FIG. 1

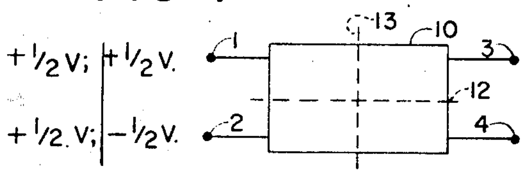


FIG. 2A

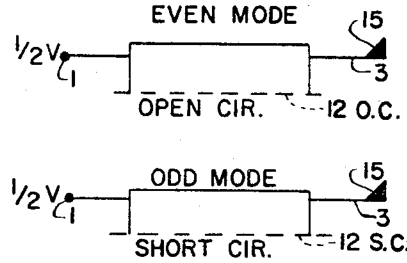
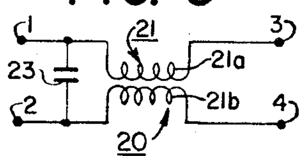


FIG. 2B

FIG. 3



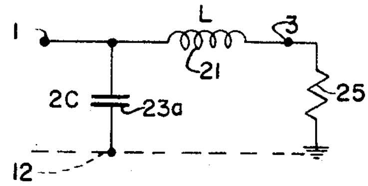


FIG. 4

FIG. 5A

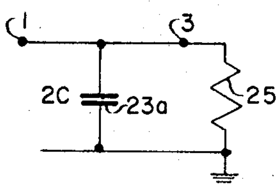


FIG. 5B

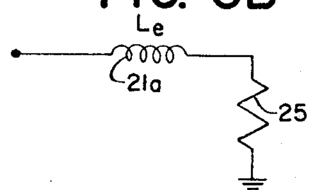


FIG. 6

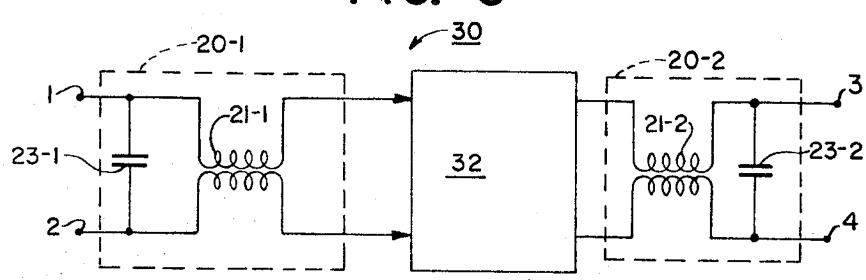


FIG. 10

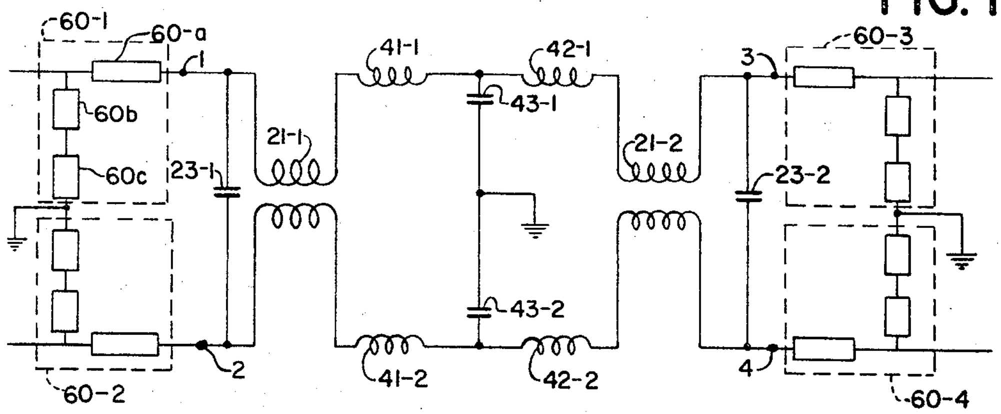


FIG. 7

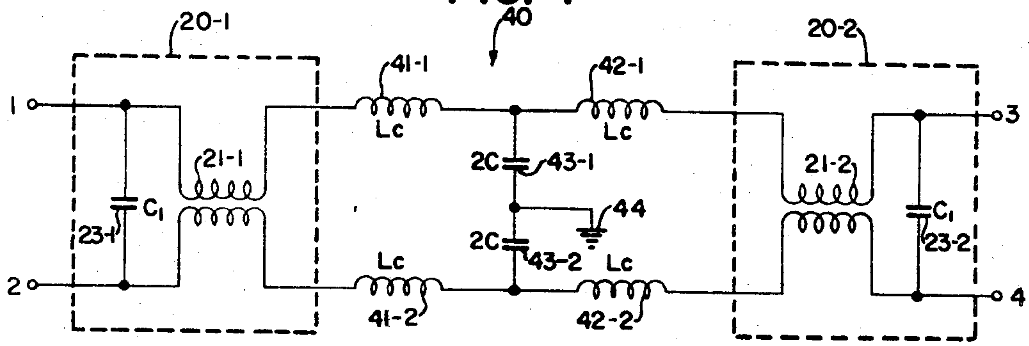
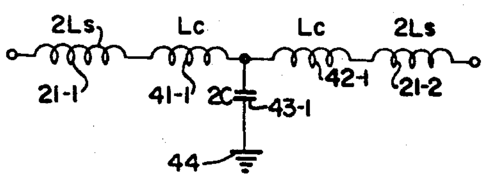
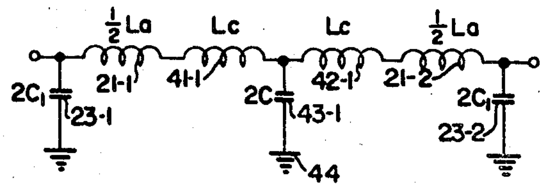


