United States Patent [19]

Hopfer

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[45] **May 21, 1974**

[54] **CONICAL SPIRAL CONDUCTOR FOR APPLYING LOW FREQUENCY SIGNALS TO A MICROWAVE STRUCTURE**

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[73] Assignee: **General Microwave Corporation,** Farmingdale, N.Y.

[22] Filed: **Aug. 18, 1972**

[21] Appl. No.: **281,994**

Related U.S. Application Data

[60] Division of Ser. No. 78,891, Oct. 7, 1970, Pat. No. 3,713,031, which is a continuation-in-part of Ser. No. 788,254, Dec. 31, 1968, abandoned.

[52] **U.S. Cl.** ..... **333/97 R,** 333/97 S

[51] **Int. Cl.** **H01p 1/00**

[58] **Field of Search** ..... 329/161, 162; 333/7 D, 333/84 M, 97 R, 97 S

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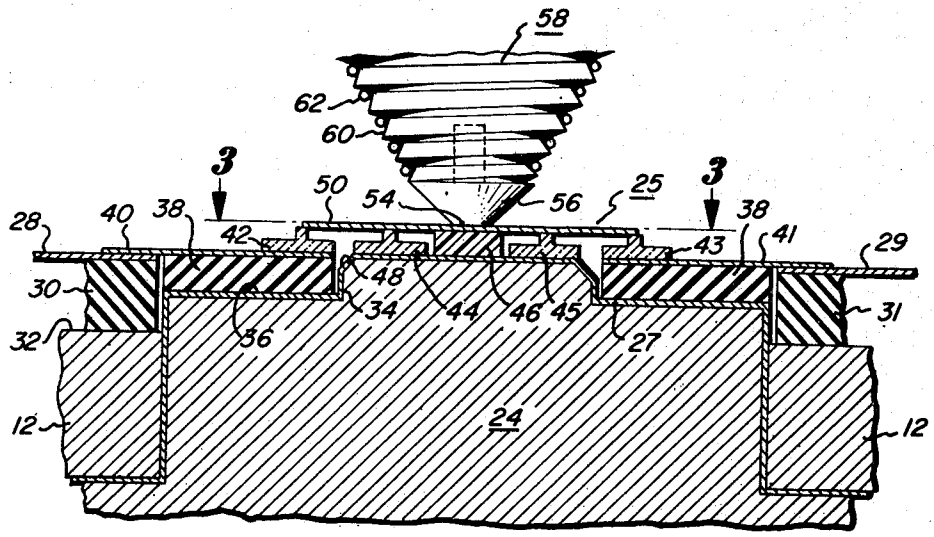
Primary Examiner—Paul L. Gensler

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[57] **ABSTRACT**

A microwave transmission system is formed by conically spiral resistive conductors that form broadband, reflectionless, non-absorbing connections. The conical spiral conductor serves as a single-wire transmission line having a high impedance to high-frequency signals over a broad band carried by an r-f transmission line and serves as an effective coupling to such an r-f line.

27 Claims, 11 Drawing Figures



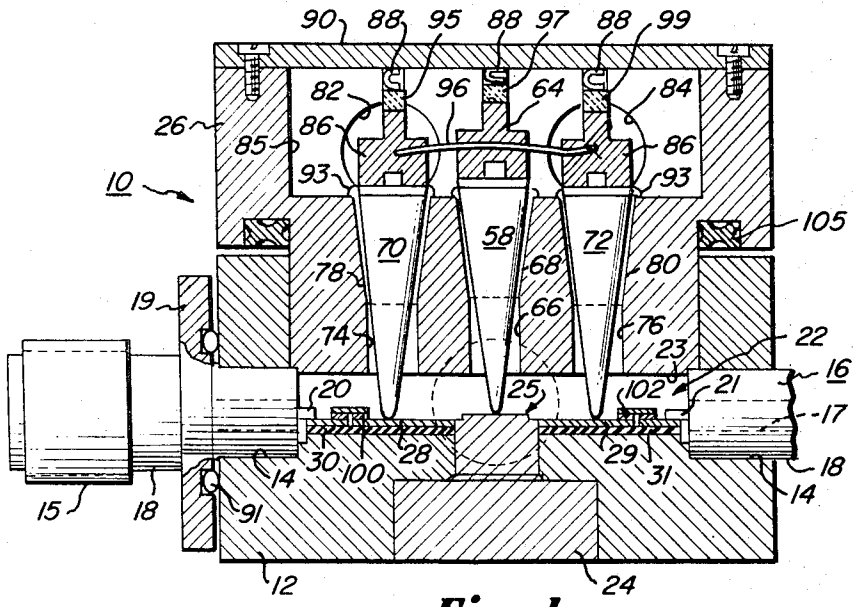

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64
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72
78
68
80
19
66
74
76
22
23
16
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FIG. 1

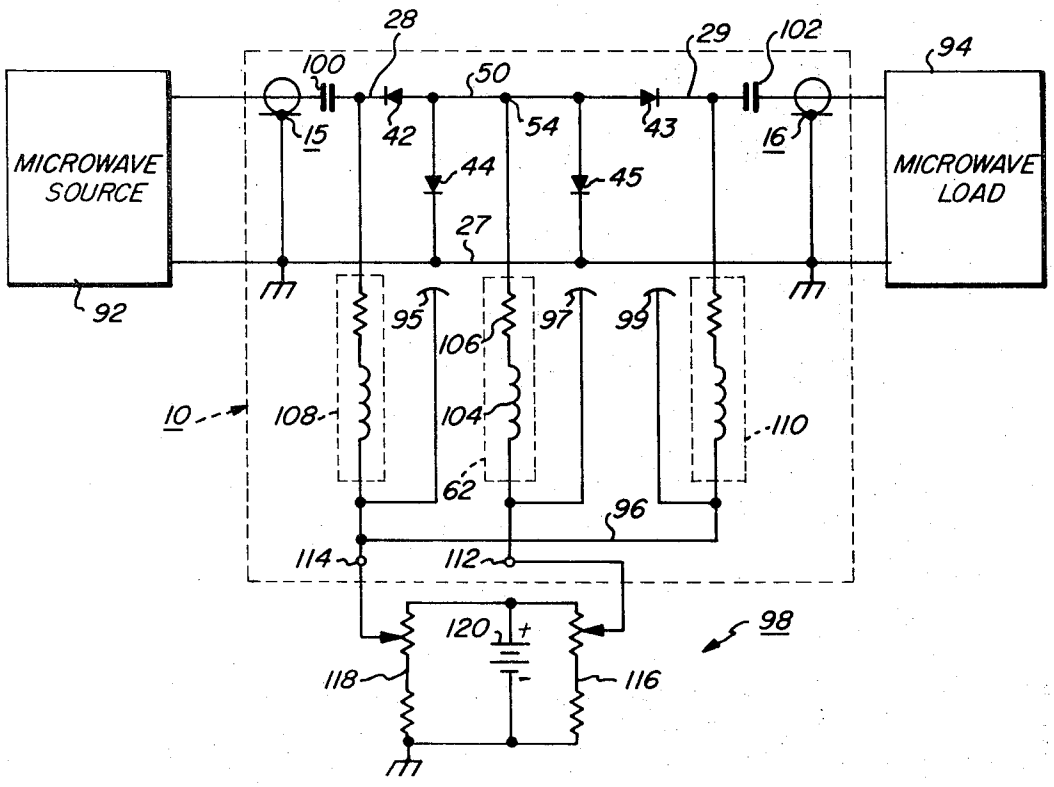
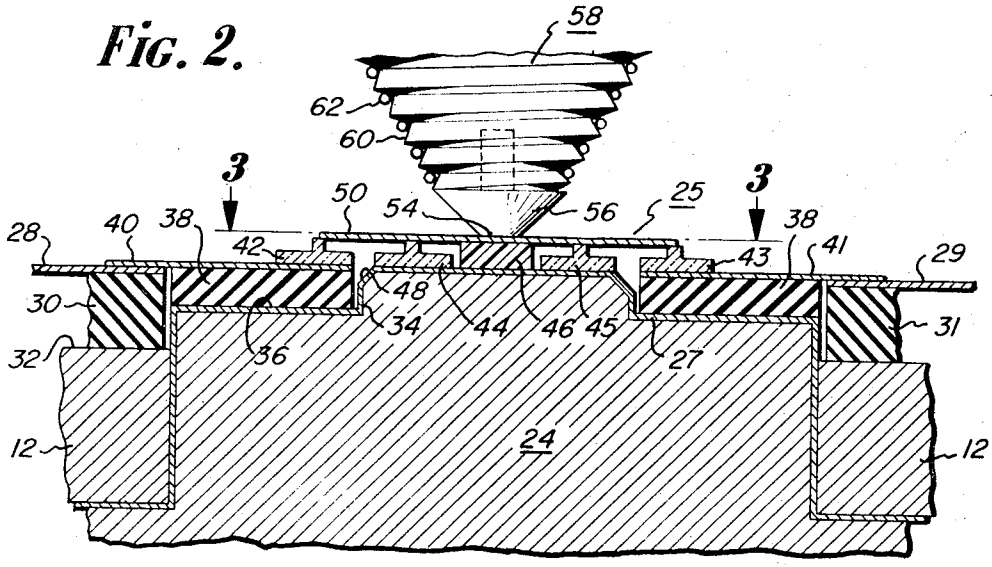

