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(54) **Method for selecting a phase shift, phase shifter, beamformer and antenna array**

(57) An RF reflection type phase shifter (RTPS) comprising a coupling device coupled to multiple reflective loads, n , respective ones of which include at least one variable reactance device with maximum reactance, $X_{\max}$, and, minimum reactance, $X_{\min}$, and at least two

impedance transformers the characteristic impedances of which are selected according to predefined criteria to provide an increase in the value of a phase shift to be applied to a signal input to the RTPS proportional to the value of n .

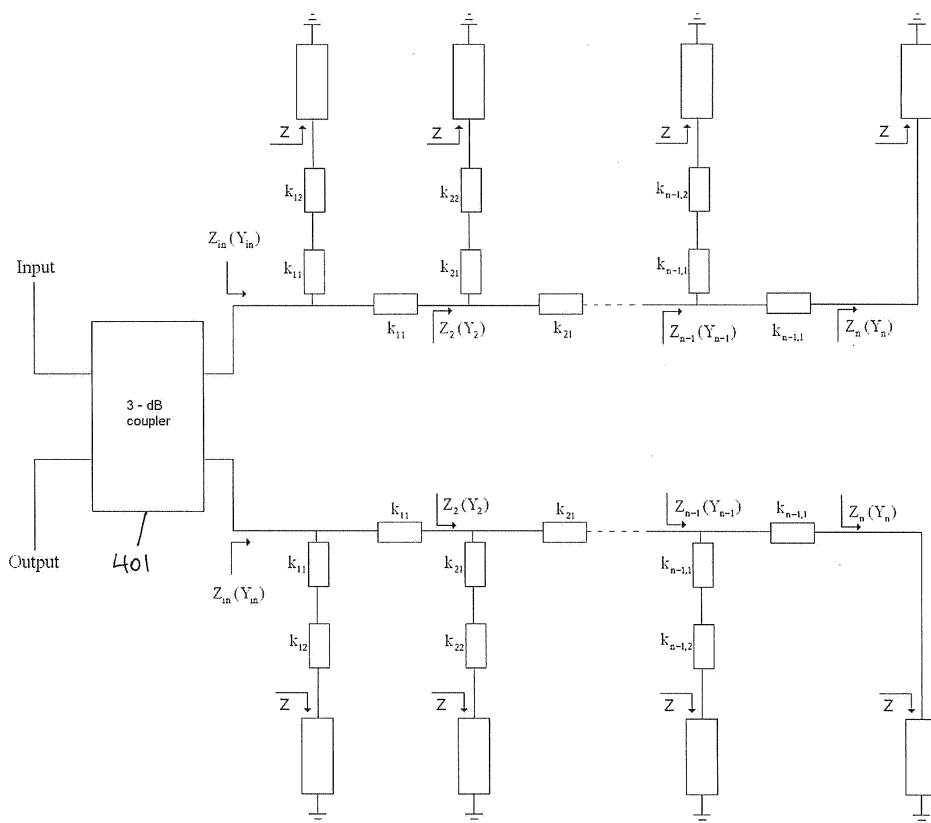


Figure 4

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Description

TECHNICAL FIELD

5 **[0001]** Aspects relate, in general, to a method for selecting a phase shift, a phase shifter, a beamformer and an antenna array.

BACKGROUND

10 **[0002]** A phase shifter provides a controllable phase shift of an input RF signal, and such devices are therefore omnipresent in telecommunications where it is desirable to modify the phase of signals. The devices also find utility in radar systems, amplifier linearization, point-to-point radio and RF signal cancellation and so on. The choice of the phase shifter for a particular application is influenced by many factors; for example, the amount of obtainable phase shift, the insertion losses and power handling capability.

15 **[0003]** For lower power handling capabilities, the variation of phase shift is typically obtained by using semiconductor technology. In particular, varactor and PIN diodes can be used as the reactance/resistance tuneable elements needed for the variation of insertion phase.

SUMMARY

20 **[0004]** According to an example, there is provided an RF reflection type phase shifter (RTPS) comprising a coupling device coupled to multiple reflective loads, n , respective ones of which include at least one variable reactance device with maximum reactance, $X_{\max}$, and, minimum reactance, $X_{\min}$, and at least two impedance transformers the characteristic impedances of which are selected according to predefined criteria to provide an increase in the value of a phase shift to be applied to a signal input to the RTPS proportional to the value of n . The characteristic impedances can be selected in accordance with a selected value of a parameter, q , determined, for a given reflective load, according to:

30
$$q = \frac{\sqrt{X_{\max} X_{\min}}}{Z_0}$$

35 where Z_0 is the characteristic impedance of the coupling device and interconnecting microstrip lines. The impedance of the or each variable reactance device can be altered by varying a DC voltage or current applied to the variable reactance device. The or each variable reactance device can be a varactor diode. The impedance transformers can be one quarter-wavelength long at a selected frequency of operation. The impedance transformers can use microstrips or stripline technology. The impedance transformers can use lumped circuit elements, such as capacitors and/or inductors for example. The coupling device can be a circulator or a 3-dB coupler.