FIG. 8A



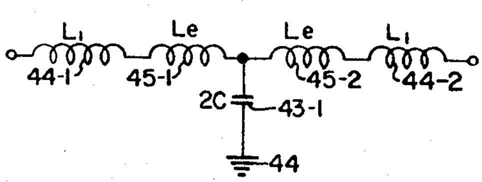
EVEN MODE BISECTION

FIG. 8B



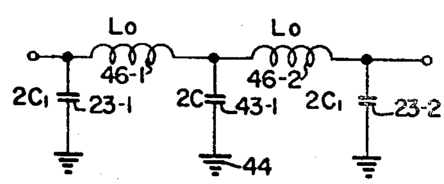
ODD MODE BISECTION

FIG. 9A



EVEN MODE EQUIVALENT

FIG. 9B



ODD MODE EQUIVALENT

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United States Patent Office

3,452,300
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FOUR PORT DIRECTIVE COUPLER HAVING ELECTRICAL SYMMETRY WITH RESPECT TO BOTH AXES

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Int. Cl. H01p *5/14;* **H03h** *1/00*
U.S. Cl. 333—10 **12 Claims**

ABSTRACT OF THE DISCLOSURE

A symmetrical four port directive coupler for controlled operation over a wide band of frequencies in which two networks formed by lumped constant parameters are connected together by an electrical coupling means. The parameters of the two networks and the electrical coupling means are selected so that the normalized input impedance of the even mode bisection of the entire symmetrical directive coupler is equal to the normalized input admittance of its odd mode bisection.

This invention relates to devices for coupling radio frequency energy and more particularly to couplers constructed in accordance with imposed conditions of duality.

An object is to provide coupling devices formed by symmetrical networks using imposed conditions of duality.

An additional object is to provide coupling devices formed by symmetrical networks and having desired coupling properties which are constructed using imposed conditions of duality in which the normalized input impedance of the even mode equivalent circuit bisection is equal to the normalized admittance of the odd mode equivalent circuit bisection.

Another object is to provide devices for coupling radio frequency energy using lumped circuit components in which the values of the components at a particular frequency of operation are selected in accordance with imposed condition of duality.

Yet a further object is to provide radio frequency energy coupling devices having a plurality of obstacles using lumped constant components for both the elements of the coupler obstacles and for connecting the plurality of obstacles together.

An additional object is to provide radio frequency coupling devices having a plurality of obstacles and using lumped constant components for both the elements of the coupler obstacles and for connecting the obstacles together, such elements being selected in accordance with imposed conditions of duality, and means for correcting for dispersion of the devices.

In accordance with the present invention, radio frequency energy coupling networks are provided which are relatively simple to construct and have desired isolation, input match, coupled output, energy transmission and frequency responsive characteristics. These couplers are constructed as symmetric networks from lumped constant components whose values are selected in accordance with conditions of duality imposed upon the network. In the preferred embodiment of the invention the duality condition is that the input impedance of the even mode bisection of the network equals the input admittance of the odd mode bisection.

Other objects and advantages of the present invention will become more apparent upon reference to the following specification and annexed drawings in which: FIG. 1 shows a four terminal symmetric coupler network; FIGS. 2A and 2B, show the even and odd mode bisections of the network of FIG. 1; FIG. 3 is a schematic diagram

of a four terminal symmetric coupler network formed of lumped constant components; FIG. 4 is a schematic diagram of the bisection of the network of FIG. 3; FIGS. 5A and 5B respectively are the odd mode and even mode equivalent circuits for the bisection of FIG. 4; FIG. 6 is a schematic diagram of a two obstacle coupler network; FIG. 7 is a schematic diagram of a two obstacle coupler using lumped constant components; FIGS. 8A and 8B respectively show the even and odd mode bisections of the coupler of FIG. 7; FIGS. 9A and 9B respectively show the equivalent transmission line type circuits for the bisections of FIGS. 8A and 8B; and FIG. 10 is a schematic diagram of a coupler of the type shown in FIG. 7 with additional elements provided for correcting dispersion.

FIGURE 1 shows in general block notation a symmetric four terminal network **10** of the type to be considered having components (not shown) which are frequency responsive to change their impedances and admittances. The network has four ports **1**, **2**, **3** and **4**, and is of the type such that when an input signal is applied to port **1**, a coupled output signal is produced at port **2**, a transmitted output signal produced at port **3** in phase quadrature with the coupled output at port **2**, and port **4** is isolated so that no output signal appears thereon. Many such networks are well known in the art.