28
29
102
94
100
50
MICROWAVE SOURCE
15
42
54
43
16
MICROWAVE LOAD
44
45
27
92
95
97
99
106
10
108
104
110
62
96
114
112
120
98
118
116

FIG. 4

FIG. 2.


58
62
60
3
3
50
54
56
25
28
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38
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43
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30
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24
12

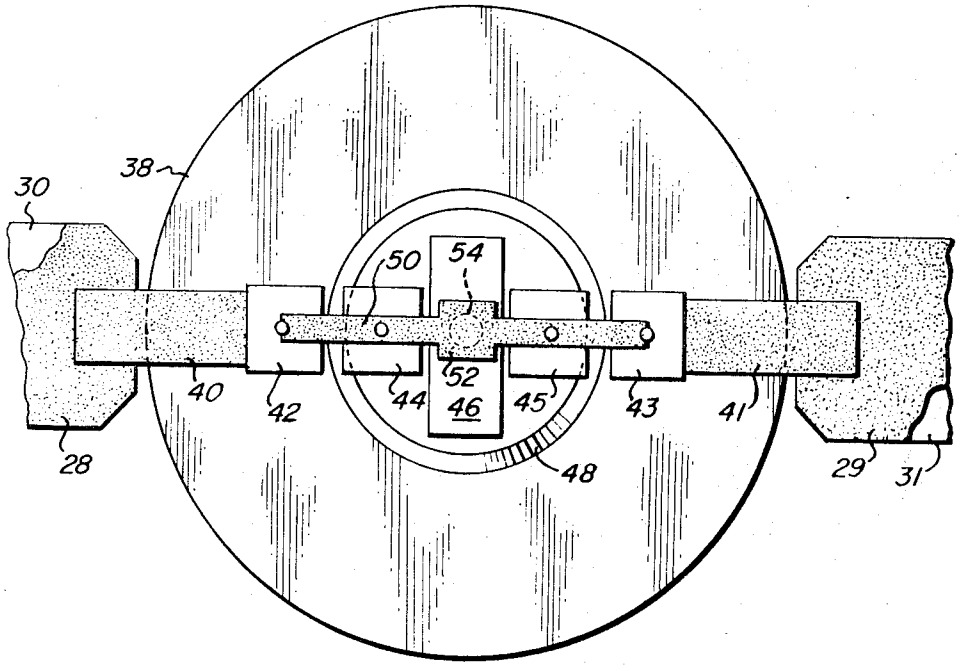

38
30
50
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52
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42
44
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43
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FIG. 3.

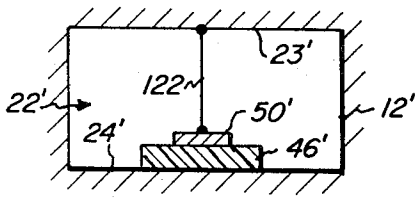


FIG. 5.

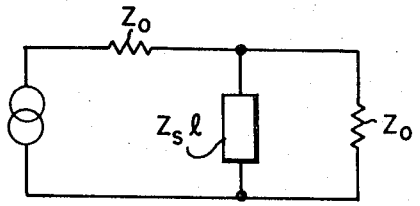


FIG. 6.

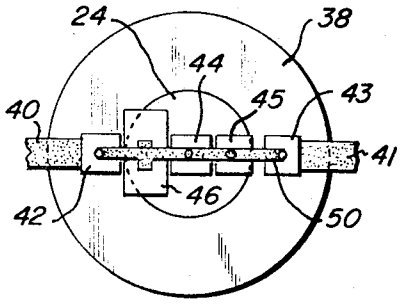


FIG. 7.

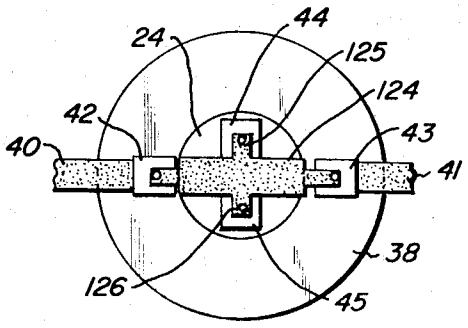


FIG. 8.

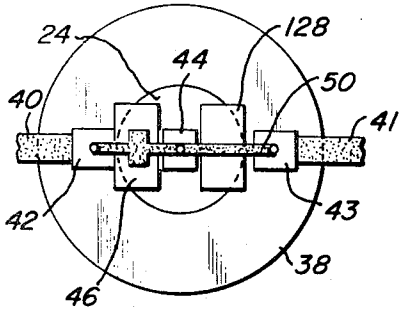


FIG. 9.

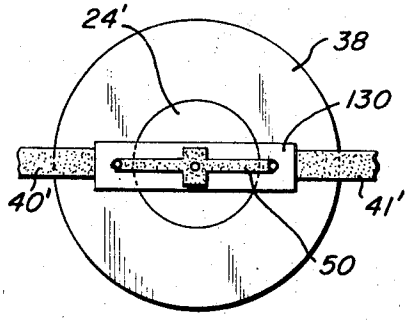


FIG. 10.

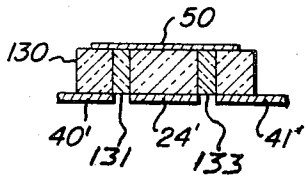


FIG. 11.

