Oct. 31, 1967 G. W. ZIEGLER, JR 3,350,666

COAXIAL CONNECTOR

Filed April 30, 1963 4 Sheets-Sheet 1

Fig. 1

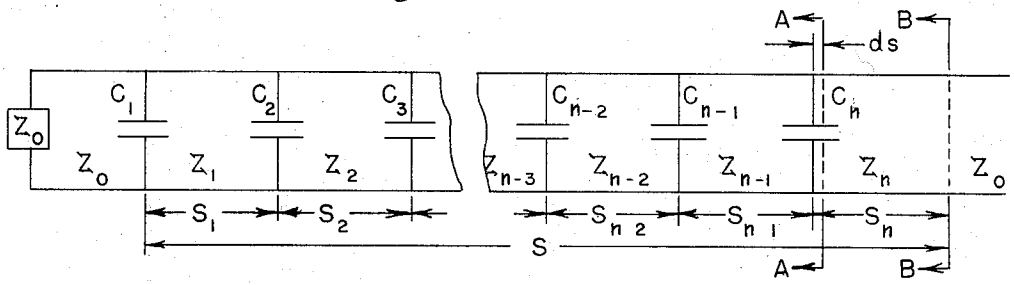


Fig. 2

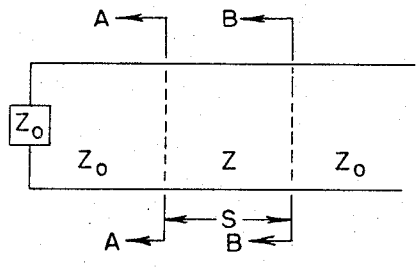


Fig. 4

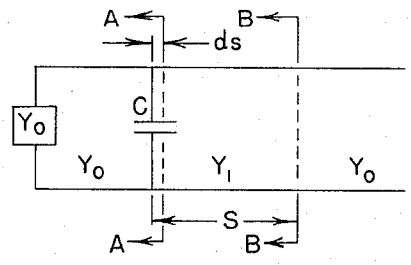


Fig. 3

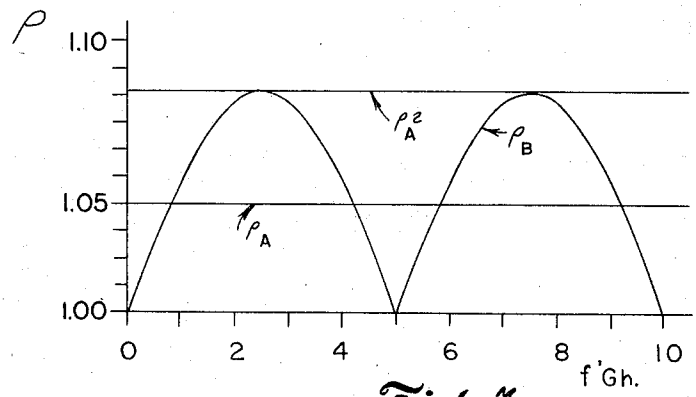
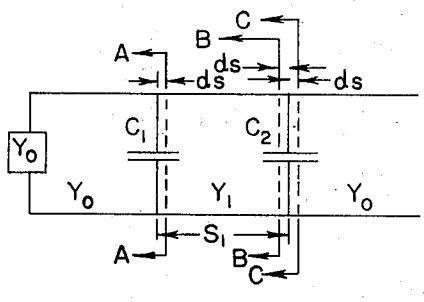


Fig. 5



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Oct. 31, 1967 G. W. ZIEGLER, JR 3,350,666

COAXIAL CONNECTOR

Filed April 30, 1963 4 Sheets-Sheet 2

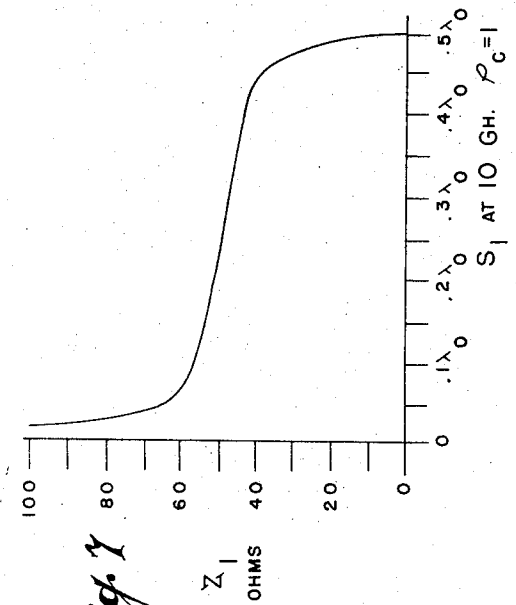


Fig. 7

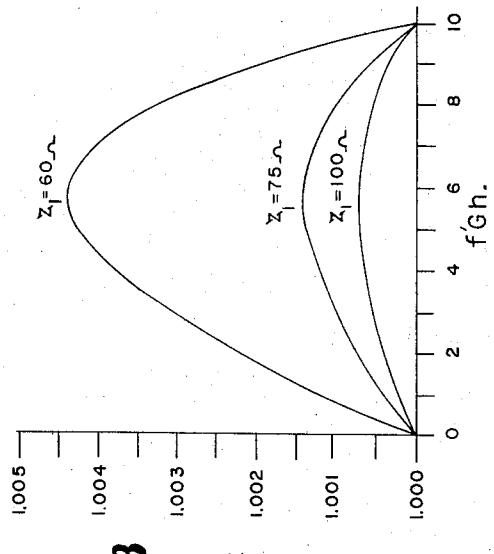


Fig. 8

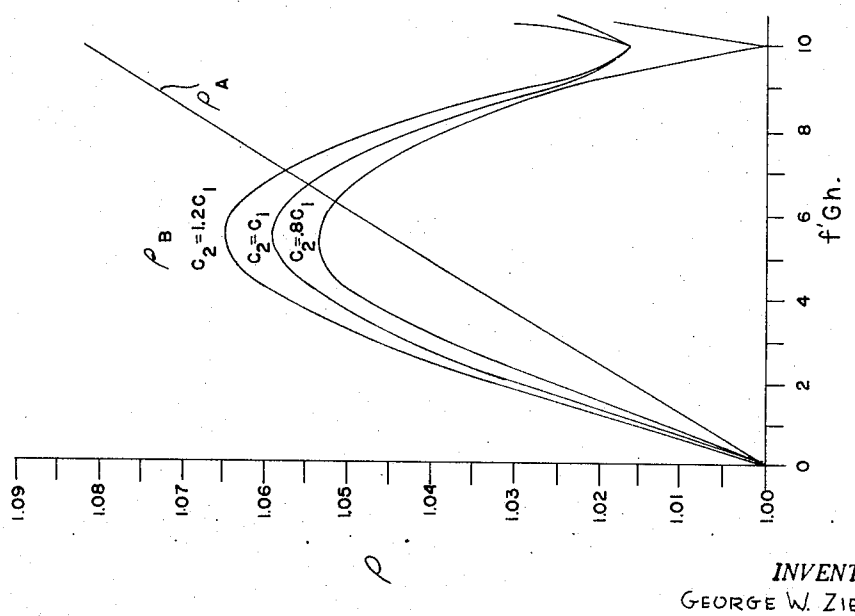


Fig. 6

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Oct. 31, 1967 G. W. ZIEGLER, JR 3,350,666