Since network **10** is symmetrical, in accordance with theory of analysis it is said to have a plane of symmetry **12** so that a voltage V applied between input ports **1** and **2**, with the other ports **3** and **4** terminated with the proper impedances, can be broken up into the two equivalent modes of excitation applied between ports **1** and **2**. These two modes are, as shown in FIG. 1, an even mode excitation of +½V and +½V applied to the respective ports **1** and **2** and an odd mode excitation of +½V and −½V applied to the same two ports. The even and odd mode analysis is shown for example in an article by S. B. Cohn entitled "Shielded Coupled-Strip Transmission Line" in the October 1955 Transactions IRE, Volume PGMTT.

Bisecting the symmetrical network **10** along its plane of symmetry **12** produces the two bisected even mode and odd mode networks shown respectively in FIGS. 2A and 2B. Reference numeral **15** indicates the characteristic impedance termination for port 3. Using the even and odd mode excitations of FIG. 1 for FIGS. 2A and 2B respectively, it can be seen that in FIG. 2A plane **12** is effectively an open circuit plane since excitations of +½V and +½V applied to terminals **1** and **2** produce no difference of potential therebetween and therefore no current flow between the two input ports **1** and **2**. When the odd mode excitation of +½V and −½V is applied to terminals **1** and **2**, plane **12** is considered to be short circuited since a difference of potential V exists between ports **1** and **2**. This is illustrated by the odd mode network bisection network of FIG. 2B.

Letting Z_{in_e} be the normalized input impedance for the even mode bisection of FIG. 2A, where the normalized input impedance equals the input impedance of the network at any one frequency divided by the characteristic input impedance of the network, it can be shown by network analysis that:

(1)
$$S_{11_e}=1/2V\left[\frac{S_{in_e}-1}{Z_{in_e}+1}\right]=1/2V\Gamma_e$$

and

(2)
$$S_{13_e}=1/2V[1-\Gamma_e^2]^{1/2}$$

where S_{11} is the reflected signal at port **1** and S_{13} is the signal transmitted between ports **1** and **3**. Γ_e is the reflection coefficient of the even mode bisection at port **1** of the S_{11_e} term and $[1-\Gamma_e^2]^{1/2}$ is the transfer coefficient between ports **1** and **3** of the S_{13_e} term. The even mode bisection gives the same results for ports **2** and **4**, so that:

$$S_{22_e}=\tfrac{1}{2}V\Gamma_e \quad (3)$$

and

$$S_{24_e}=\tfrac{1}{2}V[1-\Gamma_e^2]^{1/2} \quad (4)$$

where S_{22_e} is the reflected signal at port **2** and S_{24_e} is the signal transmitted between ports **2** and **4**. S_{11}, S_{13}, S_{22} and S_{24} are commonly called the scattering coefficients of the network.

If, in FIG. 2B, for odd mode bisection Y_{in_o} is the normalized input admittance, where the normalized input admittance equals the input admittance at a specific frequency multiplied by the characteristic input admittance of the network, then it can be shown by network analysis that

$$S_{11_o}=1/2V\left[\frac{1-Y_{in_o}}{1+Y_{in_o}}\right]=1/2V\Gamma_o \quad (5)$$

$$S_{13_o}=\tfrac{1}{2}V[1-\Gamma_o^2]^{1/2} \quad (6)$$

where S_{11_o} and S_{13_o} correspond to Equations 1 and 2 for the even mode bisection of FIG. 2A. Since the odd mode bisection anticipates $-\tfrac{1}{2}V$ at terminal **2** of the symmetrical network, then:

$$S_{22_o}=-\tfrac{1}{2}V\Gamma_o \quad (7)$$

and

$$S_{24_o}=-\tfrac{1}{2}V[1-\Gamma_o^2]^{1/2} \quad (8)$$

By imposing an condition for duality on the network **10** such that $Z_{in_e}=Y_{in_o}$, then from Equations 1 and 5.

$$\Gamma_e- -\Gamma_o$$

meaning that the reflection coefficient of the even mode bisection is equal to the negative of the reflection coefficient of the odd mode bisection. By using (9), the scattering coefficients for the odd mode bisection in terms of Γ_e become:

$$S_{11_o}=-\tfrac{1}{2}V\Gamma_e \quad (5a)$$
$$S_{13_o}=-\tfrac{1}{2}V[1-\Gamma_e^2]^{1/2} \quad (6a)$$
$$S_{22_o}=\tfrac{1}{2}V\Gamma_e \quad (7a)$$
$$S_{24_o}=-\tfrac{1}{2}V[1-\Gamma_e^2]^{1/2} \quad (8a)$$

To obtain the scattering coefficients for the entire four terminal network **10** of FIG. 1, Equations 1 through 4 and 5*a* through 8*a* are added, giving:

$$S_{11}=S_{11e}+S_{11_o}=0 \quad (10)$$
$$S_{12}=S_{22_e}+S_{22_o}=V\Gamma_e \quad (11)$$
$$S_{13}=S_{13_e}+S_{13_o}=V[1-\Gamma_e^2]^{1/2} \quad (12)$$
$$S_{14}=S_{24_e}+S_{24_o}=0 \quad (13)$$

Equations 10 through 13 define the symmetric network **10** of FIG. 1 as a directional coupler having the following characteristics:

(a) Isolation—since $S_{14}=0$ there is no signal transmission between ports **1** and **4**.
(b) Input match—since $S_{22}=0$ there is no mismatch at the port **1** input.
(c) Coupled output of Γ_e—$S_{12}=V\Gamma_e$ defines the coupling between ports **1** and **2**,
(d) Transmission of $[1-\Gamma_e^2]^{1/2}$—$S_{13}=V[1-\Gamma_e 2]^{1/2}$ defines the transmission between ports **1** and **3**.