CONICAL SPIRAL CONDUCTOR FOR APPLYING LOW FREQUENCY SIGNALS TO A MICROWAVE STRUCTURE

This is a division of application Ser. No. 78,891, now U.S. Pat. No. 3,713,037 filed Oct. 7, 1970, which was a continuation-in-part of Ser. No. 788,254, filed Dec. 31, 1968, now abandoned, having the same applicant and assignee as the present application, and describing forms of attenuators and coupling devices.

BACKGROUND OF THE INVENTION

This invention relates to wideband microwave attenuators and particularly to a wideband voltage controlled microwave attenuator employing semiconductor elements such as PIN diodes, and to coupling devices usable in such attenuators.

A circuit designer synthesizing low frequency circuits may employ devices designated resistor, capacitor, inductor without questioning these designations. As a circuit's operating range is extended into the megahertz region and into gigahertz, complex impedance characteristics of these elements become significant and must be considered. When a circuit is designed in the microwave region each circuit element must be considered as a complex network. The task of constructing the circuit which can equivalently be represented by a resistance that remains substantially constant in value over a wide band of frequencies in the microwave region becomes a sophisticated problem.

Typically, d-c currents are applied to circuit elements, for example, to bias diodes through inductive leads. Usually a coil serves as an adequate inductor to pass the d-c without loading the circuit. When the circuit must be operated in the microwave region, however, a coil ceases to be an inductor but rather tends to be a parallel tuned circuit at some frequency, and yet at higher frequencies the capacity between the turns may predominate and effectively render a conventional coil a capacitor rather than an inductor.

SUMMARY OF THE INVENTION

Therefore, it is an object of this invention to provide a microwave coupling device.

Another object is to provide a coupling network for applying low frequency signals to a microwave structure.

In accordance with one form of the invention described in detail hereinafter, a conical, spiral conductor serves as a single-wire transmission line for coupling bias signals to the diodes. The single-wire line has a relatively low impedance to the low frequency bias signals and a high impedance to radio frequency signals over a broad band carried by the attenuator.

A trough-line structure in the form of a hollow waveguide encloses a TEM microstrip line having the conical spiral coupling connected thereto whereby propagation within the waveguide is restricted to the single TEM mode.

BRIEF DESCRIPTION OF THE DRAWING

The foregoing and other objects of this invention, the various features thereof, as well as the invention itself, may be more fully understood from the following description when read together with the accompanying drawing, in which:

FIG. **1** is a sectional view in elevation of an attenuator unit embodying this invention;

FIG. **2** is a greatly enlarged view of the encircled portion of FIG. **1** and is a sectional view corresponding thereto;

FIG. **3** is a top plan view taken along the line **3—3** of FIG. **2**;

FIG. **4** is an equivalent schematic circuit diagram of the attenuator unit of FIG. **1**;

FIG. **5** is a simplified sectional view of a trough-line structure similar to the structure of FIG. **1** except with a straight line coupling conductor instead of a conical helix; the section corresponding to one at right angles to FIGS. **2** and **3**;

FIG. **6** is an idealized equivalent circuit diagram used in the explanation of this invention;

FIGS. **7, 8, 9** and **10** are top views similar in orientation to that of FIG. **3**, illustrating modified forms of this invention; and

FIG. **11** is a sectional view of a portion of FIG. **10**, taken along the horizontal diameter of the circular member therein.

In the drawings, corresponding parts are referenced by similar numerals throughout.

DESCRIPTION OF A PREFERRED EMBODIMENT

In FIG. **1**, the attenuator unit **10** is constructed with a housing of electrically conductive material such as aluminum. The housing is generally rectangular and symmetrical and includes a bottom block **12** of generally rectangular outer shape and having aligned circular openings **14** in opposite faces, in which are mounted coaxial connectors **15** and **16** having an outer conductor **18** and a central conductor **17** (part of the connector **16** is omitted from FIG. **1** for simplicity of illustration). A mounting plate **19** secured to the connector **15** is used to fasten the latter to the block.

The block **12** has a generally rectangular inner opening **22** formed between the inner and outer openings **14**, and which is enclosed from the top by the lower face of a block **26** inserted within block **12**. A stud **24** of generally circular form is inserted through a similarly shaped opening in the lower wall of block **12** and is centrally and symmetrically located between the ports **15** and **16**. The stud carries on its upper face, which enters the opening **22**, a diode network **25** illustrated in greatly enlarged form in FIGS. **2** and **3**. The stud **24** serves as part of a ground plane for this network, and is preferably formed of a conductive material such as tellurium copper, which has a gold plating **27** (FIG. **2**) for good electrical connections; a nickel plating under the gold serves as a barrier against diffusion of the gold into the copper and also is a good thermal conductor to carry heat energy away from the diode network.

The portions **20** and **21** of the central conductors projecting into the region **22** serve as a transition between the coaxial waveguide of the connectors **15** and **16** at each port and the strip-line form of TEM transmission line located within the opening **22**. The strip-line has a ground plane formed by the upper surface of the aluminum block **12** which is in direct electrical contact with the outer coaxial conductor **18**, which in use is normally a reference or ground connection. The strip-line at one port includes a dielectric layer **30**, such as teflon, in the form of a rectangular strip between the ground-plane block **12** and a thin, rectangular, copper

strip 28 which overlies the dielectric and is directly connected to the central conductor 20 of one port. A similar copper strip 29 overlies a similar dielectric layer 31 at the other port thereof and is connected to the center conductor 21 of the other port's connector 16.

The center stud 24 is cut away at its upper end (FIG. 2) to form a smaller, circular throat portion 34 with a chamfered rim 48, and on the surrounding shoulder 36 a dielectric ring 38, such as beryllia, is mounted. The elements are dimensioned so that the top surface of the ring 38 is at the same level as the top surface of the copper strips 28 and 29; and radially extending over, and in direct contact with, the latter strips and the beryllia ring on each side are gold strip-line ribbons 40 and 41.

The top of the throat portion 34 of the center stud 24 is at substantially the same level as the top surface of the strip-line ribbons 40 and 41. Two outer diodes 42 and 43 respectively rest on the ribbons 40 and 41, and two inner diodes 44 and 45 rest on the upper throat portion 34 of stud 24. A dielectric block 46, such as kapton, is located between the center diodes 44 and 45 and likewise rests on the throat 34 at the center thereof. The tops (e.g., the anodes) of all of the diodes 42–45 and of the center block 46 are at substantially the same level, and secured to these electrodes is a gold strip-line ribbon 50 which is generally uniform in thickness and in transverse width, for its electrical characteristics; a central portion 52 (FIG. 3) of this ribbon is somewhat wider for its mechanical, structural function as a bearing surface, as hereinafter described, but without effect on its electrical characteristics. The chamfered outer rim 48 of the top section of the throat 34 provides a substantial electrical spacing between the inner ends of strips 40 and 41 and the top conductive surface 27 of stud 24.