COAXIAL CONNECTOR

Filed April 30, 1963 4 Sheets-Sheet 3

Fig. 9a

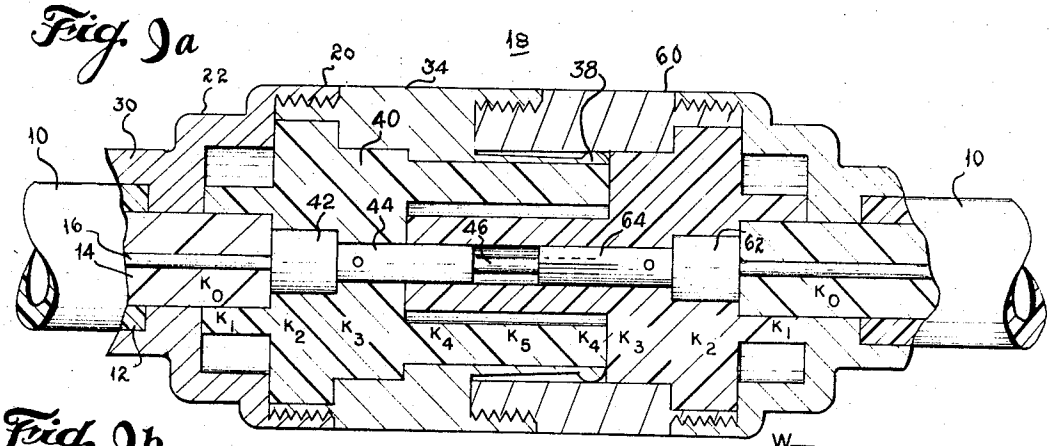


Fig. 9b

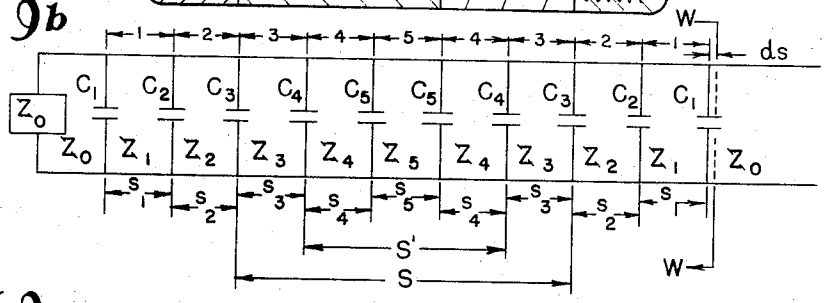


Fig. 9c

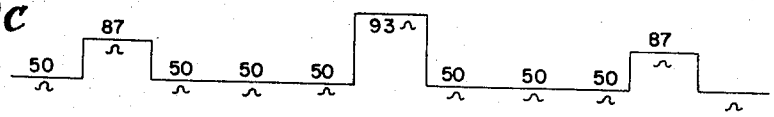
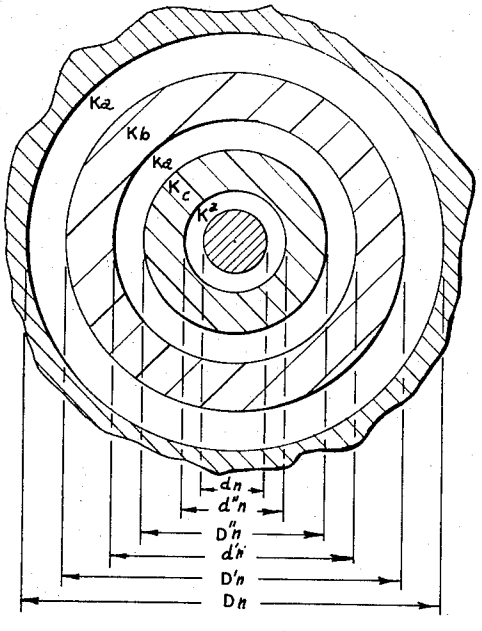


Fig. 10



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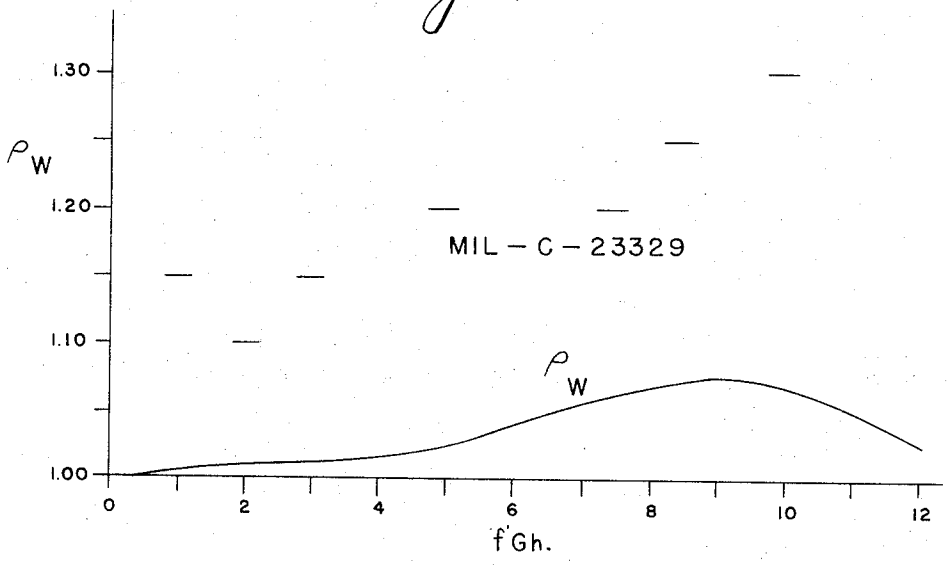
BY

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COAXIAL CONNECTOR

Filed April 30, 1963 4 Sheets-Sheet 4

Fig. 11



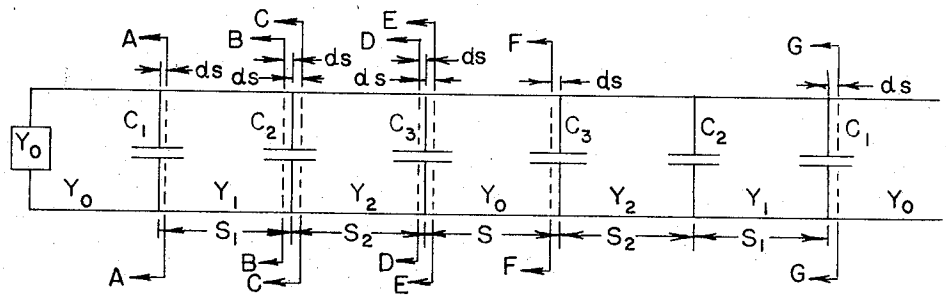


Fig. 12

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United States Patent Office 3,350,666
Patented Oct. 31, 1967

3,350,666

COAXIAL CONNECTOR

George W. Ziegler, Jr., Carlisle, Pa., assignor to AMP Incorporated, Harrisburg, Pa.