Additionally, due to the symmetry of the network about a second bisecting plane **13**, which is transverse to plane **12**, the coupled output at port **2** can be shown to be in phase quadrature with the transmitted output at port **3**.

All of the above desired characteristics are produced in the couplers of the present invention using imposed duality conditions with lumped circuit elements, such as capacitors and inductances. By suitable connection and selection of the values of the elements, couplers of a relatively simple form are produced having desired coupling properties over a range of frequencies.

The principles of the present invention are discussed with respect to the coupler **20** using lumped constant elements shown in FIG. 3, which is called a single obstacle type coupler. Coupler **20** is a symmetrical network formed by a lumped inductor **21** whose two coils respectively have their ends connected to ports **1** and **3** and ports **2** and **4**. Ports **1** and **2** are shunted by a lumped capacitor **23** of capacitance C.

The wires of the two coils forming inductance **21** are preferably twisted together and both wires are wound on a coil form or a toroidal core to form a bifilar winding.

In any bifilar inductor, such as inductor **21**, the odd, or anti-symmetric, mode inductance is less than the even, or symmetric, mode inductance. This is so because in the odd mode most of the electromagnetic field is contained between the wires while the even, or symmetric mode inductance remains large. A bifilar inductor can be wound to have a considerable difference between the even mode and odd mode inductances.

The odd mode inductance L_o of a bifilar inductor is measured as a series connection of the pair of wires, each having an inductance L_a, forming the inductor. Since the wires each of inductance L_a are connected in series, the total odd mode inductance L_o equals $\tfrac{1}{2}L_a$. The even mode inductance L_e is measured as a shunt connection of two wires each having an inductance L_s so that $L_e=2L_s$.

Since network **20** of FIG. 3 is symmetrical, it can be bisected along its plane of symmetry **12** as in FIGS. 2A and 2B to give FIG. 4. The capacitance of capacitor **23*a*** is 2C since half of the total capacitance C of capacitor **23** was taken and capacitors in series are added by adding their reciprocals. Resistor **25** designates the terminating impedance of the network, and is shown as having a value of one (1) unit. The value L is shown for the inductance of the one wire of bifilar inductor **21** assumed to be in the half of the network encompassed by the bisecting plane **12**.

The odd mode equivalent circuit for the bisected network of FIG. 4 is shown in FIG. 5A. As explained with respect to FIG. 2B, bisecting plane **12** is considered to be a short circuit plane in the odd mode meaning that the capacitor **23*a*** of value 2C is in parallel with the output impedance **25**. The inductance of the inductor **21** is neglected since it is very small in the odd mode, as explained above. FIG. 5B shows the even mode equivalent circuit for the bisected circuit of FIG. 4. Since bisecting plane **12** is an open circuit plane for the even mode, the capacitor **23*a*** has no effect and the inductance of the coil **21** is the high value even mode inductance L_e.

For the odd mode equivalent circuit of FIG. 5A the normalized input admittance is:

$$Y_{in_o}=1+jZ_o w(2C)=1+jb \quad (14)$$

Z_o designates the characteristic impedance of the circuit which is multiplied with the imaginary part of (14) rather than divided since admittance is the reciprocal of impedance. When terminated by its characteristic impedance Z_o the network **20** has an admittance looking into the circuit of $1/Z_o$.

Using Equation 5 the reflection coefficient Γ_o of the circuit of FIG. 5A is:

$$\Gamma_o=\frac{-jb}{2+jb} \quad (15)$$

and the scattering coefficient S_{11_o} is:

$$S_{11_o}=1/2\left[\frac{-jb}{2+jb}\right] \quad (16)$$

Similarly, for the even mode equivalent circuit of FIG. 5B, the normalized input impedance is:

$$(17) \qquad Z_{in_e}=1+\frac{jwL_e}{Z_o}$$

Here, the imaginary part of (17) is divided by Z_o to obtain the normalized impedance.

Since $L_e=2L_s$ in the even mode, then:

$$(17a) \qquad Z_{in_e}=1+j\frac{w2L_s}{Z_o}=1+jx$$

Using Equation 1, the reflection coefficient Γ_e of the circuit of FIG. 5B is:

$$(18) \qquad \Gamma_e=\frac{jx}{2+jx}$$

and the scattering coefficient S_{11_e} is:

$$(19) \qquad S_{11_e}=1/2\left[\frac{jx}{2+jx}\right]$$

Imposing the duality condition on the circuit of FIG. 3 that its normalized input impedance for the even mode bisection equals its normalized input admittance for the even mode bisection gives from Equations 14 and 17:

$$(20) \qquad Z_o w(2C)=\frac{w(2L_s)}{Z_o}$$

or

$$(21) \qquad \frac{L_s}{C}=Z_o^2$$

and, from Equations 14 and 17*a*

$$(21a) \qquad x=b$$

Using Equations 1 through 8 and 5*a* through 8*a* to obtain the even and odd mode scattering coefficients for the network of FIGS. 3–5 in terms of the even mode coefficient and adding the even and odd coefficients in accordance with Equations 10 through 13 to obtain the scattering coefficients for the entire symmetrical four-port network gives:

$$(22) \qquad S_{11}=0$$

$$(23) \qquad S_{12}=\frac{jx}{2+jx}$$

$$(24) \qquad S_{13}=\left[1-\left(\frac{jx}{2+jx}\right)^2\right]^{1/2}$$

$$(25) \qquad S_{14}=0$$

The coupling k in db between ports **1** and **2** of the coupler of FIG. 3 is given as:

$$(26) \qquad k=10\log\left|\frac{1}{S_{12}}\right|^2=10\log\frac{4+x^2}{x^2}=10\log\left|1+\frac{4}{x^2}\right|$$