40 **[0005]** According to an example, there is provided a method for selecting a phase shift value for an output signal of a phase shifter including a coupling device coupled to multiple reflective loads, n , respective ones of which include at least one variable reactance device with maximum reactance, $X_{\max}$, and, minimum reactance, $X_{\min}$, and at least two impedance transformers, the method comprising selecting the characteristic impedances of impedance transformers according to predefined criteria to provide an increase in the value of a phase shift to be applied to a signal input to the phase shifter proportional to the value of n . Characteristic impedances can be selected in accordance with a selected value of a parameter, q , determined, for a given reflective load, according to:

50
$$q = \frac{\sqrt{X_{\max} X_{\min}}}{Z_0}$$

55 where Z_0 is the characteristic impedance of the coupling device and interconnecting microstrip lines. The impedance of the or each variable reactance device can be modified by varying a DC voltage or current applied to the variable reactance device.

[0006] According to an example, there is provided a beamformer for use in a wireless telecommunication network to modify the transmission profile of an antenna array, the beamformer including an RF reflection type phase shifter (RTPS)

comprising a coupling device coupled to multiple reflective loads, n , respective ones of which include at least one variable reactance device with maximum reactance, $X_{\max}$, and, minimum reactance, $X_{\min}$, and at least two impedance transformers the characteristic impedances of which are selected according to predefined criteria to provide an increase in the value of a phase shift to be applied to a signal input to the RTPS proportional to the value of n . The antenna array can be composed of multiple antennas, wherein the phase shifter is operable to modify the phase of respective signals input to the multiple antennas. The phase shifter can comprise multiple parallel phase shift stages to apply respective different phase shifts to the respective signals input to the multiple antennas.

[0007] According to an example, there is provided an antenna array in a wireless telecommunication network, the antenna array including multiple antennas operable to receive input from a beamformer that is operable to modify the transmission profile of the antenna array, the beamformer including an RF reflection type phase shifter (RTPS) comprising a coupling device coupled to multiple reflective loads, n , respective ones of which include at least one variable reactance device with maximum reactance, $X_{\max}$, and, minimum reactance, $X_{\min}$, and at least two impedance transformers the characteristic impedances of which are selected according to predefined criteria to provide an increase in the value of a phase shift to be applied to a signal input to the RTPS proportional to the value of n .

BRIEF DESCRIPTION OF THE DRAWINGS

[0008] Embodiments will now be described, by way of example only, with reference to the accompanying drawings, in which:

Figure 1 is a schematic representation of a reflective type phase shifter (RTPS);

Figure 2 is a schematic representation of an RTPS device;

Figure 3 is a schematic representation of an RTPS device;

Figure 4 is a schematic representation of a generic circuit consisting of a coupling device, transformers $k_{i,j}$ and impedance z according to an example;

Figure 5 is a schematic representation of the reflective load of the generic circuit of a multiple active element reflective load RTPS as shown in figure 4;

Figure 6 is a schematic representation of a two active element per reflective load RTPS according to an example;

Figure 7 is graph of the insertion loss of a first order phase shifter according to an example;

Figure 8 is a graph of the insertion phase of the first order phase shifter according to an example;

Figure 9 is a graph of insertion loss of a second order phase shifter with $q=1$ according to an example;

Figure 10 is a graph of insertion phase of the second order phase shifter with $q=1$ according to an example;

Figure 11 is a graph of insertion loss of second order phase shifter with $q=1.6$ according to an example;

Figure 12 is a graph of insertion phase of second order phase shifter with $q=1.6$ according to an example;

Figure 13 is a schematic representation of a three active element per reflective load RTPS according to an example;

Figure 14 is a schematic representation of the dependence of the characteristic impedances k_{11} , k_{11} and k_{22} on k_{21} for the case when $Z_0=50\Omega$ and $q=1$ according to an example;

Figure 15 is a graph of the insertion loss of a third order phase shifter according to an example with $q=1$;

Figure 16 is a graph of insertion phase of third order phase shifter according to an example with $q=1$;

Figure 17 is a graph of insertion loss of third order phase shifter according to an example with $q=1.6$; and

Figure 18 is a graph of insertion phase of third order phase shifter according to an example with $q=1.6$.