Filed Apr. 30, 1963, Ser. No. 276,714
20 Claims. (Cl. 333—33)

This invention relates to high performance coaxial connectors and to a method of compensating connectors for broad-band signal frequency use.

It is an object of the invention to provide an improved method of compensating coaxial connectors for operation with minimized signal distortion and loss over a broad range of signal frequencies.

It is a further object of the invention to provide improved coaxial connector constructions having a relatively low, flat VSWR over a broad range of signal frequencies.

The continuing and expanding development of communication equipments utilizing signal frequencies in the kilomegacycle range has produced a complementary need for high performance coaxial connectors of all types and sizes. The history of the technical development of coaxial connectors has been to adapt available connectors by modification to the extended signal frequency ranges and to the newly developed coaxial cables for such ranges. While the features of mechanical interconnection utilized in prior art devices are often satisfactory with respect to connectors utilized in the higher frequency ranges, the electrical characteristics of connectors designed for operation in use with signal frequencies ranging from several hundred to several thousand megacycles have proven to be less than optimum or even inadequate due to limitations forced on the connector design by mechanical considerations.

The present invention provides a novel method of connector design and an improved connector construction related to such method and various parts thereof. In order to fully develop the novel method of the invention and the connector construction contemplated, a theoretical treatment of problems and the contemplated solution will first be given, followed by an exemplary connector construction utilizing aspects of the method of the invention.

In the drawings:

FIGURE 1 is a schematic diagram of a generalized lossless connector including examples of mismatch sources;

FIGURE 2 is a schematic diagram depicting characteristic impedance mismatches;

FIGURE 3 is a plot of standing wave ratios resulting from impedance mismatches over the frequency range 0–12; gHz.

FIGURE 4 is a schematic diagram representing compensation of a discontinuity capacitance;

FIGURE 5 is a schematic diagram representing compensation with discontinuity capacitance pairs;

FIGURE 6 is a plot of standing wave ratios identified with different capacitance relationships for the capacitance pairs of FIGURE 5;

FIGURE 7 is a plot showing the effect of impedance on length of matching sections as related to an optimum standing wave ratio at a frequency of 10; gHz.

FIGURE 8 is a plot showing the effect of impedance on standing wave ratio where the length of a given section is optimized in accordance with FIGURE 7;

FIGURE 9*a* is a schematic diagram of a connector including intermating halves and typical diameter changes necessary to achieve mechanical interconnection;

FIGURE 9*b* is an equivalent circuit diagram for the connector of FIGURE 9;

FIGURE 9*c* is a representation of impedance levels in different sections of an exemplary connector embodiment made in accordance with the invention and having the configuration of the connector of FIGURE 9;

FIGURE 10 is a connector cross-section showing radial laminations of various materials included to facilitate manipulation of the method of the invention;

FIGURE 11 is a plot of standing wave ratio versus frequency showing the performance of the connector example included in the specification as compared with a typical performance curve for the prior art; and

FIGURE 12 is a schematic diagram of a partly compensated connector.

The objective of the coaxial connector is to provide a transmission path between coaxial cable paths, which is at least as good with respect to signal distortion and energy loss as that of the particular cable utilized. Ideally, a given connector should be physically identical to an incremental length of the particular cable to satisfy this requirement. Unfortunately, because of the requirement that cable halves be joined by a connection at least as strong, with respect to pull test, as the cable itself, connector design must deviate from the suggested ideal with various means of interlocking cable halves required. The more usual approach is to utilize threaded or bayonet type connector assemblies adapted to be crimped or otherwise affixed to each half of cables to be joined with some additional structure, including a threaded or bayonet type portion having intermating male and female parts to complete paths for the inner and outer conductors of the cable. The practical result of this requirement is that numerous changes in diameter and in conductive path spacing must be made along the length of the connector. An example of this is shown in U.S. Patent No. 2,540,012, granted Jan. 30, 1951, to Dr. O. M. Salati.

It will be realized that, for each change in diameter of conductive paths or dielectric material laminations, there may occur a change in the characteristic impedance of the connector representing a characteristic impedance mismatch.

Abrupt changes in conductor diameters represent discontinuities which behave as capacitances shunting the conductive paths at each point in a connector at the point at which diameter changes occur. Both factors act as defects which, in varying degrees, produce reflections resulting in interferences between incident and reflected waves carrying a standing wave pattern. This phenomena can seriously reduce the efficiency of energy flow.

In FIGURE 1 a broad generalization of any connector is schematically represented to include any type of discontinuity or mismatch source. The length of the connector represented as S is shown to be made up of a series of incremental or finite lengths S_1, S_2, S_3, . . . S_n. Length S exists between a transmission path such as a coaxial cable having a characteristic impedance Z_0 and a further transmission path which could be any suitable coaxial connector interface including a further coaxial cable half having the characteristics impedance Z_0. Along the length S of the connector are capacitances C_1, C_2, C_3 . . . C_n representing discontinuity capacitances occurring at the possible points of conductor diameter changes. Between each of these capacitances are impedance sections Z_1, Z_2, Z_3, . . . Z_n representing the characteristic impedances of each of the sections **1**, **2**, **3**, . . . *n*. The section lines A—A and B—B represent possible points looking into the connector shown in FIGURE 1 from lengths *ds* and S_n, respectively.

In general, the maximum standing wave ratio, SWR, resulting from *m* individual sources, of any nature, is the product of the individual contributions of each source. This is known as the combination principle as discussed

by J. W. E. Griemsmann, in the publication, "Handbook of Design and Performance of Cable Connectors for Micro-Wave Use," Bu-Ships, Index No. N.E.-110718, May 1956. A number of small energy reflection sources may, through their phase relationship, produce a large SWR. This is expressed as:

$$\rho_{max.}=\prod_{i=1}^{m}\rho_i \qquad \text{Eq. 1}$$

For an impedance match at B—B, FIGURE 1, any Z_1, and C_1 or any combination of elements may be considered to complete the match; and any is the equivalent of others. This is demonstrated as follows. Consider the length of the transmission line, S_n, of the connector of FIGURE 1, to be the matching device. With respect to admittance, Y, conductance, G, and susceptance, B, for the admittance at point A—A in FIGURE 1, the general relationship is $Y_A=G_A+jB_A$.