The transmission loss L in db between ports **1** and **3** is given by:

$$(27) \qquad L=10\log\left|\frac{1}{S_{13}}\right|^2=10\log\left(\frac{4+x^2}{4}\right)=10\log\left(1+\frac{x^2}{4}\right)$$

This completely defines the coupler. Given the coupling value k necessary, Equation 26 is used to solve for x which, from Equation 21*a*, is equal to b. For any predetermined frequency of operation and input impedance level the value of the inductance **21** and capacitor **25** can be determined.

For example, consider that a 3 db coupler is to be constructed. Using 3 db as the value for k in (26) and solving for X gives $X=b=2$. Since, from (20)

$$(28) \qquad X=\frac{w2L_s}{Z_o}=2$$

and

$$b=Z_o w2C=2$$

then

$$(29) \qquad L_s=\frac{Z_o}{w}$$

and

$$C=\frac{1}{Z_o w}$$

For an operating center frequency f of 30 mc. ($w=2\pi f$) and a characteristic impedance Z_o of 50 ohms, $L_s=.2653$ μh and $C=106$ μμf. L_s gives the value of each coil of the bifilar inductor **21**.

While the single obstacle coupler described above is quite useful and has many applications, it has a relatively narrow bandwith, like any other frequency responsive circuit with only a few components. To provide greater bandwidth, a two obstacle coupler is used. This coupler **30** is shown in FIG. 6. Here, two symmetrical single obstacle networks **20–1** and **20–2**, of the same construction as network **20** of FIG. 3, are connected together by a coupling element **32**. Network **20–1** has a capacitor **23–1** (C_1) shunted across the ports **1** and **2** and a bifilar inductor **21–1** connected in the same manner as network **20** of FIG. 3. Network **20–2** is the same as **20–1** and has a bifilar inductor **21–2** and a capacitor **23–2**, the latter of which is connected in shunt across output ports **3** and **4** and to the ends of the two wires of inductor **21–2.**

The ideal elements **32** for coupling networks **20–1** and **20–2** together are, of course, transmission lines of the proper length and characteristics for matching the characteristic impedances of the two networks, so that the imposed duality condition is not destroyed. A pair of such lines would be used, one connecting each of the corresponding coils of the two bifilar inductors **21–1** and **21–2.** If the impedances of element **32** are non-dispersive, that is, have substantially linear phase shift and have no attenuation over the operating frequency range, then designing the two obstacle coupler **30** is a straight forward aplication of the design principles for the single obstacle coupler described above. At some frequencies, particularly higher frequencies, sort lengths of transmission lines possess substantially non-despersive properties, and are an ideal choice for elements **32.** Such short line lengths would not adversely affect operation of the coupler **30** and the duality design criterion of $Y_{in_o}=Z_{in_e}$ can be achieved.

At intermediate frequencies (I.F.), such as in the range from 1 to 100 mc., the lengths of the two transmission line coupling elements **32** between networks **20–1** and **20–2** needed to preserve the duality condition would be fairly long and would make the coupler considerably bulky. To keep the coupler **30** compact, lumped constant components are preferably used for the coupling elements **32.** At these "lower" I.F. frequencies the transmission lines have dispersive properties. Therefore, the coupler must also have the same dispersive properties at a particular frequency of operation as the equivalent lengths of transmission lines, in order to maintain the imposed duality conditions to achieve the desired characteristic for the coupler. Such a network is illustrated in FIG. 7.

FIG. 7 illustrates a symmetrical two obstacle coupler **40** with a coupling network formed by lumped constant components in the form of inductors and capacitors. The same reference numerals used previously are used here again for the same components. The lumped constant component equivalent for each connecting transmission line of FIG. 6 is shown as a T-circuit having two inductors **41** and **42** and a shunt capacitor **43.** The upper T-circuit has series connected inductors **41–1, 42–1** and a shunt capacitor **43–1,** whose lower end is connected to a point of reference potential such as ground **44,** connecting the ends of the upper wires of bifilar inductors **21–1** and **21–2.** Inductances **41–1** and **42–1** are single coil inductors. A similar T section formed by single coil series connected inductors 41–2, 42–2 and shunt capacitor 43–2 connects the ends of the lower wires of bifilar inductors 21–1 and 21–2. Capacitors 23–1 and 23–2 each have the same value C_1, while capacitors 43–1 and 43–2 each have a value 2C. The single coil inductors 41–1, 41–2, 42–1 and 42–2 each have a value L_c.