DESCRIPTION

[0009] Example embodiments are described below in sufficient detail to enable those of ordinary skill in the art to embody and implement the systems and processes herein described. It is important to understand that embodiments can be provided in many alternate forms and should not be construed as limited to the examples set forth herein.

[0010] Accordingly, while embodiments can be modified in various ways and take on various alternative forms, specific embodiments thereof are shown in the drawings and described in detail below as examples. There is no intent to limit to the particular forms disclosed. On the contrary, all modifications, equivalents, and alternatives falling within the scope of the appended claims should be included. Elements of the example embodiments are consistently denoted by the same reference numerals throughout the drawings and detailed description where appropriate.

[0011] The terminology used herein to describe embodiments is not intended to limit the scope. The articles "a," "an," and "the" are singular in that they have a single referent, however the use of the singular form in the present document should not preclude the presence of more than one referent. In other words, elements referred to in the singular can number one or more, unless the context clearly indicates otherwise. It will be further understood that the terms "comprises," "comprising," "includes," and/or "including," when used herein, specify the presence of stated features, items, steps, operations, elements, and/or components, but do not preclude the presence or addition of one or more other features, items, steps, operations, elements, components, and/or groups thereof.

[0012] Unless otherwise defined, all terms (including technical and scientific terms) used herein are to be interpreted as is customary in the art. It will be further understood that terms in common usage should also be interpreted as is customary in the relevant art and not in an idealized or overly formal sense unless expressly so defined herein.

[0013] When a varactor diode is used as a tuneable reactance element in a phase shifter or phase shift device, a parameter of importance is the ratio of the maximum and minimum capacitance, referred to as $r_c = C_{\max}/C_{\min}$. Normally, the greater this ratio is, the greater is the amount of obtainable phase shift. A typical value for this ratio is usually between 3-10. On the other hand, when a PIN diode is used (usually as a switch) the parameter of importance is the ratio between the "ON" and "OFF" resistance (referred to as $r_r = R_{\max}/R_{\min}$).

[0014] Due to cost effectiveness, a reflective type phase shifter 100 as depicted schematically in figure 1 is widely used in telecommunications. Its structure is relatively simple and consists of a 3-dB coupler 101 and varactor diodes 103 as part of the circuitry of the reflective loads.

[0015] Typically, according to a relatively simplified analysis, it is assumed that the 3-dB coupler is lossless and that the "coupled" and "through" arms of the 3-dB coupler have reflection coefficients given by Γ_1 and Γ_2 respectively. Thus, the overall reflection and transmission coefficients s_{11} , s_{21} of the phase shifter 100 become:

$$\begin{aligned} S_{11} &= 0.5(\Gamma_1 - \Gamma_2); \text{ and} \\ S_{21} &= -j0.5(\Gamma_1 + \Gamma_2) \end{aligned} \quad (1)$$

[0016] If the reflective loads are the same, i.e. $\Gamma_1 = \Gamma_2 = \Gamma$, it follows that $S_{11} = 0$ and $S_{21} = -j\Gamma$, inferring that the structure formed in this way is always properly impedance matched (within the bandwidth of the 3-dB coupler) and that the reflection coefficient at the loads, Γ , becomes the transmission coefficient of the phase shifter. As such, any changes in the phase of the reflection coefficient Γ directly translates to phase changes of the transmission coefficient of the phase shifter, i.e. S_{21} .

[0017] The transmission coefficient can be written as:

$$S_{21} = -j\Gamma = \frac{Z - Z_0}{Z + Z_0} = \left| \frac{Z - Z_0}{Z + Z_0} \right| \cdot e^{j\left(-\frac{\pi}{2} + \arctan\left(\frac{\text{Im}(Z)}{\text{real}(Z) - Z_0}\right) - \arctan\left(\frac{\text{Im}(Z)}{\text{real}(Z) + Z_0}\right)\right)}$$

(eqn. 2)

where, Z is the variable impedance of the reflective load, $Z = R + jX$. This impedance may come from the varactor diode alone or from a series/parallel configuration of a varactor diode and lumped elements for example.