Resting on the center of strip 52 is the circular tip 54 of a metallic cone 56 which functions as a contact and which is secured to a cone 58 of dielectric material (such as rexolite) as the lower tip thereof. Formed around the conical surface of cone 58 is a helical groove 60, and inserted fully therein is a resistive wire (e.g., nickel) 62, which wire at its lower end is secured to the metal contact 56, and at its upper end is attached to a generally cylindrical metal contact 64 (FIG. 1) which sits on top of the dielectric cone 58. The conical metal tip 56 is employed for the extremely small contact radius of about 0.01 inch or less; alternatively, the dielectric cone 58 may be formed with a conical tip having a metallic contact coating connected to the spiral wire 62.

The upper block 26 of the housing has a cylindrical central hole 66 through which the lower portion of cone 58 passes. The passage 66 communicates with a counterbore opening 68 which generally mates with the outer surface of the cone 58 and supports the upper portion thereof. Thus the cone 58 is retained within the parallel conical opening 68 through a good portion of its length, and the cylindrical passage 66 provides a gradual transition in spacing between the wire 62 and grounded block 26.

The other cones 70 and 72 (FIG. 1) are constructed and wound with wires similarly to cone 58. Cones 70 and 72 are located in cylindrical openings 74 and 76 on opposite sides of the central opening 66 and aligned in a straight line therewith. Similar conical counterbore openings 78 and 80 respectively connect with the upper portions of the cylindrical openings 74 and 76. The counterbore openings 78 and 80 are similar, but have somewhat greater depth than the central counterbore opening 68 so that the cones 70 and 72 extend deeper (e.g., about 0.005 inch) into the central opening 22 and the metal contact tips thereof rest on the copper strip-line portions 28 and 29, which are at a lower level than strip 50, corresponding to about the thickness of the diodes 42–45, as may be seen in FIG. 2. Two transverse openings 82 and 84 in the upper block 28 pass into an upper rectangular opening 85 in block 26 into which the upper portions of the cones extend. The openings 82, 84 provide passage for two connectors (not shown) that are connected to the wires of the cones; one such connector (having a terminal 112 as indicated in the equivalent circuit of FIG. 4) is connected to the contact 64 of the central cone, and the other connector (terminal 114) is connected to one of the similar contacts 86 on the top of the other cones 70 and 72. A bypass capacitor 95, 97, 99 is respectively connected between the upper end of the spirals on each of the cones 70, 58 and 72 and the ground plane of the aluminum housing. Ceramic capacitors are used, and each is secured on the top of the metal contacts 64 and 86 and electrically connected in position by means of metallic C-springs 88 pressed between the respective capacitor and the cover plate 90. A wire 96 interconnects the contacts 86 of the outer cones; other wires connect one of the latter contacts and the central contact to the terminals 114 and 112, respectively (FIG. 4). Each of the cones 58, 70 and 72 is fastened to and sealed in the upper block 26 by means of a flexible epoxy or silicone rubber in the form of a separate ring 93 adjacent the upper rim of each of these cones. The resilience in this adhesive 93 is such that when the block 26 is inserted in the block 12, and the lower tips of the cones engage the respective contact points on ribbon 50 and ribbons 28 and 29, the adhesive permits a slight upward movement of the cones caused by the pressure of engagement at the contact points. The dimensions are so chosen that such movement of the cones results in sufficient pressure being achieved in the mounting of the cones, and the resiliency of the adhesive 93 ensures that good electrical contact at the lower tips 56 is achieved. The center passage 22 of the housing is additionally sealed, when the unit is assembled, by means of resilient O-rings 91 at the mounting plates 19 and a quad-ring 95 between the upper and lower blocks 26 and 12.

Each PIN diode 42–45 is a semiconductor element having a double-diffused junction consisting of P and N regions separated by a thin layer of undoped or intrinsic (I) semiconductors; the intrinsic layer may be sharply defined if fabricated epitaxially, or it may be ill-defined due to diffusion. In its forward-biased state, the PIN diode, it is known, behaves as a resistive element over a wide range of the microwave spectrum. The dynamic resistance between its two electrodes is controllable by a bias current (which may be d-c or at audio or video frequencies) to change, for example, from about 10,000 ohms to 1 to 2 ohms. The dynamic resistance is substantially independent of frequency so long as the period of the signal is short relative to the lifetime of the minority carriers in the semiconductor.

There are no sharp upper or lower frequency limits in the microwave region beyond which the PIN diode ceases to function as a controllable resistive element, and they have been used to **18** gigahertz and above. PIN diodes have associated parasitic reactances which produce deviations from their resistive nature at the high frequency end of the microwave spectrum.

The chips are fabricated with a gold terminal at the top (which may be the anode or cathode and which illustratively is the anode) surface and with a thin layer of gold solder at the bottom surface, whereby good electrical connections are formed (e.g., by welding or thermocompression bonding) with the gold conductors **27, 40, 41, 50** in the strip-line construction. The square diode chips typically are about 0.015 to 0.020 inch long on a side and about 0.005 inch thick. The PIN diode chips **42–45** are illustrated (FIG. **2**) with their upper corners etched away, which construction serves to reduce the capacitance between the diode electrodes.

The schematic circuit diagram of FIG. **4** illustrates in the broken-line box **10** an equivalent circuit diagram of the attenuator unit of FIGS. **1–3,** and shows the connection of the unit **10** to an appropriate source **92** of microwave signals (e.g., in the broad band of 0.1 to 18 GHz) via the input port **15,** and to a microwave load **94** via the output port **16.** Biasing is supplied to the unit **10** by means of circuits **98.** Blocking capacitors **100** and **102** are provided in the input and output sections of the unit by any appropriate means; for example, a dielectric element (a chip of barium titanate) may be inserted in a break in the strips **28** and **29** as shown in FIG. **1** (or inserted in a break in the central conductor of the connectors **15** and **16** in a manner known in the art). For example, as indicated in FIG. **1,** it has been found suitable to form a break of about 0.003 inch in each strip **28** and **29** and on top of these strips **28** and **29** a dielectric chip **100** and **102** is set and an overlay **103, 105** of gold ribbon is used to bridge the breaks.

The equivalent circuit may be traced from the central conductor **20** of the input port **15** via capacitor **100** and the conductive strips **28** and **40** to the cathode of series diode **42,** the anode of which is in electrical contact with the ribbon strip **50,** as are the anodes of the other diodes **43–45.** The cathode of diode **43** is connected via the conductive strips **41** and **29** and capacitor **102** to central conductor **20** of port **16.** The cathodes of shunt diodes **44** and **45** are connected to the ground plane of gold plating **27** at the top surface of the stud throat **34.** The wire **62** of the central cone **58** has a direct electrical connection to the center of the gold ribbon **50,** and electrically it has a substantial inductance **104** produced by the multi-turn conical format, and a substantial resistance **106** due to the use of a resistive nickel wire, as explained below. The wires **108** and **110** (FIG. **4**) encircling the cones **70** and **72** are similarly characterized by a substantial inductance and resistance. The wires **108, 62** and **110** are coupled by the ceramic capacitors **95, 97** and **99,** respectively, which serve to bypass the r-f to ground.