FIGS. 8A and 8B respectively show the even and odd mode bisections of the coupler 40 of FIG. 7. As in FIGS. 3 and 4, the even mode inductance L_e of each bifilar inductor 21–1 and 21–2 is 2 L_s. The odd mode inductance L_p for each of 21–1 and 21–2 is $L_o=\frac{1}{2}L_a$. The inductances of coupling coils 41 and 42 in each mode is L_c while the capacitance of capacitor 43 is 2C. Since the even mode bisection plane 12 is an open circuit plane, capacitors 23–1 and 23–2 do not appear in the even mode bisection of FIG. 8A. In the odd mode bisection of FIG. 8B, where plane 12 is a short circuit plane, capacitors 23–1 and 23–2 have a value of 2C. It should again be pointed out that the even and odd mode bisections are circuits with identical properties, although the circuits are shown in somewhat different form, since network 40 is symmetrical.

FIGS. 9A and 9B respectively show the even and odd mode equivalent circuits of FIGS. 8A and 8B. Referring first to FIG. 9A, for purposes of analysis, the even mode equivalent circuit has been redefined as an equivalent transmission line T-circuit with a pair of series connected inductors 44–1 and 45–1 of values L_1 and L_e, respectively, shunt capacitor 43–1 of value 2C connected to ground, and a second pair of series connected inductors 45–2 and 44–2 of values L_e and L_1 connected to the junction of inductor 41–1 and capacitor 43–1. Here, the two inductors 45–1 and 45–2 of value L_e give the even mode inductance of the equivalent of T-section transmission line. The value L_e also includes the inductance of the bifilar transformers 21, which is not negligible in the case where two networks 20 are connected together. The two inductors 44–1 and 44–2 of value of L_1 are the coupling obstacles (inductances) of the equivalent transmission line. From FIGS 8A and 9A it is clear that:

(30) $$2L_s+L_c=L_1+L_e$$

The odd mode bisection of FIG. 8B is also shown in FIG. 9B for purposes of analysis as an equivalent transmission line T section. This section is formed by series connected coils 41–1 and 46–2 of value L_o connected at their junction by capacitor 43–1 of value 2C to ground. The unconnected ends of coils 46 are shunted to ground by the capcitors 23–1 and 23–2 of value $2C_1$. Here, capacitors 23–1 and 23–2 are the coupling obstacles of the equivalent line T sections and:

(31) $$\tfrac{1}{2}L_a+L_c=L_o$$

The even and odd mode bisections of FIGS. 8A and 8B as well as their corresponding equivalent circuits of FIGS. 9A and 9B are uncoupled to each other, due to the symmetry of original network 40 and the nature of the bisecting planes 12. These bisected circuits of FIGS. 8A and 8B are also identical, so that the even and odd mode T-section transmission line equivalents of FIGS. 9A and 9B are also identical so that:

(32) $$L_e=L_o$$

Here, coils 41 and capacitors 23 (L_1 and C_1) are to be the coupling obstacles which are to be dualized with respect to each other and with respect to the transmission line T-section equivalent circuits of FIGS. 9A and 9B to impose the condition of $Z_{\text{in}_e}=Y_{\text{in}_o}$. Since the transmission lines are uncoupled from the coupling obstacles L_1 and C_1 and the entire network has a characteristic impedance Z_o, which is the same as the terminating impedance 25 of FIG. 4, Z_{in_e} of the even mode equivalent circuit of FIG. 9A is $w(L_1)/Z_o$ and Y_{in_o} of the odd mode equivalent circuit is $Z_o w(2C_1)$.

In FIGS. 8 and 9 L_1 is specified by subtracting (31*a*) and (31*b*) to give:

(33) $$L_1=2L_s-\frac{L_a}{2}$$

and L_c is determined from (31*b*) as:

(34) $$L_c=L_o-\frac{L_a}{2}$$

As indicated above, the equivalent transmission line coupling elements between the two networks 20–1 and 20–2 are to have dispersive properties. To continue with the analysis presented in terms of an equivalent transmission line, such as the T-sections of elements 43 and 45 in FIG. 9A and 43 and 46 in FIG. 9B, the characteristic impedance of such lines can be defined as a quantity Z_T such that

(35) $$Z_T=Z_o\cos\theta$$

where

$$Z_o=\sqrt{\frac{L}{C}},\ \omega_c=\frac{1}{\sqrt{LC}}$$

and

$$\sin 2\theta=\frac{\omega}{\omega_c}$$

where Z_p is the characteristic impedance of the system at zero frequency and θ is the length of the line in electrical degrees. Using Z_T as the characteristic impedance, then the coupling obstacles 44–1 and 44–2, of value L_1 and 23–1 and 23–2, of value $2C_1$ must be dual to each other and dual to the T-sections to make $Y_{\text{in}_o}=Z_{\text{in}_e}$.

Since the even and odd mode equivalent circuit transmission line sections are T-networks of the constant k type, the inductor 44 (L_1) can be made to have the same dispersion as these sections, by defining this inductor's dispersion as:

(36) $$X_{L_1}=K\tan\theta/2,$$

then

(37) $$L_1=\frac{Z_T}{\omega}K\tan\theta/2,$$

where K is a constant and θ is the line length in electrical degrees.

