diameter and the design frequency are given, while the proportions and materials used may be varied within broad limits set by engineering considerations, such as mechanical strength, dielectrics available, and the range of permeabilities available. Inasmuch as the attenuation constant α_1 is dependent on a great number of parameters, it is impractical to write out the mathematical expression for α_1 . Instead, optimizing of the parameters to achieve a minimum α_1 may be conveniently accomplished by a process of successive approximations, as follows:

When all of the parameters are held constant except one, such as, for example, the dielectric constant ϵ_2 of insulator 5, ϵ_2 is varied until a minimum α_1 is reached. The process is then repeated, varying one of the other parameters, until all of the parameters have been optimized to the first approximation. The process is then repeated until successive approximations cease to have any appreciable effect on the attenuation constant.

A more efficient procedure for minimizing α_1 than the above method is found in "On the Experimental Attainment of Optimum Conditions," by G. E. P. Box and K. B. Wilson, Journal of the Royal Statistical Society, Series B, vol. 13, No. 1, pp. 1-45, 1951. In that paper, it is shown that the quantity to be minimized should be evaluated for several different choices of the variables in the neighborhood of the expected minimum. A second degree function of the variables is then fitted, by the method of least squares, to the computed points. By way of illustration, if there were only one variable factor, α_1 would be evaluated at three or more points and a parabolic approximating curve fitted to the points. Setting the first partial derivatives of the approximating function equal to zero, a set of simultaneous linear equations is obtained which, when solved, yield the minimum of the approximating function. If this minimum lies outside of the region which contains the computed points, the variables are adjusted in the direction of the minimum and a new approximating function is fitted. This process is repeated until the minimum of the approximating function lies within the region of computed points, which point is then the minimum value of α_1 , and the values of the variables may be, in effect, read off.

As an example of the application of the foregoing, assume that all conductors have the conductivity of copper, namely $g = 5.8 \times 10^7$ mhos per meter, that

$$\mu_1 = \mu_2 = \mu_0 = 4\pi \times 10^{-7}$$

henries per meter, and that ϵ_1 is fixed. Since attenuation is proportional to the square root of the dielectric constant, it is obviously desirable to select a material having the lowest possible dielectric constant. In the instant case, we are dealing with two dielectric materials having different dielectric constants, one of which (ϵ_1) will always be less than the other (ϵ_2). By selecting a material with as low a dielectric constant as is available, and designating this as ϵ_1 , we are then assured that by holding ϵ_1 constant, ϵ_2 will also be minimized for the particular case under consideration. The remaining parameters are combined in dimensionless ratios defined as follows:

$$\frac{D}{\delta_0} = 384.4 D \sqrt{f_0} \quad (20)$$

$$\Delta = \frac{\epsilon_2}{\epsilon_1} - 1 \quad (21)$$

$$BW (\text{bandwidth}) = \frac{(\alpha_0)^2}{(\alpha)} = \frac{f_0}{f} \quad (22)$$

where D is the outside diameter of the cable in inches, f_0 is the design frequency in megacycles, δ_0 is skin depth of the conductors in inches, and α_0 is the attenuation constant of a conventional coaxial cable whose conductors have the optimum ratio 3.59:1. At the frequency f_0 the improved cable has an attenuation α and an ordinary coaxial cable has an attenuation α_0 , while f is the lower frequency at which the ordinary coaxial cable has an attenuation α . It can be seen that "bandwidth" as here used does not refer to a range of frequencies, but to a ratio of frequencies which defines the improved attenuation characteristics of the cable of the invention relative to an ordinary coaxial cable. By using the foregoing method of optimizing, the optimum relationships were

found for different values of the ratio (20). The results are shown in the following table:

Table I

$\frac{D}{\delta_0}$	200	400	600
b_1/b_3	0.3303	0.3089	0.3005
b_1/b_2	0.2184	0.2224	0.2249
t_1/δ_0	1.5183	1.5444	1.5529
t_2/δ_0	0.7350	0.7494	0.7549
t_3/δ_0	1.5324	1.5568	1.5655
Δ	0.06777	0.04683	0.03645
α/α_0	0.8117	0.7690	0.7525
BW	1.5179	1.6912	1.7675