[0018] In order for an RTPS presented by (2) to be variable, the reactive part, X , can be controlled by electronic means,

such as by application of a dc voltage or current for example. If the resistive part of the impedance Z is constant, and if the maximum and minimum reactive part, X , of impedance Z are denoted by X_{max} and X_{min} , respectively, the amount of phase shift obtained from a single active element RTPS can be written as:

$$\Delta\Phi = \left\{ \arctan\left(\frac{X_{max}}{R-Z_0}\right) - \arctan\left(\frac{X_{max}}{R+Z_0}\right) \right\} - \left\{ \arctan\left(\frac{X_{min}}{R-Z_0}\right) - \arctan\left(\frac{X_{min}}{R+Z_0}\right) \right\} \quad (3)$$

[0019] The insertion loss arising from the resistive part of the impedance, Z , can be written as;

$$|S_{21}| = \sqrt{\frac{(R-Z_0)^2 + X^2}{(R+Z_0)^2 + X^2}} \quad (4)$$

[0020] In the case when the resistive part, R is zero, insertion phase (3) and insertion loss (4) become:

$$\Delta\Phi = 2 \left[\arctan\left(\frac{X_{max}}{Z_0}\right) - \arctan\left(\frac{X_{min}}{Z_0}\right) \right], \text{ and}$$

$$|S_{21}|_{R \rightarrow 0} = 1 \quad (5)$$

[0021] Thus, if the variable impedance device (varactor diode in this case) is ideal, i.e. the real part of impedance Z is 0, only changes to the reactive part of the impedance Z induce a change in phase. However, this is typically not achieved in practice since all semiconductor devices have a finite parasitic resistance, no matter how small. Further, in addition to the insertion loss arising from the finite resistance of the varactor diodes, a typical 3-dB coupler exhibits an insertion loss in the range of 0.3 - 0.5 dB, although this applies to a signal travelling in one direction.

[0022] In the case of reflective type configurations, the input signal travels through the 3-dB coupler, reaches the reflective loads and gets reflected from them to travel through the 3-dB coupler again and towards the output port. As such, since the signal travels twice through the 3-dB coupler, the insertion loss of a 3-dB coupler in a RTPS device is two times higher than its rated value. As such, the total insertion loss contribution of a 3-dB coupler in the RTPS configuration can easily occupy a significant portion of the total insertion loss (which consists of the losses of a 3-dB coupler and the losses due to the finite resistance of the load impedance, Z).

[0023] In the design of an RTPS, the varactor diode alone in the circuit of the reflective load usually provides a limited amount of phase shift. A reason for this is that the capacitance ratio (defined as the ratio of the maximum to minimum capacitance) of the varactor diode cannot be indefinitely increased, since this value is technology dependent. The capacitance ratio of a typical varactor diode is normally a single digit number, and almost always lower than 10. In order to increase the insertion phase of an RTPS, the varactor diode is connected in series with an inductor; however, increased phase shift comes at the expense of increased insertion loss due to the parasitic resistance of the additional inductor. Nevertheless, resonating a varactor with an inductor has become regular practice and is often pursued in the industry.

[0024] In some instances, the values of phase shift obtained by an inductor resonated varactor diode need to be even further increased. One approach to achieve this can be to use a cascaded connection of two RTPS devices, which inevitably doubles the amount of phase shift; however, it also doubles the insertion losses.

[0025] Increasing the amount of phase shift of an RTPS device without necessarily increasing the insertion losses is typically conditioned by one aspect. That is, for lower insertion losses and cost reduction, the number of 3-dB couplers needs to be kept to a minimum - one in this case. Bearing this condition in mind, the amount of obtainable phase shift can be increased by using two varactor diodes per reflective load of a 3-dB coupler, as shown in [3] and [4], and figures 2 and 3 for example.

[0026] Figure 2 is a schematic representation of an RTPS device. In the RTPS 200 of figure 2, the characteristic impedances of transformers, which in the case of figure 2 are microstrip lines, are set to $Z_{01}=50\Omega$ and $Z_{02}=70.7\Omega$ in

order to double the amount of phase shift compared to the case of a single varactor diode per reflective load. However, no analysis has ever been performed to show the full potential of the circuit in terms of the increase in phase shift and whether the characteristic impedances $Z_{01}=50\Omega$ and $Z_{02}=70.7\Omega$ of the transformers are optimal (or not).

[0027] Similarly, figure 3 is a schematic representation of an RTPS device 300 that provides an increased amount of phase shift. However, again, no clear indication of the full potential of the circuit has been determined before, and it has never been determined whether the arrangement of the diodes is optimal or not. Instead it was only stated a transformer, Z_c , needs to be added so as to compensate for the high losses of the resonant circuit formed using two varactor diodes and inductors. However, this does not quantify in any way the full capabilities of the circuit.