For the even mode equivalent circuit of FIG. 9B it can be shown that its loss function L is:

(38) $$L=1+\left(\frac{\Delta}{2}\right)^2$$

where

(39) $$\Delta=2X\cos\theta-X^2\sin\theta$$

and

(40) $$X=K\tan\theta/2.$$

The function Δ is derived from a matrix analysis of a series element X (here 44–1), a connecting transmission line section (here 45–1, 43–1 and 45–2) of dispersion

$$Z_T=Z_o\cos\theta,$$

and another series connected element X (here 44–2).

For every coupling condition desired, where the coupling is defined a

$$\left(\frac{L-1}{L}\right)$$

and L (loss function) is given in (38), different values of K and θ exist. For the specific case of

$$\frac{dL}{d\theta}=0$$

which gives the stationary point of the loss function of (38) with respect to line length θ, a plot of

$$\left(\frac{L-1}{L}\right)$$

vs. θ_o and K vs. θ_o can be made about the values of K, where θ_o is the line length in electrical degrees for the coupling value desired. From this plot K and θ are obtained for the particular coupling value.

With K and θ determined for the particular coupling values, the other parameters are determined by using the well-known relationships of a constant K type T network, which are from:

(41*a*) $$\omega=\omega_c \sin \theta/2$$

(41b) $$Z_T=\sqrt{\frac{L_e}{C}} \cos \theta/2$$

(41*c*) $$\omega=\sqrt{\frac{1}{L_eC}} \sin \theta/2$$

Here in Equations 41*a*, 41*b and* 41*c*, L_e corresponds to an inductance **45–1** or **45–2** of FIG. 9A, C corresponds to one-half the value of capacitor **43–1**, and ω_c is the cut-off frequency of the T-section network.

Using Equations 41*b* and 41*c* in (37) gives:

(42) $$K \tan \theta/2=\frac{\omega L_1}{Z_T}=\frac{L_1}{L_e} \tan \theta/2$$

from which

(43) $$L_1=KL_e=2L_s-\frac{L_a}{2}$$

from (34).

Also, from (42):

(44) $$L_e=\frac{Z_o \tan \theta/2}{\omega}$$

and therefore

(45) $$C=\frac{L_e}{Z_o^2}$$

from (41*c*).

The value of C_1 (elements **23–1** and **23–2** of FIG. 7) is derived in the same manner as L_1 since, from the imposed duality condition

(46) $$\frac{\omega(L_1)}{Z_o}=Z_o\omega(2C_1)$$

that Z_{in_e} (shown in FIG. 9A and given by the lefthand side of the equation) is equal to Y_{in_e} (shown in FIG. 9B and given by the righthand side of the equation). This is similar to the one obstacle coupler described above, with the exception of the different values of Z_{in_e} and Y_{in_o} for the equivalent circuits of FIGS. 9A and 9B. Therefore, from (46):

(47) $$2C_1=\frac{L_1}{Z_o^2}$$

which from (43) gives:

(48) $$2C_1=\frac{KL_e}{Z_o^2}=KC$$

from (45).

Thus, for any given coupling value, all of the various parameters of the network of FIG. 7 are specified. As an example, using the 3 db coupling value given above in a 50 ohm system (Z_o) operating at 30 mc., from the curves $\theta_o=21.1°$; $K=8.17$; X_o from (42) is 1.52158; $\omega_o=188.496\times10^6$ cycles; L_1 from (42) is $.40361\times10^{-6}$ henries; C_1 from (43) is 180.72×10^{-12} farads; L_e from (43) is $.0494\times10^{-6}$ henries; and 2C from (45) is 39.52×10^{-12} farads. To get L_c, if a bifilar inductor is wound to get

$$2L_s-\frac{L_a}{2}$$

equal to the value of L_1 ($.4036\times10^{-6}$ henries), L_C can be calculated such that

$$L_c=L_e-\frac{L_a}{2}$$

Any set of values of L_s and L_a can be used to fabricate L_1 so long as $L_a/2$ is equal to or less than L_e. A perfect winding is when

$$\frac{L_a}{2}=L_e$$

so that $L_c=0$ and four inductors are eliminated from the network.

The network **40** of FIG. 7 is dispersive, that is, its characteristic impedance is not constant and its phase shift is not linear with frequency, but the effects of dispersion ($Z_T=Z_o \cos \theta$) can be readily accounted for. Since θ is known for any given coupling and Z_o is known, Z_T for the elements **32** coupling the two networks **20–1** and **20–2** together is also known. The same dispersion value Z_T can be realized by connecting equivalent dispersive elements to the network **40**. Since the entire network **40** is a four terminal device with T-sections having constant *k* and frequency responsive properties, that is, it has a passband, it is effectively a constant K prototype filter. Therefore, other filters can be placed at each terminal **1–4** of network **40** to reduce the dispersion and make the network more linear. A preferred filter is the so-called *m*-derived end section (or half section) which can be connected to network **40** to correct the dispersion of the system.

FIG. 10 shows the network **40** of FIG. 6 with an *m*-derived end section **60** connected to each port. Each end section **60** is shown in the conventional manner by the three generalized impedances **60*a***, **60*b*** and **60*c*** of the *m*-derived π configuration half section. By suitably selecting the values of *m* the dispersion of the network **40** can be significantly reduced over a selected frequency range of operation. Design of *m*-derived filters is highly conventional in the art, so the details thereof are not given here.