Terminal connectors **112** and **114** (inserted via openings **82** and **84** in the upper block **26**) are respectively connected to potentiometers **116** and **118,** which are energized by battery **120.** Thereby, a biasing current supplied to the node **54** between the shunt diodes **44** and **45** may be varied separately from the direct current bias supplied to the cathodes of the series diodes **42** and **43.**

The operation may be summarized as follows: The variation of the bias supplied across each of the diodes effectively modifies the operating characteristic of the diode to a different portion of the nonlinear resistive characteristic thereof. With the resistances of the series and shunt diodes varied, the overall attenuation of the unit **10** is varied. For example, with an extremely large resistance in series diodes **42** and **43** and a relatively small resistance in shunt diodes **44** and **45,** the unit has an overall high attenuation. Similarly, when the resistance of the shunt diodes is extremely large, and that of the series diodes small, the overall attenuation is small. Between those two extremes of attenuation, variations in attenuation may be achieved by intermediate levels of bias current. The unit may be operated between the two extreme levels of attenuation in the manner of an on-off switch; or, alternatively, it may be operated at various intermediate levels corresponding to variations in the bias current supplied to the diodes. Further details of operation and of FIGS. **7** to **10** are set forth in copending application Ser. No. 78,891, now U. S. Pat. No. 3,713,037.

The conical spiral wires **62, 108** and **110,** as indicated above, are used to supply a d-c or other relatively lower frequency signal into the strip-line structure at the central nodal region **50** and at adjacent points spaced from each other by a fraction of an inch. The conical spiral wires are designed so as not to introduce any disturbance to the radio frequency line over an extremely broad bandwidth. For example, a quarter wavelength stub can be effective to feed in a signal in this fashion, but it is limited in that the microwave bandwidth that can be attenuated is much less than twice the frequency determined by the stub dimensions. The use of a direct metal connection to the contact points would generally be equivalent to a low impedance path and cause large reflections; for example, FIG. **5** shows a cross-section of a TEM transmission line of the microstrip type (i.e., a signal strip **50'** and ground plane **24'** with dielectric **46'** form the line and opposing ground plane **23'** is effectively infinitely remote and carries negligible current). FIG. **5** is somewhat representative in simplified form (and viewed as a transverse medial section) of the microstrip line of FIG. **2,** except that the spiral wire **62** of FIG. **2** is replaced by straight wire **122,** and an opening (e.g., cylindrical-conical opening **66, 68**) in top plane **23'** for passing the coupling wire **122** is omitted. In FIG. **5,** parts corresponding to those shown in FIGS. **1** and **2** are referenced by similar numerals with the addition of a prime (').

With the spiral, the looping around many times affords more length, and the non-periodic looping such as that on the cone **58** shown in FIG. **2** (or by an Archimedes spiral in a plane) avoids the resonances found in periodic structures, and therefore produces a broadband performance. The spiral looping develops an inductance in the nature of a coil, which is distributed inductance, and the length of the spiral is a distributed resistance. The spiral metal is very thin and the capacitance between each loop is small. There are many capacitances, but they are in series so that the net capacitance is but that of a single capacitance for the r-f signal. The input impedance for this spiral is the sum of the impedances of the individual loops, and the latter are determined by the reactive components. The r-f resistance of the spiral that is developed due to the skin effect may be quite large (e.g., 2000 to 4000 ohms),

many times that of the characteristic impedance (50 ohms) of the microstrip line and many times that of a corresponding resistance wire (such as the wire **122** in FIG. 5) of length equal to the height of space **22** in the housing.

The d-c bias connections that are formed by the spiral wires **62, 108** and **110** are effectively shunt connections to the ground line (via the bypass capacitors **95, 97, 99** that isolate the r-f signal from the d-c source **120**). It is therefore desirable that these shunt paths should not pass any large amount of r-f energy. If these shunt paths were purely reactive, there would be no absorption but large reflections could be induced, especially where the reactance was small compared to the 50-ohm impedance of the line. These large reflections would then be objectionable, and effectively would impair the characteristic impedance of the microstrip line. If such reactive shunt paths had a large impedance, they would nevertheless resonate at various frequencies, which would then produce the same low impedance effect. The ideal is for the spiral lines to have a high impedance, with some loss or absorption of r-f energy in the shunt lines so as to prevent or substantially reduce the inevitable resonances (in a bandwidth from megahertz to gigahertz where the frequency extremes may be in the ratio of 200 to 1). That is, the resistance of the spiral paths should not be so large as to lose bias power and produce excessive heat. On the other hand, the r-f resistance of the spiral should be large enough so that the suck-outs of the r-f signal (at repeated ¼ wavelengths) are small and substantially smoothed out. Thus, at a normal resonant frequency, where one might expect to have a 40 or 50 percent shunting of r-f energy, the actual absorption produced by the resistive spiral **62** is only 1 or 2 percent. That is, the ratio of the microstrip impedance of 50 ohms to the r-f resistance of 2000 (or up to 4000) ohms of the spiral corresponds to about 2 percent; which is about a tenth db.

The resistance of the spiral should preferably be from 20 to 40 times the characteristic impedance of the transmission line (about 10 times might be good enough for some purposes). Nickel has the desired resistive characteristic (a number of times greater than copper), and its overall length is chosen to be sufficiently great for the desired overall resistance of the wire. Nickel has the additional characteristic of being ferromagnetic, and therefore its r-f skin depth is still smaller than non-ferromagnetic materials. Accordingly, its r-f resistance is still higher. On the other hand, for the d-c bias currents the nickel has a reasonably small resistance.

The impedance of the coupling spiral (**62, 108** or **110**) relative to that of the microwave transmission line is always high, and (except at the lowest frequencies) the attenuation is large enough so that the character of the termination (e.g., a short) of the spiral does not affect the impedance level seen looking out from the microwave line. Thus, the use of the conical spiral or helix makes it possible to achieve, in the extremely small available space **22** in the housing, a broadband, substantially reflectionless connection for supplying the bias currents and without absorption thereof. Though the height of the cone **58** is only about 0.4 inch, about 75 turns can be wound on it at a rate of about 200 turns per inch. Each of the turns is almost parallel to the strip-line plane and thus substantially at right angles to the electric field at the strip **50.** The cone **58** carries the spiral transmission line **62** vertically away from the strip-line ground plane so that the capacitance to ground decreases as the spiral rises and increases in length, while the inductance and impedance to r-f increase. The conical configuration makes it possible to build up as many turns and as much length as may be required, and by its nature, varies the length per turn of the spiral so that it meanders aperiodically. The cylindrical passage **66** of the grounded top block **26,** through which the intermediate cone portion passes, brings the ground plane gradually toward the cone to match with it and remain parallel to it so as to gradually increase attenuation and produce a good termination of the spiral line. The characteristic impedance of the spiral line thereby is reduced, and the attenuation factor (which increases with resistance and inversely with the characteristic impedance) increases.

As explained below, the conical spirals **62, 108, 110** each can be analyzed as a transmission line made up of a single wire. In such a line the electric lines of force originate at one point on the single line and terminate at another point. There is no separate line or ground; the transmission line functions as though the ground were at infinity. For the microwave signals carried by the microstrip line at r-f frequencies, this single-wire line **62** has a high characteristic impedance, but for bias signals at d-c or at lower modulating frequencies (for example, up to 100 megacycles) the wire has a low or moderate impedance. To achieve the high Z_o at high r-f frequencies, a very thin wire (e. g., 1 mil) is provided, since the impedance (or surface resistance) varies inversely as the radius, as explained below. In addition, as also shown below, this surface impedance at the microwave-signal frequencies varies directly as the permeability at those high frequencies of the wire relative to that of free space. Thus a high magnetic permeability wire is provided whose permeability is substantially greater (for example, 10 times as great) than the free space permeability, and as high as possible (for example, 100 or more times). This magnetic permeability has the effect of additionally concentrating the microwave signal in a thin layer of the wire; that is, the skin depth of the wire is effectively made very small, and therefore the wire is very lossy at high frequencies.