In terms of S_n and the phase constant β_n the foregoing may be expressed as:

$$Y_A=G_A+jB_A=\frac{Y_n(Y_0-jY_n\tan\beta_nS_n)}{Y_n-jY_0\tan\beta_nS_n} \qquad \text{Eq. 2}$$

The real parts provide the conditions for a conductance match.

$$\tan\beta_nS_n=\frac{Y_n(Y_0-G_A)}{B_AY_0} \qquad \text{Eq. 3}$$

For a susceptance match, the imaginary parts provide the required condition.

$$\tan\beta_nS_n=\frac{B_AY_n}{G_AY_0-Y_n^2} \qquad \text{Eq. 4}$$

For an admittance match

$$B_A=[(1-G_AZ_0)(G_AY_0-Y_n^2)]^{1/2} \qquad \text{Eq. 5}$$

In the above relationships the susceptance, B_A, may assume any positive or negative value. The conductance, however, must be limited to positive values, $G_A\geq 0$.

In general, G_A and B_A are different functions of frequently expressed as:

$$B_A=h_1(f) \qquad \text{Eq. 6}$$

$$G_A=h_2(f) \qquad \text{Eq. 7}$$

The match is frequently sensitive except for singular points, as indicated by the relationship obtained from Eqs. 5, 6, and 7

$$B_A=\frac{G_Ah_1(f)}{h_2(f)}[(1-G_AZ_0)(G_AY_0-Y_n^2]^{1/2} \qquad \text{Eq. 8}$$

Because of this, and in general, a perfectly flat line, SWR=1, as viewed from point B—B over a broad-band of frequency, cannot be obtained by compensating mismatches. In other words, by following the above relationships, compensation to substantially $\rho_B=1$ may be obtained at few, if any, of the frequencies present in a broad-band frequency spectrum. However, within the rigorous restriction of Eq. 5 a practical approach to a flat line may be made. The method of the invention contemplates achieving a connector having a relatively flat standing wave ratio over a broad range of frequencies by the use of novel arrangements of compensating sections and spacing therebetween. This approach contrasts materially from that of the prior art, as for example, in the Salati patent above mentioned.

In FIGURE 2 a schematic diagram of a connector is shown having a length S and an impedance Z to interconnect transmission line paths having characteristic impedances Z_0. The standing wave ratio, ρ_A as viewed at A—A may be expressed as:

$$\rho_A=\frac{1+|\Gamma_A|}{1-|\Gamma_A|}=\frac{Z}{Z_0},\ Z>Z_0 \qquad \text{Eq. 10}$$

where Γ_A represents the reflection coefficient at A—A. The characteristic impedance Z being greater than Z_0 represents a mismatch when viewed from B—B such that

$$Z_B=\frac{Z(Z_0\cot\beta_S+jZ)}{Z\cot\beta_S+jZ_0} \qquad \text{Eq. 11}$$

FIGURE 3 shows ρ_A and ρ_B in a plot of ρ versus f' wherein f' is in gHz.; $Z_0=50$ ohms; $Z=52$ ohms; $S=0.1f'\lambda$. As will be apparent from the curves in FIGURE 3, when

$$S=a\frac{\lambda}{4},\ a=1, 3, 5, \ldots \qquad \text{Eq. 12}$$

then

$$\rho_B=(Z/Z_0)^2=\rho_A^2,\ Z>Z_0 \qquad \text{Eq. 13}$$

This is in conformity with Eq. 1 above. The variation of ρ_B with frequpency is limited to $1\leq\rho_B\leq\rho_A^2$. As will be apparent from FIGURE 3, the envelope, ρ_A^2, is the locus of points satisfying Eq. 13 for all values of S. The contribution of a Z-type mismatch may be relatively large at low frequencies, especially in an electrically long connector wherein a multicplicity of maxima may occur over a given frequency range. An electrically short connector is desirable for the fewest maxima and careful design with respect to the particular Z_0 of the cable must be made to provide minimum values of ρ_{max}.

Considering now discontinuity capacitances, FIGURE 4 depicts a schematic diagram with a connector of length S having characteristic admittance Y_1 placed between transmission paths having characteristic admittances Y_0 with C representing a particular discontinuity capacitance. Considering $\omega=2\pi f$ and β as the phase constant for the section S, the conditions for an admittance match at B—B may be expressed as:

$$Y_A=Y_0+j\omega C=\frac{Y_1(Y_0-jY_1\tan\beta_S)}{Y_1-jY_0\tan\beta_S} \qquad \text{Eq. 14}$$

For a susceptance match, the imaginary parts of Eq. 14 give

$$\tan\beta_S=\frac{\omega CZ_0}{\left[\frac{Z_1}{Z_0}-\frac{Z_0}{Z_1}\right]} \qquad \text{Eq. 15}$$

From this, when S is small so that

$$\tan\beta S=\beta S \text{ radians} \qquad \text{Eq. 16}$$

then S may be expressed as follows:

$$S=\frac{\omega CZ_0}{\beta\left[\frac{Z_1}{Z_0}-\frac{Z_0}{Z_1}\right]}=\frac{c_0CZ_0}{\left[\frac{Z_1}{Z_0}-\frac{Z_0}{Z_1}\right]\sqrt{K_1}} \qquad \text{Eq. 17}$$

where $c_0=2.99776\times10^{10}$ centimeters per second and K_1 represents the effective dielectric constant of connector section length S. This is known as the Griemsmann relationship. For a conductance match, the real parts of Equation 14 yield

$$\omega CY_0\tan\beta S=0 \qquad \text{Eq. 18}$$

Unfortunately, this condition can be realized only in trite cases and the beneficial effect of obtaining a susceptance match from Eq. 15 above, can be largely offset by the conductance mismatch introduced by the compensating section of length S.