The addition of *m*-derived sections makes the characteristic impedance of the network essentially constant at its terminals and makes its phase shift more nearly linear over the frequency band of concern. Thus, adding the *m*-derived networks at the terminals of the coupler makes the response of the coupler more nearly that of a linear system.

While preferred embodiments of the invention have been described above, it will be understood that these are illustrative only, and the invention is limited solely by the appended claims.

What is claimed is:

1. A four port directive coupler for controlled response over a broad band of frequencies, said coupler possessing electrical symmetry with respect to both axes and comprising:

first network means formed substantially of lumped constant frequency responsive components, means connecting said components so that said first network means has a pair of input ports and a pair of output ports, one of said pair of input ports serving as the input port of the coupler and the other serving as the coupler coupled output port,

second network means formed substantially of lumped constant frequency responsive components having a pair of input ports and a pair of output ports, one of said output ports serving as the output port of the coupler and the other serving as the isolated port of the coupler,

said lumped constant components of said first and second network means formed by a pair of highly magnetically coupled conductors and a capacitor in shunt with two of the ends of the conductors,

and means for electrically coupling the pair of output ports of the first network means to the pair of input ports of the second network means, the parameters of the lumped constant components of the first and second network means together with the parameters of the coupling means forming a four port symmetrical directional coupler between the input ports of the first network means and the output ports of the second network means in which the two port networks formed by the even mode bisection of the entire coupler about the symmetry axis is substantially the dual of the two port networks formed by an odd mode bisection of the entire coupler about its symmetry axis over said band of frequencies.

2. A coupling network as set forth in claim 1 wherein said electrical coupling means includes at least one transmission line.

3. A symmetrical four port coupling network as in claim 1 further comprising means connected to at least one of said first and second network means for reducing the dispersion of the symmetrical coupler.

4. A symmetrical four port coupling network as in claim 1 wherein said electrical coupling means comprises a pair of transmission line means with a line being connected between the third and first and the fourth and second ports of the first and second network means respectively.

5. A coupling network as set forth in claim 4 wherein each of said transmission line means is substantially nondispersive over a substantial portion of the operating band of the directive coupler.

6. A symmetrical four port network as set forth in claim 1 wherein the components of each of said first and second network means comprises a bifilar wound inductor having one end of each winding electrically connected to a respective port of one of the pairs of input and output ports, a lumped capacitor means shunted across said one end of each of the two windings, and means connecting the other end of each winding of the bifilar inductor to a respective port of the other one of the pairs of input and output ports.

7. A symmetrical four port network as set forth in claim 6 wherein the electrical coupling means connecting the output ports of the first network means and the input ports of the second network means comprises a respective transmission line means connecting each of a pair of ends of the bifilar inductor of one network means to each of a pair of ends of the bifilar inductor of the other network means.

8. A symmetrical four port network as set forth in claim 6 wherein the other end of each of the windings of the bifilar inductor of the first network means is connected to a respective one of the pair of output ports and the

other end of each of the windings of the bifilar inductor of the second network means is connected to a respective one of the pair of input ports of the second network means, said electrical coupling means between said first and second network means comprising a pair of series connected coils connected between an output port of the first network means and an input port of the second network means, and a capacitor connected between the junction of each pair of series connected coils and a point of reference potential.

9. A symmetrical four port network as set forth in claim 6 wherein the electrical coupling means connecting the output ports of the first network means and the input ports of the second network means comprises a lumped constant transmission means formed of a cascade of series inductors and shunt capacitors connected between an output port of the first network means and an input port of the second network means.

10. A coupling network as set forth in claim 7 in which each said transmission line means is substantially nondispersive over a substantial portion of the operating band of the directive coupler.

11. A symmetrical four port network as set forth in claim 8 further comprising means connected to a port of one of said network means for reducing the dispersion of the symmetrical coupler.

12. A coupling network as set forth in claim 11 wherein said means for reducing dispersion comprises an *m*-derived filter end section.

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U.S. Cl. X.R.

333—24, 70

UNITED STATES PATENT OFFICE
CERTIFICATE OF CORRECTION

Patent No. 3,452,300 June 24, 1969

Joseph D. Cappucci et al.

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

Column 2, equation (1) should appear as shown below:

$$S_{11_c} = 1/2V\left[\frac{Z_{in_e} - 1}{Z_{in_e} + 1}\right] = 1/2V\Gamma_e$$

Column 3, line 36, at the beginning of the line insert -- (9) --; same line 36,

$\Gamma_e - - \Gamma_o$ should read $\Gamma_e = -\Gamma_o$

same column 3, equation (6a) should appear as shown below:

$$S_{13_o} = 1/2V\left[1 - \Gamma_e^2\right]^{1/2}$$

same column 3, line 66, section (d) should appear as shown below:

$$[1-\Gamma_e^2]^{1/2} — S_{13} = V[1-\Gamma_e^2]^{1/2}$$

Column 8, line 27, "where Z_p is the" should read -- where Z_o is the --.

Signed and sealed this 29th day of June 1971.

(SEAL)
Attest:

EDWARD M.FLETCHER,JR.
Attesting Officer

WILLIAM E. SCHUYLER, J
Commissioner of Patent