Analysis of the above table reveals several interesting facts. For optimum performance, the inner and outer conductors 1 and 3 of Figs. 1 and 2 must be approximately 1.5 times the skin depth at the design frequency, while the intermediate conductor 2 must be approximately 0.75 times the skin depth. In addition, ϵ_2 will have a different value from ϵ_1 . When these conditions obtain, attenuation at the design frequency is seen to decrease to as little as 75 percent of that of a conventional coaxial cable within the range of calculations performed in the example, while bandwidth increases by as much as 75 percent within the same range. It is obvious from Table I that the three conductor cable of Figs. 1 and 2 offers greatly improved performance over that of the conventional coaxial cable. In Fig. 4 there is shown a plot of the attenuation-frequency characteristic of the cable of Figs. 1 and 2 (i.e., the thin shell embodiment) for $D/\delta_0=400$, and, for comparison purposes, the equivalent characteristic of a standard coaxial cable, where α_0 is the attenuation constant of the standard coaxial cable with dielectric constant ϵ_1 , at the design frequency f_0 . It can be seen from an examination of Fig. 4 that the cable of Figs. 1 and 2 gives a tremendously improved attenuation vs. frequency characteristic, as manifested by the exceptional bandwidth wherein the attenuation remains substantially constant.

In Fig. 3 there is shown a second preferred embodiment of the present invention. For simplicity in applying the mathematical relations, elements and dimensions in Fig. 3 are designated by the same reference characters as like elements and dimensions in Fig. 2. The cable of Fig. 3 differs from that of Fig. 2 in that it has a center conductor of solid wire and an outer conductor made of an electrically thick tube, thus providing both mechanical strength and shielding for the cable. For the cable of Fig. 3, the foregoing mathematical development is applicable except in the equations of Z_{aa} and Z_{bb} . When the radius b of a solid cylindrical conductor is large compared to the skin depth, the surface impedance is given by:

$$Z_{bb} = \frac{\eta}{2\pi b} \frac{P_0(\sigma b)}{P_1(\sigma b)} \quad (23)$$

where

$$P_0(\sigma b) = 1 + \frac{1}{8\sigma b} + \frac{9}{128\sigma^2 b^2} + \frac{225}{3072\sigma^3 b^3} + \dots \quad (24)$$

$$P_1(\sigma b) = 1 - \frac{3}{8\sigma b} - \frac{15}{128\sigma^2 b^2} - \frac{315}{3072\sigma^3 b^3} - \dots \quad (25)$$

and

$$\eta = \sqrt{\frac{j\omega\mu}{\sigma}} = \frac{(1+j)}{\sigma} \quad (26)$$

Also the surface impedance of an electrically thick, homogeneous, hollow conductor of inner radius a is given by:

$$Z_{aa} = \frac{\eta}{2\pi a} \frac{Q_0(\sigma a)}{Q_1(\sigma a)} \quad (27)$$

where

$$Q_0(\sigma a) = 1 - \frac{1}{8\sigma a} + \frac{9}{128\sigma^2 a^2} - \frac{225}{3072\sigma^3 a^3} + \dots \quad (28)$$

$$Q_1(\sigma a) = 1 + \frac{3}{8\sigma a} - \frac{15}{128\sigma^2 a^2} + \frac{315}{3072\sigma^3 a^3} - \dots \quad (29)$$

Using the optimizing process explained in the foregoing, the following results are obtained for the cable of Fig. 3, where D is now the inside diameter of the outer conductor.

Table II

$\frac{D}{\delta_0}$	200	400	600
b_1/a_3	0.3280	0.3047	0.2955
b_1/a_2	0.2125	0.2166	0.2191
t_2/δ_0	0.7902	0.8020	0.8071
Δ	0.0736	0.05092	0.03995
α/α_0	0.8475	0.8076	0.7913
BW	1.3922	1.533	1.5970

Since the cable of Fig. 3 is substantially like a coaxial cable with the addition of a thin walled intermediate conductor, the advantageous effect of the intermediate conductor can be readily appreciated by an examination of Table II. It is evident that the third conductor, used with the required two dielectrics in an optimally proportioned cable, adds roughly 50 percent to the bandwidth of the coaxial cable. In the case of the cable of Figs. 1 and 2, another 15 percent increase is obtained. A study of Table II reveals certain facts which were evident in the study of Table I. Thus, the thickness of the intermediate conductor is again within the range of 70–80 percent of the skin depth. Also, the dielectric constants ϵ_1 and ϵ_2 must differ by a small amount. In Fig. 5 there is shown a plot of the attenuation versus frequency characteristic of the cable of Fig. 3 for $D/\delta_0=400$, and, for comparison purposes, the equivalent characteristic of the standard coaxial cable having the same constants as the standard coaxial cable of Fig. 4. It can readily be seen that the cable of Fig. 3 gives an increased bandwidth over the standard coaxial cable.