The solution to this is shown in FIGURE 5, which is similar to the schematic diagram of FIGURE 4, but includes an additional capacitance C_2 spaced from a discontinuity capacitance C_1 by a distance S_1. The two capacitances C_1 and C_2 can be selected to interact to produce an admittance match at C—C by making $Y_c=Y_0$, such that the relationship is

$$Y_c=Y_0=\frac{Y_1[Y_0+j(\omega C_1+Y_1\tan B_1S_1)]}{(Y_1-C_1\tan\beta_1S_1)+jY_0\tan\beta_1S_1}+j\omega C_2 \qquad \text{Eq. 19}$$

For a conductance match, the real parts of Eq. 19 give

$$C_1=C_2=C \qquad \text{Eq. 20}$$

For a susceptance match, the imaginary parts of Eq. 19 give

$$\tan \beta_1 S_1=\frac{\omega(C_1+C_2)Y_1}{Y_0^2-Y_1^2+\omega^2 C_1 C_2} \qquad \text{Eq. 21}$$

For an admittance match, combining Eqs. 20 and 21

$$\tan \beta_1 S_1=\frac{2\omega C Y_1}{Y_0^2-Y_1^2+(\omega C)^2} \qquad \text{Eq. 22}$$

When S_1 is small, applying Eq. 16 to Eq. 22

$$S_1=\frac{\omega(2C)Z_0}{\beta_1\left[\frac{Z_1}{Z_0}-\frac{Z_0}{Z_1}+Z_0 Z_1(\omega C)^2\right]}$$

In applying Eq. 17 above, to the connector arrangement of FIGURE 5, it is the practice when C_1 and C_2 are not equal to merely add their values. Refer to the Griemsmann article, above mentioned and to the work by M. Pomerantz, "Modification of Existing R.F. Coaxial Connectors," TM–1896, September 1957, Fort Monmouth, N.J. This is a satisfactory approximation when C_1 and C_2 are not substantially different. It is important to note that the form $(\omega C)^2$ in Eq. 22 does not appear in the Griemsmann relationship above expressed. The term $(\omega C)^2$ can be much larger than $(Y_0^2-Y_1^2)$, and thereby the controlling term, when: firstly, Y_0 and Y_1 do not differ greatly; secondly, when C is large; thirdly, when $\omega=2\pi f$ is large owing to high frequency; and fourthly, by a combination of the first three effects. From FIGURE 5 the absolute quantity of the reflection coefficient may be expressed such that

$$|\Gamma|=\left[1+\left(\frac{2Y_0}{\omega C_1}\right)^2\right]^{-\frac{1}{2}} \qquad \text{Eq. 24}$$

The interaction between C_1 and C_2 is shown in FIGURE 6 wherein ρ is plotted against f' in Gh.; $C_1=0.025\mu\mu f$; $Z_1=Z_0=50$ ohms; and S adjusted for admittance match at 10 Gh. As can be seen, the standing wave ratio, ρ_A, increases monotonically with frequency in contrast with the single Z-type discontinuity as heretofore indicated in FIGURE 3; compare ρ_A, FIGURE 3, with ρ_A, FIGURE 6. The interaction of two equal capacitances with increasing frequency produces a series of minima as required by Eq. 22, in the standing wave ratio as viewed from C—C alternating with maxima of increasing magnitude. By applying Eq. 22 at a high frequency, as for example, 10 gHz. as indicated in FIGURE 6, a single maximum at an intermediate frequency may be obtained. The magnitude of the standing wave ratio maximum decreases with a decrease of total capacitance C_1+C_2. From Eq. 20 and from FIGURE 6, the provision of equal capacitances of the smallest possible value provide the lowest possible minima with respect to ρ_C. In practice, a twenty percent difference in the capacitances is acceptable for non-critical applications. For special high-performance applications wherein the cost of tolerance control is acceptable, efforts should be made to make the capacitances as close to equal as possible.

It is to be realized that avoidance of defects causing energy loss is the best practice, but with the practicalities of manufacturing a disconnect type coaxial connection, the best approach is to provide as few discontinuities as possible with the individual discontinuities being as small as possible. It has been found that the closer the diameter of the center conductor of a connector is to that of the cable center conductor, the smaller the discontinuity capacitances of both inner and outer connector conductors. As can be seen from FIGURE 3 and its associated discussion, it is highly desirable to provide a Z_0 match over almost the entire length of the connector. This necessitates an outer conductor diameter discontinuity for every major center conductor diameter discontinuity.

Considering $Z_0=50$ ohms, a plot may be made depicting Z_1 versus connector section length S_1 at 10 Gh. for a

standing wave ratio $\rho_C=1$ with $C_2=C_1=0.025\mu\mu f$; as viewed from C—C in FIGURE 5. S_1 is expressed in terms of λ_0, the wave length in free space in centimeters. As can be seen from FIGURE 7, very short sections S_1 are required when Z_1 is made to be much greater than Z_0. Due to the slope of the curve it will be apparent that for $Z_1>100$ ohms the compensation will be sensitive to normal production tolerances for compensator length S_1. In FIGURE 8 a plot is shown of the standing wave ratio ρ_C versus f' for different values of Z_1 with S_1 adjusted according to FIGURE 7. It will be apparent that by increasing Z_1, the value of ρ_C may be made to approach 1.000.

The preceding considerations delineate the general requirements for broad-band compensation. From the curves and equations above given, it should be apparent that the total discontinuity capacitance of the connector should be minimized and most of the total should be made to reside in four equivalent or nearly equivalent discontinuities. Two of these discontinuities representing a pair associated with one connector half are matched with a high Z spacer of a length computed at a high frequency giving a small mismatch maximum at a lower frequency to achieve the improved ρ as indicated in FIGURE 8. The other pair of discontinuities are similarly matched. The pairs of discontinuity capacitances are matched against each other through being separated by a section having an effective characteristic impedance equal to that of the line Z_0, with the section length computed to compensate at the lower frequency. These lengths are calculated from Eq. 21, above given, by substituting the high or low frequency values in accordance with $\omega=2\pi f$, where f is the particular frequency. The remaining, small discontinuity capacitances (those omitted in obtaining FIGURE 12 from FIGURE 9*b*) located between the end pairs of discontinuities are matched out in pairs separated by sections of lengths computed at frequencies intermediate to the high and low frequencies when such capacitances are substantial. The foregoing may be more fully brought to mind by following an exemplary procedure in accordance with the method of the invention.

The example hereinafter given includes additional discontinuities to those which will be present in certain simplified connector designs in order to illustrate the various aspects of the method of the invention. In certain instances, depending upon the physical characteristics of the cable to be interconnected, and the mechanical specification for the connector, little compensation will be necessary, whereas in other situations the various steps of the method of the invention will be mandatory to achieve superior electrical characteristics.

Consider now that a connector is to be designed for a coaxial **10**, as shown in FIGURE 9*a*, having an outer conductor **12**, a dielectric core **14** and a center conductor **16** coaxially disposed within **12**. A typical cable of this type has a $Z_0=50\pm2$ ohms with a center conductor OD dimension of 0.0359 inch and an outer conductor dimension equal to 0.1254 inch. The dielectric material has a constant $K^*=2.25$.