The effect of an increased length of line (by reason of a conical spiral instead of the straight line in FIG. 5) is that the electrical intensity due to the microwave signal at any point along the line is decreased, which decreased intensity results in a decreased coupling between the microstrip **50** and the conical line **62,** so that the microwave current in the line **62** is small. Accordingly, by using a helix, the length of the line can be increased substantially. The spiral helix as a particular form of meandering line has an additional advantage: The component of electrical intensity along each turn of a helix is very much less than the intensity at right angles to the microstrip, due to the shallow lead angle of the helix. In practice, the pitch is made to be about 200 turns per inch, which is about 5 mils per turn (for practical purposes, the pitch is made constant since the machining of support **58** is easier with a constant pitch). The lead angle of the helix is the ratio of the pitch to the radius of the turn, and accordingly, as the diameter of the helical coil becomes greater, the lead angle becomes smaller. In practice, the lead angle is

about 5° near the tip, and effective operation can be provided at least up to 10°.

The pitch (that is, the spacing between turns) should not be too small, because the capacitive effects increase with closer turns and bypass effects of the microwave signal may take place. It has been found that for a 1 to 2-mil wire, a 5-mil spacing from center to center of the wire is desirable, but 2-mil spacing may be too close. In the latter case, the effect of the bypass capacitance may become substantial. The coupling of the helix into the microstrip **50** is made at a point where the microwave signal is weakest, i.e., on the top surface of the microstrip ribbon **50** (it is strongest in the opposite direction, along the under surface of ribbon **50** facing ground plane **27**). In addition, the diameter of the contact tip **54** should be smaller than the cross-sectional width of the microstrip **52**, so that the tip is shielded by the microstrip ribbon **52** from the ground plane **24**, and thereby is shielded from the principal microwave field that is formed between the under surface of the ribbon **50** and that ground plane.

From a practical standpoint, the solderless connection of the helical wire **62** to the microstrip **59** is extremely important. The connection, achieved by a spring force applied by C-spring **88** to the conical support **58** of the spiral wire, is both simple and effective. The conical opening **68** positions the conical support **58** to locate its extremely small tip **56** (e.g. 0.01 inch) on the small ribbon **50** and also serves as a ground plane shield. The ground plane shielding of the helical line **62** by passage through the opening **66, 68** in grounded cover **26**, insures freedom from spurious responses due to the various modes that can propagate along such a helical line. The shielding insures that conditions suitable for the propagation of the higher order modes are not satisfied over the operating frequency range. Thus the shielding preserves the character of the single-wire line as one which propagates a surface wave having circular magnetic lines. Due to the shielding, the helix **62** within opening **68** may operate in the TEM mode rather than the surface-wave mode with circular magnetic lines. However, this is without harmful effect. Only a portion of the coil is shielded, and a portion is unshielded (about one-third of the height of the helix), in order to keep the cover ground plane **23** as far away as possible from the microstrip ribbon **50**; and the fully shielded portion also seems to be effective in preventing propagation of other modes than the desired surface-wave mode. In order to prevent undue reflections produced by a transition from the unshielded portion of the helix to the shielded portion, the shield opening is formed with a cylindrical section **66**, so that the conical spiral approaches it gradually, to provide a smooth transition.

The conical dielectric support **58** is provided with deep V-shaped grooves so that the uninsulated wire **62** seats therein with a substantial spacing from the conical ground plane opening **68.** This spacing avoids any shorting of the uninsulated wire. In addition, due to the dielectric constant of the conical support **58,** and the deep imbedding of the wire within that, support so that most of the electric field is located within the dielectric, the effective length of the helical line is made substantially greater. This helix involves some **24** inches of wire. This conical spiral bias line is especially useful in its application to the above described attenuator, but it is not limited thereto and can be used for coupling into any radio-frequency device; for example, where it is desired to supply electrical energy for modulation. Thus, if one wanted to modulate up to 100 megacycles to achieve a 10-nanosecond response, this conical spiral line would provide a suitable transmission line for that purpose, although there would be some loss of the modulating signal due to the substantial resistance of the spiral. This spiral line is especially useful for coupling into microwave devices of very small dimension, as in the microstrip attenuator described above.

The difference between this device and a high frequency choke or coil should be noted. Such chokes are closely wound, insulated wires, but they act as a metallic sheath due to the interwinding capacitance at high frequencies, and their operation is unpredictable when used to couple into a microwave line handling frequencies above 100 megacycles.

The propagating space **22** in housing block **12** is generally in the form of a rectangular U cross-section (as indicated in FIG. **5**) formed by the walls of block **12** (in the center region, stud **24** forms the bottom wall of the U). Thus, the walls of this U-shaped space **22** are in the nature of a "trough" ground plane, in which a metallic ribbon or microstrip **50** extends along the bottom length of the trough and is spaced by a dielectric from the bottom wall of the trough and is remotely and equally spaced from the vertical walls on either side. In this construction, the dimensions are chosen to form a transmission line propagating in the TEM mode; such a transmission line may be called a "trough line." This trough line construction of the present invention is provided with a cover ground plane in the form of the lower face **23** of member **26**, which forms a rectangular waveguide with the side and bottom walls of the trough and extends therealong for the length of propagating space **22.** This top ground plane **23,** as described above, is sufficiently remote from the microstrip **50** as to have negligible effect on the microwave signals, and also provides a shield and ground plane for the upper portion of the spiral line **62** in its transition from a single-wire transmission line.

This hollow waveguide construction formed by the cover ground plane **23** and the trough is designed to control the spectrum of the next propagating mode. That is, the dimensions of the trough space **22** (formed by the side walls of block **12,** the upper face of stud **24,** and face **23**) are chosen so that the resulting hollow waveguide cannot propagate a rectangular waveguide mode up to the highest operating frequency of the microwave attenuator (e.g. 18 GHz). The unit is thereby designed so that the highest operating frequency which the attenuator microstrip line is designed to propagate is below the frequency of any hollow waveguide mode that the closed trough line can propagate within its walls as a hollow waveguide. Thus, the trough line is limited to propagation of the single TEM mode for which it is designed. The rectangular shape of the hollow waveguide has some advantages over other shapes (e.g., circular) in that in the former there is a larger spacing of the next higher modes so that the waveguide dimensions can be chosen to exclude more effectively such higher modes from propagation.

The single-wire continuous transmission line formed by spiral **62** functions as a very compact miniature low pass filter which above cut-off frequency maintains a high input impedance, relative to that of the 50-ohm transmission line, over the extreme frequency range of

100 mc to 18 GHz, and has a moderately low impedance in the low pass region. This single-wire line operates in a single mode over the frequency range of interest, so that no lumped or periodic discontinuities are introduced, a condition which assures a generally smooth behavior over the entire frequency band. Furthermore, it provides a natural way of making connection to the signal conductor strip of a microwave line, such as the above described attenuator, without introducing other ground connections.