Comparison of the curves of Figs. 4 and 5 reveals that the cable of Figs. 1 and 2 gives an improved performance over the cable of Fig. 3. In practice, where both mechanical strength and adequate shielding are highly desirable, the performance characteristics of the cable of Figs. 1 and 2 may be achieved by making both the inner and outer conductors of some appropriate material such as, for example, iron, and plating the outer surface of the inner member and the inner surface of the outer member with a highly conductive material such as copper to the required thickness for the cable of Figs. 1 and 2.

The foregoing optimizations were based upon most of the important parameters being variable. In practice, it might be necessary to confine the insulating materials, for instance, to easily obtainable materials, such as polyethylene and polystyrene. Thus ϵ_1 could be 2.26 (polyethylene) and ϵ_2 could be 2.45 (polystyrene). In this case, Δ would have the value 0.084. Furthermore, in applications where the center conductor is to carry D.C. power current or where the center conductor is copper-plated steel for mechanical strength, it may become necessary to make the center conductor approximately equal in size to the center conductor of a coaxial cable. Thus, for example, the inner and outer conductors may have the same outside diameters as a coaxial cable, and

$$b_1/b_3 = \frac{1}{3.59} = 0.2785$$

for the cable of Figs. 1 and 2, and

$$b_1/a_3 = 0.2785$$

for the cable of Fig. 3. With these limitations on the proportions of the cable, optimizing yields the following results:

Table III (cable of Figs. 1 and 2)

$\frac{D}{\delta_0}$	200	400	600
b_2/b_0	0.3687	0.3303	0.3146
t_1/δ_0	1.5308	1.5598	1.5698
t_2/δ_0	0.6882	0.7265	0.7355
t/δ_0	1.5330	1.5577	1.5671
α/α_0	0.8265	0.7878	0.7751
BW	1.4640	1.6114	1.6646

Table IV (cable of Fig. 3)

$\frac{D}{\delta_0}$	200	400	600
b_2/a_0	0.3758	0.3344	0.3176
t_2/δ_0	0.7249	0.7639	0.7753
α/α_0	0.8643	0.8276	0.8154
BW	1.3385	1.4599	1.5042

While the results are not as good as the results obtainable when Δ may be varied freely, nonetheless, as a practical matter, the results obtainable are much better than the performance of a conventional coaxial cable. As was the case with Tables I and II, the thickness of the intermediate conductor varies in the neighborhood of 75 percent of the skin depth, while the outer and inner cylindrical shells in the cable of Figs. 1 and 2 are approximately 1.5 times the skin depth.

Throughout the analysis of the characteristics of the embodiments shown in the figures, the permeabilities of the insulating materials have been taken to be equal to each other and to the permeability of air. Examination of the equations, and particularly Equation 18, reveals that μ is a factor equally as important as the dielectric constant ϵ . The same or similar results can be obtained by varying the permeabilities instead of the dielectric constants, or, where such materials may be available, both factors may be made to differ. Applicants do not intend to limit themselves to the case where only the dielectric constants of the insulators are made to differ, but intend to include within the scope of the invention the case where the permeabilities are made to differ, and also the case where both the permeabilities and the dielectric constants are made to differ. Inasmuch as the characteristic wave propagation velocity of a material is proportional to

$$\frac{1}{\sqrt{\mu\epsilon}}$$

where μ is the permeability of the material and ϵ is the dielectric constant thereof, it can readily be seen that by making the permeabilities or the dielectric constants of the insulating materials differ, the insulating materials will have different characteristic wave propagation velocities.

Since in practice all proportions of the cable are likely to deviate from their optimum values, it is of interest to note that fairly reasonable tolerances are permissible. For example, in the second column of Table I, if only Δ were variable, for a 0.5 percent variation in α/α_0 the ratio ϵ_2/ϵ_1 could lie anywhere between 1.040 and 1.053. In general, the increase in α/α_0 for a small variation of any one of the cable parameters away from its optimum value is proportional to the square of the variation of the parameter.

While applicants have shown and described two embodiments of their invention, it is obvious to those skilled in the art that numerous other embodiments are possible without departing from the spirit and scope of the appended claims.