The connector **18** shown in FIGURE 9*a*, is comprised of halves **20** and **60**, adapted to intermate through complementary male and female parts, The connector half **20** is comprised of a stepped metallic shell **22** adapted to be threaded onto a forward metallic shell **34**. Shell **22** includes a shell extension **30**, which may be crimped or otherwise secured to cable **10** in a standard fashion. Shells **22** and **34** include a number of internal diameter changes of the type used to secure the internal parts. The forward internal end of shell **34** includes typical spring finger members **38** which serve to complete the outer path electrical connection between half **20** and half **60**. Included in each of the halves **20** and **60** is a dielectric insert, such as **40**, shown with respect to half **20**. Half **20** further includes a central contact member **42**, which is suitably attached to the center conductor **16** of cable **10** and which includes a forward step portion **44** and a contact pin **46**. Pin **46** forms

the center path electrical connection with complementary portions of the contact member **62** through fingers **64** thereof.

The particular means employed to interconnect the connector halves to the cable **10** and the center contact members to the central conductor thereof may be achieved by a wide variety of connector designs. The important consideration with respect to connector **18**, is depicted by the schematic diagram thereunder representing the various lengths S_i, discontinuity capacitances C_i and impedance values Z_i for each section $i=1,2,3, \ldots n$.

In accordance with the theoretical procedure heretofore given and the step-by-step example to follow, the connector shown in FIGUIE 9*a* was dimensioned and compensated to produce the following values where n represents the stations shown in FIGURE 9*a* and the diameters are related to FIGURE 10. FIGURE 9*b* shows roughly the impedance levels existing along the length of the connector of FIGURE 9*a*.

NOMINAL DIMENSIONS IN INCHES

n	Conductor Diameters		Dielectric Bead Diameters				Lengths S_n''
	D_n	d_n	D_n'	d_n'	D_n''	d_n''	
1	0.2613	0.0359	0.2140	0.1350	0.1254	0.0359	0.0210
2	.2943	.0900	.2923	.0910	d_2''	D_2''	.0300
3	.2613	.0800	.2593	.0810	d_3''	D_3''	.2650
4	.2542	.0800	.2522	.1940	.1840	.0810	.0970
5	.2542	.0400	.2522	.1940	.1840	.0810	.0080

ELECTRICAL PARAMETERS

n	Z_n Ω	C_n $\mu\mu f$	p_n ($B_n=p_n f'$ mhos)	q_n ($S_n=q\, f'\, \lambda\, n$)	K_n
1	86.72	0.0218	0.0001370	0.00244	1.372
2	50.00	.0302	.0001898	.00361	1.421
3	50.00	.0052	.0000327	.03187	1.419
4	50.00	.0006	.0000038	.01140	1.386
5	92.76	.0020	.0000126	.00081	1.195

A connector so arranged produced a standing wave ratio, ρ_W from W—W, FIGURE 9*a*, 0.2 gHz. intervals over a frequency range, f', of 0–12 gHz., as shown in FIGURE 11. The ρ_W computed is well below 1.10 at all frequencies throughout the chosen range. Also shown in FIGURE 11 is a specification for a type-N connector.

As a helpful adjunct to the practice of the method of the invention, there exists a procedure which simplifies the calculation of the effective dielectric constant through any section of a connector or cable of the type shown in FIGURE 9*a*. This is shown in FIGURE 10, wherein various radial laminations of dielectric material are repsented by K_a, K_b, K_c, and the diameters of the nth section are D_n, D_n', d_n', D_n'', d_n'', d_n. From this the effective dielectric constant of the nth section, K_n, may be computed from the expression:

$$K_n=\frac{K_aK_bK_c \log \frac{D_n}{d_n}}{K_bK_c \log \frac{D_n}{d_n}-K_c(K_b-K_a) \log \frac{D_n'}{d_n'}-K_b(K_c-K_a) \log \frac{D_n''}{d_n''}} \quad \text{Eq. 25}$$

Using this value for K_n, the characteristic impedance of the nth section may be computed from:

$$Z_n=\frac{138.05}{\sqrt{K_n}} \log \frac{D_n}{d_n}=Y_n^{-1} \quad \text{Eq. 26}$$

Following now the procedure of the invention and viewing FIGURE 9*a*, the first step of the method of the invention is to arrange all discontinuities to facilitate compensation. From the above given theoretical treatment this means that each of the connector halves **20** and **60**

A further part of the above step of the method of the is made symmetrical with respect to discontinuities such that for any necessary change in diameter in half **20**, there is a suitable change in half **20** in accordance with Eq. 26 and identical changes in half **60**. In essence, the halves **20** and **60** should be arranged to be mirror images with respect to electrical characteristics.

Part of this step is to arrange the discontinuities of the connector to facilitate compensation. This is achieved by concentrating seventy-five percent or more of the total discontinuity capacitance of the connector in four capacitances (C_1,C_2), locating a pair at each end of a connector half. Referencing FIGURE 9*b*, it will thus be seen that the discontinuity capacitances C_1 and C_2 are, respectively, represented by a change in the diameter of the outer connector shells with respect to the center conductor **16** to provide C_1 and a change in the outer conductor connector shells with a change in the effective diameter of the center conductor at the largest portion of contact members **42** and **62**, to provide C_2.

invention is to space the discontinuity capacitance pairs (C_1,C_2) by a high Z section, Z_1.

The next step is to determine the smallest diameter of the contact member portions **44** and **64**, that will provide adequate attachment to the cable center conductor and sufficient spring force, respectively. While the contact members **42** and **62** are shown enlarged to exaggerate the relative dimensions, the shoulder diameters joining the various portions should be minimized or eliminated if possible. The diameter of the most forward portion of **42**, shown as pin **46**, should be maximized such that the diameter d_5 approaches the diameters $d_3=d_4$. The enlargement of the diameter d_5 is limited by the thickness of the spring members of **64** which must provide spring force for the inner contact between the contact members **42** and **62**.

The non-compensating sections Z_2, Z_3 and Z_4 may be made to closely approximate the characteristic impedance Z_0 of the line. Considering Z_0 to be 50 ohms and using Eqs. 25 and 26, above given, the diameters D_2, D_3, D_4, may be calculated by successive approximation to provide the proper Z values. With respect to the air gaps existing in the sections **2**, **3**, and **4**, nominal spacings may be included such as 0.0020 inch between the head and outer conductor and 0.0010 inch between the bead and center conductor with 0.0100 inch being provided at the bead overlap.

With respect to the compensating sections Z_1 and Z_5, such should be adjusted to a suitable value higher than the characteristic impedance of the cable. For many applications, $Z_n \geq 1.1Z_0$ is sufficient. For the general case

an adjustment wherein $1.1Z_0 \leqslant Z_n \leqslant 2Z_0$ is preferable. In special applications wherein cost is secondary to performance, a choice of $Z_n > 2Z_0$ may be made and suitable tight tolerances imposed.

The next step is to calculate values for all discontinuity capacitances using the method of Whinnery and Jamieson; Proceedings of IRE, vol. 32, pp. 98–114 and 695–709, 1944. The provision of $C_1 = C_2$ may be made by successive approximation altering relative diameters.