A consideration of the spiral coupling **62** in terms of the characteristics of a single wire transmission line may be helpful. As discussed by Stratton in "Electromagnetic Theory," McGraw-Hill, 1941, pages 524–537, the surface wave of interest is the symmetric $n = 0$ mode which is a TM mode, with the magnetic field being entirely transverse to the axis of the wire. This mode has the lowest attenuation of all modes and is the only one which survives after some small initial distance from the point of excitation. The surface impedance Z_s is by definition the ratio of the axial component of the electric field at the surface of the wire and the total current flowing in it. Accordingly, the surface impedance Z_s = (volt/length)/amp is expressed in ohms per unit length of line. If the depth of penetration d is small compared to the radius of the wire, then the surface impedance is approximately given by

$$Z_s = 1.33/a\ (P_r S_r f)^{1/2}\ (1 + j)\ \text{ohms/inch} \qquad (1)$$

where a = radius of wire in mil
P_r = relative permeability (at f)
S_r = relative resistivity of copper
f = frequency in GHz.

At the low frequencies, where the skin depth becomes large relative to the radius of the wire, equation 1 does not apply. In this case, Z_s reduces to the d-c resistance of the wire. Thus, at the high frequencies, where this equation applies, it is seen that Z_s is complex; in fact, that it is inductive, the inductive reactance per unit length being equal to the resistance per unit length. Inspection of equation 1 shows that Z_s increases inversely with a, and is proportional to the square root of P_r, S_r, and f. Thus, if it is desired to make Z_s large, one should make P_r as large as possible, since this quantity does not enter into the d-c resistance and thus makes for an efficient design. With respect to the choice of S_r and a, it is seen that $Z_s \approx (S/a^2)^{1/2} \approx R_o^{1/2}$ is proportional to the square root of the d-c resistance R_o, so that an increase in Z_s via these parameters will also increase the d-c resistance, but with relatively minor effect on the bias signals. Thus, for instance, a copper wire 1 mil in diameter at 300 mc has a surface impedance Z_s of slightly over 2 ohms per inch, whereas nickel wire with P_r about 150 and S_r equal to 4.55, has a surface impedance Z_s of about **54** ohms per inch, and a d-c resistance of 3.9 ohms per inch. The use of equation 1 is justified since d, the skin depth, for the assumed values turns out to be $d = 0.026$ mil, which is very much less than $a = 0.5$ mil.

In the simplified cross-sectional diagram of FIG. **5**, it is assumed that over the frequency range of interest, no mode other than the TEM mode can propagate inside the space bounded by the grounded walls of **12'**, **23'** and **24'**. Representation of the effect of the loading of the single wire **122** on the microstrip **50'** may be difficult to present rigorously for although the homogeneous solution for the single-wire line **122** (i.e., the characteristics of the freely propagating mode along an axial cylinder) is known, the counterpart for a single-wire line immersed in an external RF microwave field as in FIG. **5** is not known. Nevertheless, there are some important deductions that can be made. At low frequencies where the length of the wire line **122** is very small relative to wavelength of the microwave frequencies, the current in wire **122** is substantially uniform. Evaluating the line integral of the electric intensity E along the line **122** from the strip **50'** to ground **23'**, one obtains

$$\overline{\varphi AB} = \int_A^B \bar{E} \cdot \bar{ds} = \int_A^B Z_s \bar{\imath} \cdot \bar{ds} = Z_s i l \qquad [2]$$

where l is the length of the line **122**. Since the field is Laplacian in the cross-sectional plane of FIG. **5**, ϕAB is the voltage between the strip **50'** and ground **24'**, and thus, for this condition, the equivalent circuit of FIG. **6** is applicable, where the characteristic impedance of the microstrip and the matched generator **92** are each represented by Zo and the shunting impedance of wire **122** is represented by $Z_s l$. The insertion loss L(db) under the above conditions is given by

$$L(\text{db}) = 20 \log_{10} / 1 + Z_s' 1 / 2\ Z_s' 1 / \qquad (3)$$

where Z_s' is the normalized (to Zo) surface impedance per unit length; which is the result that should be obtained as the d-c conditions are approached. Over the range where equation 3 applies, small insertion losses can only be obtained if $| Z_s' | l >> 1$. However, in view of the above-noted impedance values obtained with copper and nickel, which is feasible if l is made very much larger than the cross-sectional dimension. But this design of wire **122** was predicated on the validity of the equivalent circuit of FIG. **6**, which now comes into question inasmuch as the shunting impedance $Z_s l$ cannot be treated as a lumped element, but must be treated as a distributed transmission of length l terminated in a short circuit. The finding that by increasing the length of the single wire, one path reduces its loading effect on the 50-ohm transmission line may be explained by the coupling to the surface wave being correspondingly decreased. Since the amplitude of the induced surface wave is proportional to the unperturbed electric field intensity of the external microwave field along the initial path of wire **122**, it follows that by covering the same vertical distance in FIG. **5** from strip **50'** to some intermediate point via a longer path thus permitting the unperturbed field to intersect this path more closely at right angles, the initial E_s, and thus the coupling to the surface wave, is reduced. Although the circuit of FIG. **6** is not directly applicable at the higher frequencies, it nevertheless gives qualitatively the correct dependence, i.e., the coupling to the surface wave is properly expressed by the product of $Z_s l$. If $| Z_s l | >> Z_o$(50 ohms), the attenuation along this line is sufficiently large so that the termination at ground **23'** is unimportant. However, at the lower frequencies, below 500 mc, and using about 20 inches of wire length, the reflections cannot be ignored. The energy of the surface wave is carried in the electric field space surrounding the wire, and thus, if the line is conductively or capacitively connected (an effective r-f short)

to a metallic surface at ground **23'**, the wave is fully reflected at this point. Consequently, the external (external to the wire) impedance of the surface wave $Z_z = ErH_\theta$ in the propagating direction at point **56'** (where z, r and θ are the cylindrical coordinates) increases as the frequency is approached where the line length l equals a quarter-wavelength and decreases as the frequency is approached where l equals a half-wavelength. Since the surface current I_s in the propagating direction is directly proportional to H_θ at the surface, it is clear that the current loading of the line at the microstrip **50'** must go through these same variations. Consequently, one should expect increased loading effects at frequencies where l is an integral number of half-wavelengths, and decreased loading where l is an odd number of quarter-wavelengths. However, as previously mentioned, since Z_s increases with frequency, the line attenuation increases so that these variations tend to disappear. Experimental results show, in fact, that the disappearance of these higher resonances is more rapid than would be expected on the basis of the increase in Z_s treated as a transmission line. This may be due to the fact that the external field, which in the cross-sectional plane is in equiphase, is getting more out of phase with the propagating surface wave as the frequency increases. Thus, it is reasonable to assume that only an initial fixed electrical length, probably less than $\pi/2$, is responsible for launching the surface wave. Since this length is inversely proportional to the frequency, it is seen that the current loading due to this effect and Z_s should at the higher frequencies vary as $f^{-3/2}$ rather than $f^{-1/2}$.

The length of coupling wire **122** can be increased beyond the internal dimensions so as to allow only the surface wave to propagate and to launch this wave in such a way as to bring the full length of the line into play. In the case of the conventional coaxial structures, this may be accomplished by constructing the line in the form of an Archimedes spiral (as shown and described in the aforementioned paper of applicant, here incorporated by reference). In this way, the electric intensity E_s is reasonably well distributed over the length of the spiral, so that the external field intersects the spiral more and more at right angles. Experimental results indicate that over the frequency range where the coaxial line operates in the TEM mode alone, the spiral tends to operate in the surface mode alone, particularly if the pitch of the spiral is equal to about five wire diameters. In a stripline construction, a spiral coupling line is not preferred because of the desired electric field configuration and the miniature dimensions of the microstrip line. The conical spiral, as discussed above, has been found to be fully suitable.