What is claimed is:

1. A low-loss transmission line for electromagnetic waves comprising inner and outer members of conducting material, a single thin member of conducting material between said inner and said outer members, said thin member being separated from said inner member by first insulating material and being separated from said outer member by second insulating material, the characteristic wave propagation velocity of said first insulating material being greater than that of said second insulating material, and the thickness of said thin member is between one-half and one skin depth at the highest frequency of operation of the transmission line.

2. A low-loss transmission line for electromagnetic waves as claimed in claim 1 wherein the dielectric constant of said first insulating material is different from the dielectric constant of said second insulating material.

3. A low-loss transmission line for electromagnetic waves as claimed in claim 1 wherein the permeability of the first insulating material is different from the permeability of the second insulating material.

4. A low-loss transmission line for electromagnetic waves as claimed in claim 1 wherein both the dielectric constant and the permeability of the first insulating material differ from the corresponding constants of said second insulating material.

5. A low-loss transmission line for electromagnetic waves comprising an inner cylindrical member of conducting material, and an outer cylindrical member of conducting material coaxial with said inner member, a single thin member of conducting material between said inner and outer members and coaxial therewith, said thin member being separated from said inner member of first insulating material and being separated from said outer member by a second insulating material, the intrinsic wave propagation velocity of said first insulating material being greater than that of said second insulating material, said thin member having a thickness greater than one-half and less than one skin depth at the highest frequency of waves to be transmitted.

6. A low-loss transmission line for electromagnetic waves as claimed in claim 5 wherein the dielectric constant of said first insulating material is different from the dielectric constant of said second insulating material.

7. A low-loss transmission line for electromagnetic waves as claimed in claim 5 wherein the permeability of the first insulating material is different from the permeability of the second insulating material.

8. A low-loss transmission line for electromagnetic waves as claimed in claim 5 wherein both the dielectric constant and the permeability of the first insulating material differ from the corresponding constants of said second insulating material.

9. A low-loss transmission line for electromagnetic waves as claimed in claim 5 wherein the inner conducting member is a solid cylindrical conductor and the outer conducting member is a hollow cylinder.

10. A low-loss transmission line for electromagnetic waves as claimed in claim 5 wherein both the inner and outer conducting members are hollow cylindrical conductors.

11. A low-loss transmission line for electromagnetic waves according to claim 5 wherein the spacing between said thin member and said inner conducting member is less than the spacing between said thin member and said outer conducting member.

12. A low-loss transmission line for electromagnetic waves comprising an inner hollow cylindrical member of conducting material, an outer cylindrical member of conducting material coaxial with said inner member, a single thin intermediate member of conducting material between said inner and outer members and coaxial therewith, said thin member being separated from said inner member by a first insulating material and being separated from said outer member by a second insulating material,

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the intrinsic wave propagation velocity of said first insulating wave material being greater than that of said second insulating material, said thin member having a thickness greater than one-half and less than one skin depth at the highest frequency of waves to be transmitted, said inner member having a thickness greater than one and less than two skin depths thick at said frequency and said outer member having a thickness greater than one and less than two skin depths thick at said highest frequency.

13. A low-loss transmission line for electromagnetic waves comprising a solid inner member of conducting material, an outer cylindrical member of conducting material coaxial with said inner member, a single thin intermediate member of conducting material between said inner and outer members and coaxial therewith, said thin member being separated from said inner member by a first insulating material and being separated from said outer member by a second insulating material, the intrinsic wave propagation velocity of said first insulating material being greater than that of said second insulating material, said thin member having a thickness greater than one-half and less than one skin depth at the highest frequency of waves to be transmitted, the outside diameter of said intermediate member being greater than one and one-third times the outside diameter of said inner member.

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14. A low-loss transmission line for electromagnetic waves comprising a solid inner member of conducting material, the surface of said inner member being coated with a conducting material of greater conductivity than that of the material of said inner member, an outer cylindrical member of conducting material coaxial with said inner member, the inner surface of said outer member being coated with a conducting material of greater conductivity than that of the material of said outer member, a single thin intermediate member of conducting material between said inner and outer members and coaxial therewith, said thin member being separated from said inner member by a first insulating material and being separated from said outer member by a second insulating material, the intrinsic wave propagation velocity of said first insulating material being greater than that of said second insulating material, said thin member having a thickness greater than one-half and less than one skin depth at the highest frequency of waves to be transmitted.

15. A low-loss transmission line as claimed in claim 14 wherein the coatings on the inner and outer members each have a thickness greater than one and less than two skin depths at said highest frequency of operation.

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