Experience has shown that the section S_2 should be made as short as is possible. With respect to S_5, owing to the interaction of discontinuities of the reentrant type, the value of C_5 is dependent upon S_5. When

$$\frac{2S_5}{D_5 - d_5} > 0.25$$

C_5 decreases almost linearly with S_5. It is preferred to make S_5 as small as is permitted by the accumulation of tolerance in length of the outer conductor parts, allowing bead resilient to take up any remaining tolerances. If the diameter of **46**, d_5, has been made as large as possible relative to the portion **44** (diameter $= d_3 = d_4$), then C_5 should be a negligibly small value. If for design reasons it is not possible to so size portion **46** and the length S_5, other procedures hereinafter given may be carried out.

If C_4 is small, as it frequently may be, the length S' may be made any arbitrary value to provide bead overlap. If C_4 is appreciable, then an alternate approach may be taken to determine S'. This will be hereinafter explained relative to compensation at intermediate points of the frequency range.

In many applications it has been satisfactory to make $S = \frac{1}{2}\lambda$ at 5 gHz. Alternatively, S may be computed in the manner hereinafter given for the computation of S.

The compensation for the major discontinuity capacitances C_1 and C_2 may be made by adjusting the length S_1 in accordance with Eq. 21, above given, wherein the frequency substituted in the ω term of the equation is the highest frequency of the range for the connector; i.e., in the example above given, 10 gHz. In the event that the C_3 cannot be made negligible, a different choice of S_1 will compensate. Viewing now FIGURE 12, which includes the partinent portions of the schematic diagram of FIGURE 9*b*, it may be considered that the omitted discontinuities are compensated or otherwise reduced to a negligible effect. For a susceptance match at B—B with an admittance match at E—E, the relationship is

$$Y_c = \frac{Y_2[Y_0 - j(\omega C_3 + Y_2 \tan \beta_2 S_2)]}{(Y_2 - \omega C_3 \tan \beta_2 S_2) - jY_0 \tan \beta_2 S_2} = G_c + JB_c$$

Eq. 27

$$\beta_1 S_1 = \tan^{-1}\left[\frac{\omega(C_1 + C_2)Y_1 - B_c Y_1}{G_c Y_0 - Y_1^2 + \omega^2 C_1 C_2 + \omega C_1 B_c}\right] = 360 q_1 f' \text{ degrees}$$

Eq. 28

From this there will be a conductance mismatch when

$$Y_2 \neq Y_0;\ C_1 \neq C_2;\ \text{or}\ C_3 \neq 0.$$

It will be seen that the mismatch will be small when

$$Y_2 = Y_0 \text{ and } C_3 \text{ is} \leqslant C_2 = (1 \pm 0.2) C_1$$

In general

$$S_n = (0.084730\ S_n'' \sqrt{K_n}) f' \lambda_n = q_n f' \lambda_n \quad \text{Eq. 29}$$

$$S_1 = q_1 f' \lambda_n \quad \text{Eq. 30}$$

As above indicated, S_5 is best minimized. For compensation of the discontinuity capacitances C_5, a first procedure is to solve by successive approximation, which is necessary due to the interaction previously described. A graphical method plots C_5 vs. S_5 over the practical range of S_5 together with C_5 vs. an S_5 which is computed using Eqs. 31, 32 and 33. The point of intersection of the curves will provide the desired S_5 directly.

$$\beta_5 S_5 = \tan^{-1}\left[\frac{2\omega C_5 Y_5}{Y_0^2 - Y_5^2 + (\omega C_5)^2}\right] = 360\ q'\ 5f' \text{ degrees}$$

Eq. 31

$$S_5 = q'5^{f'} \lambda 4 \quad \text{Eq. 32}$$

$$q'5 = 0.084730 S_5'' \sqrt{K_4} \quad \text{Eq. 33}$$

If the grhaphical method for solving for C_5 and S_5 produces curves that do not intersect with values of S_5 which may be achieved in practice, the curves can be made to intersect by changing the diameter of the pin **46** (diameter d_5). It matters not which direction this diameter is changed as long as intersection is obtained and practical dimensions are achieved. Following the change in diameter, the curves may be replotted to locate a point of intersection to yield S_5.

The following method is sure to produce compensation but has the disadvantage of complicating the affected connector parts. A simultaneous decrease in Z_5 and increase in C_5 result from decreasing the diameter D_5 over the length S_5 until a suitable plot intersection is obtained as above. This procedure provides an annular ring as in the Series N connector. Application of this method to the Series N connector requires, for a match to $Z_0 = 50.00$, that the ring length be 0.0100 inch. The actual specified value is 0.026 inch, which is thus in disagreement with the instant method.

As heretofore indicated, the major discontinuities of the connector are compensated at the maximum frequency and about half the maximum frequency; in the example shown, 10 gHz. and 5 gHz. This compensation is at least effective at intermediate frequencies. The remaining small discontinuities may be spaced to compensate the small mismatch from the compensated major discontinuities at about 7.5 gHz., or spaced to compensate each other as follows. Again, from FIGURE 9*b*, assuming negligible mismatch elsewhere in the connector, S' and S_4 may be computed.

$$\beta' S' = \tan^{-1}\left[\frac{2\omega C_4 Y_4}{Y_0^2 - Y_4^2 + (\omega C_4)^2}\right] = 360 q' f' \text{ degrees}$$

Eq. 34

$$S' = q' f' \lambda_4 \quad \text{Eq. 32}$$

$$q' = 0.084730 (S')'' \sqrt{K_4} \quad \text{Eq. 36}$$

$$S_4 = \tfrac{1}{2}(S' - S_5) = q_4^{f'} \lambda_4 \quad \text{Eq. 37}$$

As will be apparent from FIGURES 6 and 8, maxima occur above 5 gHz. for the capacitance pairs involved. The exact frequency at which the maxima occur can be determined for a given ensemble of discontinuities and compensation can be made at that frequency. As a practical matter, 5 gHz. is often satisfactory. Reasonable deviation from the optimum frequency provides flexibility in adjusting the overall length of the connector.