Various other modifications of this invention will be apparent from the above description and the illustrated embodiments and from applicant's aforementioned prior application Ser. No. 788,254. For example, instead of a solid spiral wire **62**, thin films of high resistance (such that the surface impedance becomes purely resistive) may be used where the bias currents are small. Such thin films may be deposited in the spiral groove of a conical dielectric support such as the cone **58**. Also, more than two shunt diodes may be used. A wideband conically spiral coupling device is effective to supply low frequency bias currents to a compact structure such as a microstrip without impairing the operation at microwave frequencies. The overall unit is designed to operate over a wide band of microwave operating frequencies including multi-gigahertz, and a generally flat response characteristic is achieved over that band. A closed trough line structure in the form of a hollow waveguide is provided for enclosing a TEM microstrip line having a coupling line connected thereto, and the structure is so dimensioned as to restrict propagation to the single TEM mode.

What is claimed is:

1. In a microwave transmission system having a characteristic impedance, a coupling device for supplying low frequency electrical energy to a part of said system and for presenting a large impedance to microwave signals from said system part, said coupling device comprising:

an electrical conductor characterized by resistance at microwave operating frequencies a plurality of times greater than said characteristic impedance, said conductor having an aperiodic meander;

separate terminal means connected to spaced portions of said meander conductor for receiving said electrical energy and for connection to said system;

and means for mounting said conductor with its meander path transverse to the electric field of the transmission system.

2. In a microwave transmission, a coupling device as recited in claim **1** wherein said conductor is in the form of a conical helix.

3. In a microwave transmission system including a strip signal conductor, means providing a ground plane and a dielectric therebetween;

means for coupling to said strip conductor for passing low frequency signals and for presenting a high impedance to high frequency microwave signals transmitted by said strip conductor, said coupling means including a conical helical conductor having the small diameter end thereof connected to said strip conductor.

4. A microwave transmission system as recited in claim **3** wherein at least a portion of said conical helical conductor is a single-wire transmission line.

5. A microwave transmission system as recited in claim **3** wherein said helical conductor is of magnetic material.

6. A microwave transmission line as recited in claim **3** wherein the lead angle of each turn of said helical conductor is less than 10°.

7. A microwave transmission system as recited in claim **3,** wherein a grounded shield surrounds said coupling conductor over a major fraction of the length thereof.

8. A microwave transmission system as recited in claim **3,** wherein said helical conductor is characterized by a resistance at microwave frequencies which is a plurality of times greater than the characteristic impedance of said microwave transmission system.

9. A microwave transmission system as recited in claim **3** wherein the axis of said helical conductor is substantially perpendicular to said strip conductor.

10. A microwave transmission system as recited in claim **9** wherein said helical conductor is connected to the surface of said strip opposite from said ground plane means and extends away from said ground plane means.

11. A microwave transmission system as recited in claim **10** wherein said coupling means includes means

at said small helical end having a width less than the width of said strip.

12. A microwave transmission system as recited in claim **3** wherein said coupling means includes a conical member formed of non-conductive material and said helical conductor is wound thereon.

13. A microwave transmission system as recited in claim **12** wherein said coupling means includes resilient means for biasing said conical member toward said strip conductor to connect said helical conductor thereto.

14. A microwave transmission system as recited in claim **12** wherein said conical member has a helical groove around the conical surface thereof, and said helical conductor is embedded within said groove.

15. A microwave transmission system as recited in claim **12,** and further comprising metallic means connected to said ground plane and having a conical opening surrounding a portion of said conical member and of said embedded helical conductor as a shield therefor.

16. A microwave device comprising input and output means having signal and ground conductors, a trough-shaped metallic housing coupled to said ground conductor, a strip conductor coupled to said signal conductor and extending along the bottom wall of said trough and spaced therefrom by a dielectric and spaced from the side walls of said trough housing, said housing being dimensioned to limit the mode of transmission, a dielectric support member, a resistive coupling conductor aperiodically meandering around said support member for supplying low frequency signals to said strip conductor and extending through said trough and having a resistance at microwave frequencies a plurality of times greater than the characteristic impedance of said microwave device, and a grounded metallic shield spaced from and surrounding said coupling conductor between the walls of said trough and extending along said coupling conductor for a substantial length thereof and spaced from said strip conductor, said shield having an opening for receiving said dielectric support member therein and for positioning said support member and said coupling conductor with respect to said strip conductor.

17. A microwave device as recited in claim **16,** wherein said dielectric support has grooves around the outer surface thereof approximately parallel to said strip conductor, and said coupling conductor is turned into said grooves.

18. A microwave device as recited in claim **17,** wherein said dielectric support has a conical outer surface, and said shield opening is conical to receive and position said dielectric support.

19. A microwave device as recited in claim **18,** wherein the conical tip of said dielectric support is metallic to provide a conductive contact with a surface of said strip conductor, and said coupling conductor has an end connected to said metallic support tip.

20. A microwave device as recited in claim **19,** and further comprising means for mechanically biasing said dielectric support to maintain contact with the surface of said strip conductor.

21. A microwave device comprising input and output means having signal and ground conductors, a trough-shaped metallic housing coupled to said ground conductor, a strip conductor coupled to said signal conductor and extending along the bottom wall of said trough and spaced therefrom by a dielectric and spaced from the side walls of said trough housing, said housing being dimensioned to limit the mode of transmission, a coupling conductor for supplying low frequency signals to said strip conductor and extending through said trough, and a grounded metallic shield surrounding said coupling conductor between the walls of said trough and extending along said coupling conductor for a substantial length thereof and spaced from said strip conductor, said coupling conductor being a spiral helix.

22. A microwave device as recited in claim **21** wherein said shield includes an opening having at least part thereof in a conical shape for surrounding and shielding a portion of said spiral helix.

23. A microwave device as recited in claim **22** wherein said shield opening includes a cylindrical part surrounding and shielding another portion of said spiral helix and providing a transition between an unshielded portion of said helical conductor and said portion shielded by said conical opening.

24. In a microwave transmission line including a metallic ground plane housing and a conductor within and spaced from said housing for transmitting microwave signals over a wide band;

means for coupling to said microwave conductor to supply low frequency signals thereto, said coupling means having a high impedance to said microwave signals a plurality of times greater than that of the characteristic impedance of said microwave line, and including:

a single-conductor transmission line having a resistive conductor extending through the space of said housing to said microwave conductor along an aperiodic meander,

and a grounded metallic shield within said housing surrounding said single-conductor transmission line over a substantial length thereof and spaced from said microwave conductor whereby modes of propagation by said single-conductor line are limited.

25. A microwave transmission line as recited in claim **24,** wherein said single-conductor transmission line has a substantial resistance at microwave frequencies a plurality of times greater than the characteristic impedance of said microwave conductor.

26. A microwave transmission line as recited in claim **25,** wherein said resistive conductor is a coil.

27. A microwave transmission line as recited in claim **26,** wherein said coil is a spiral helix.

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