Referring to FIGURE 12:

$$Y_c = \frac{Y_1[Y_0 + j(\omega C_1 + Y_1 \tan \beta_1 S_1)]}{(Y_1 - \omega C_1 \tan \beta_1 S_1) + jY_0 \tan \beta_1 S_1} + j\omega C_2 = G_c + jB_c$$

Eq. 38

$$Y_D = \frac{Y_2[G_c + j(B_c + Y_2 \tan \beta_2 S_2}{(Y_2 - B_c \tan \beta_2 S_2) + jG_c \tan \beta_2 S_2} \quad \text{Eq. 39}$$

$$y^E = (Y_D + jwC_3) Z_0 = g_E + jb_E \quad \text{Eq. 40}$$

$$\beta_S = \tan^{-1}\left[\frac{2b_E}{g_E^2 + b_E^2 - 1}\right] = 360\ qf' \text{ degrees}$$

Eq. 41

$$\beta_3 S_3 = \tfrac{1}{2}(\beta_S - 360\ q'f') = 360\ q_3 f' \text{ degrees} \quad \text{Eq. 42}$$

$$S_3 = q_3 f' \lambda_3 \quad \text{Eq. 43}$$

Following the arrangement of the various connector lengths, discontinuity capacitances and impedance sections, the standing wave ratio for the connector, ρ_W, may

be computed at several frequencies. This may be followed by a computation at short frequency intervals such as 0.2 gHz. The result of this is indicated in FIGURE 11.

By following the above method, connectors may be designed having compensation to provide a low, flat standing wave ratio over a broad range of frequencies.

Changes in construction will occur to those skilled in the art and various apparently different modifications and embodiments may be made without departing from the scope of the invention. The matter set forth in the foregoing description and accompanying drawings is offered by way of illustration only. The actual scope of the invention is intended to be defined in the following claims when viewed in their proper perspective against the prior art.

I claim:

1. An improved broad band compensated coaxial connector for cable of characteristic impedance Z_0, including a connector body having inner and outer conductive portions of different diameters and sections along the length of the body forming distinct discontinuity capacitances and having a given total discontinuity capacitance, the said body having a length S comprised of a number of sections of different lengths with one or more sections comprising nearly the total length S of a characteristic impedance Z_0, and with the remaining length of S of characteristic impedances each being substantialy greater than Z_0 and with the said distinct discontiuity capacitances being located at predetermined points along the length of said body.

2. The connector of claim **1**, wherein most of the total discontinuity capacitance of the connector resides in spaced pairs of discontinuity capacitances.

3. The connector of claim **1**, wherein most of the total discontinuity capacitance of the connector resides in spaced pairs of discontinuity capacitances with pair members being of substantially equal discontinuity capacitance.

4. The connector of claim **1**, wherein seventy-five percent or more of the total discontinuity capacitance of the connector resides in four substantially equal discontinuity capacitances.

5. The connector of claim **1**, wherein seventy-five percent or more of the total discontinuity capacitance of the connector resides in six substantially equal discontinuity capacitances.

6. The connector of claim **1**, wherein the connector is comprised of at least two portions, each of which includes discontinuity capacitances and section lengths having impedance values such as to be substantially mirror images.

7. The connector of claim **1**, wherein the connector includes two halves with most of the total discontinuity capacitance of the connector residing in at least three pairs of discontinuity capacitances with the pairs positioned such that one pair is proximate the physical center of the connector length and effected by the difference in diameters required to intermate the halves, and one of the other pairs positioned at the major diameter discontinuities in each half.

8. An improved transmission path for coaxial cable having a characteristic admittance Y_0 or characteristic impedance Z_0 where $Z_0=1/Y_0$ comprising in combination, an outer conductive path and an inner conductive path disposed coaxially therein, a dielectric therebetween, discontinuity capacitances C_1 and C_2 spaced by a section of length S_1, having a phase constant β_1, a characteristic admittance Y_1 or characteristic impedance Z_1 where

$$Z_1=1/Y_1$$

the length S_1 being of a value determined at a specific frequency f within the band of frequencies for which the transmission path is to be utilized such that

$$\tan B_1S_1=\frac{2\pi f(C_1+C_2)Y_1}{Y_0^2-Y_1^2+(2\pi f)^2C_1C_2}$$

9. The transmission path of claim **8**, wherein C_1 is substantially equal to C_2.

10. The transmission path of claim **9**, wherein Z_1 is greater than Z_0.

11. The transmission path of claim **9**, wherein Z_1 has a value in the range of 1.1 to 2.0 Z_0.

12. The transmission path of claim **9**, wherein Z_1 is greater than 2 Z_0.

13. The transmission path of claim **8**, wherein C_2 is not equal to C_1.

14. The transmission path of claim **13**, wherein Z_1 is greater than Z_0.

15. The transmission path of claim **13**, wherein Z_1 has a value in the range of 1.1 to 2.0 Z_0.

16. The transmission path of claim **13**, wherein Z_1 is greater than 2 Z_0.

17. An improved broadband compensated coaxial connector for a cable of characteristic value Y_0 including a connector body having inner and outer conductive portions of different diameters and sections along the length of the body forming distinct pairs of discontinuity capacitances of values C′ and C″ spaced apart by a section of length S of characteristic value Y and phase constant β determined at a frequency f through the relationship

$$\tan \beta S=\frac{2\pi f(C'+C'')Y}{Y_0^2-Y^2+(2\pi f)^2C'C''}$$

where the said distinct pairs of discontinuity capacitances with given values of Y and β existing between pair members are spaced by a length S determined at a high frequency f and the lumped discontinuity capacitance of each distinct pair is compensated against the lumped discontinuity capacitance of another pair with given values of Y and β existing between the pairs spaced by a length S determined at a lower frequency f.

18. The connector of claim **17** wherein the said distinct pairs of discontinuity capacitances have given values of Y and β existing between pair members spaced by a length S determined at a high frequency f substantially equal to the design maximum frequency for which the connector is to be used and wherein the lumped discontinuity capacitance of each distinct pair is compensated against the lumped discontinuity capacitance of another pair with given values of Y and β existing between the pairs spaced by a length S determined at a frequency f substantially equal to half the said design maximum frequency.

19. The connector of claim **17** including other pairs of discontinuity capacitances located between said distinct pairs or adjacent thereto with an ensemble comprised of a said other pair and a distinct pair compensated by a length S at values Y and β at the high frequency f and with the resulting ensemble compensated by a length S at values Y and β at a lower frequency f to give overall broadband compensation matching to the cable utilized.

20. In a broadband compensated coaxial connector for use with cable of characteristic value Y_0, a connector body having inner and outer conductor portions of different diameters in sections along the length of the body forming distinct discontinuity capacitances of values C′ and C″ spaced apart by a section of length S of characteristic value Y and phase constant β determined at a frequency f from the relationship.

$$\tan \beta S=\frac{2\pi f(C'+C'')Y}{Y_0^2-Y^2+(2\pi f)^2C'C''}$$

where the said pairs of discontinuity capacitances have values of Y and β therebetween spaced by a length S determined at a high frequency f, a further pair of discontinuity capacitances with further values of Y and β existing between said further pair spaced by a further length S determined at same or different high frequency f, the said distinct pairs and said further pair being compensated by spacing the lumped discontinuity capacitance of each distinct pair with respect to a member of said further pair by a length S of characteristic value Y and

phase constant β determined at a lower frequency f so as to provide overall broadband compensation of the connector to the cable.

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