[0028] According to an example, a generic circuit of multiple active element load RTPS using only one 3-dB coupler is provided, which can provide optimum values of phase shift. In an example, proper selection of impedance transformers, $k_{i,j}$, which allow increase of phase shift using only one 3-dB coupler is provided. Further, since the increase of phase shift is achieved using only one 3-dB coupler, the insertion losses of the circuit according to an example are not a multiple integer of the order of the phase shifter, but are increased by only a small amount due to the presence of additional active elements per load.

[0029] Figure 4 is a schematic representation of a generic circuit 400 consisting of a 3-dB coupler 401, transformers $k_{i,j}$ and impedance z according to an example. Impedance z is used to represent any variable impedance which can be in the form of a varactor diode alone or any series/parallel combination with lumped elements for example.

[0030] The input admittance of the circuit of figure 4, Y_{in} , can be represented as:

$$Y_{in} = \frac{k_{12}^2}{k_{11}^2} Y + \frac{1}{k_{11}^2 Y_2} \quad (6)$$

Where,

$$Y_2 = \frac{k_{22}^2}{k_{21}^2} Y + \frac{1}{k_{21}^2 Y_3}, \quad Y_3 = \frac{k_{32}^2}{k_{31}^2} Y + \frac{1}{k_{31}^2 Y_4} \dots \quad (7)$$

Or, in general,

$$Y_{n-1} = \frac{k_{n-1,2}^2}{k_{n-1,1}^2} Y + \frac{1}{k_{n-1,1}^2 Y_n}, \text{ where } Y_n = Y \quad (8)$$

[0031] Here, $k_{i,j}, i=2 \dots n, j=1,2$ represent the impedance transformers, n represents the order of the absorptive filter and $Y=Z^{-1}$. It can be inferred from (7) - (8) that the input admittance, Y_{in} , can be represented in the form of a generalized continued fraction:

$$Y_{in} = \frac{k_{12}^2}{k_{11}^2} Y + \frac{1}{\frac{k_{11}^2 k_{22}^2}{k_{21}^2} Y + \frac{k_{11}^2}{\frac{k_{21}^2 k_{32}^2}{k_{31}^2} Y + \dots}} \quad (9)$$

Or equivalently,

$$Y_{in} = b_0 + \frac{a_1}{b_1 + \frac{a_2}{b_2 + \dots}} = b_0 + \sum_{k=1}^m \frac{a_k}{b_k}, k=1\dots m, m = n-1 \quad (10)$$

Where $b_0 = \frac{k_{12}^2}{k_{11}^2} Y$, $a_1=1$, $b_k = \frac{k_{k,1}^2 k_{k+1,2}^2}{k_{k,1}^2} Y$, $\forall n \geq 2, k=1\dots n-1$ and $a_k = k_{k-1,1}^2$, $\forall n \geq 3, k=2\dots n-1$

[0032] The input admittance of the n-th order reflective load can now be represented as:

$$Y_{in,n} = \frac{A_{m-1}}{B_{m-1}} \quad (11)$$

where $A_{m-1}=b_{m-1}A_{m-2}+a_{m-1}A_{m-3}$ and $B_{m-1}=b_{m-1}B_{m-2}+a_{m-1}B_{m-3}$. Solving (11) yields the n-th order admittance polynomial from which the expression for the n-th order polynomial expression for the transmission coefficient of the n-th order RTPS (with the 3-dB coupler included), S_{21} , can be derived:

$$S_{21} = -j \frac{Y_0 - Y_{in}}{Y_0 + Y_{in}} \quad (12)$$

where Y_0 is the characteristic admittance of the 3-dB coupler.

[0033] Substituting (11) into (12) and converting the admittance parameters into their impedance counterparts, i.e.

$Y_0 = Z_0^{-1}$ and $Y = Z^{-1}$, the expression for the transmission coefficient, S_{21} , as a function of impedance parameters according to an example can be represented as:

$$S_{21} = -j \frac{Z_{in} - Z_0}{Z_{in} + Z_0} \quad (13)$$

[0034] The n-th order identical and real zeroes of such a polynomial yield the expressions for the transformer impedances, $k_{i,j}, i = 2\dots n, j = 1,2$. However, polynomials of order 5 and above cannot be analytically uniquely solved for real and identical zeroes. As such, for such high orders a numerical approach can be used.

[0035] In order for (13) to offer phase shift increase commensurate with the number of pairs of active elements in the reflective loads, (13) can be represented in the following form:

$$S_{21} = -j\Gamma_{in} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} = \left| \frac{Z - Z_0}{Z + Z_0} \right|^n \cdot e^{jn \left(\frac{\pi}{2} + \arctan \left(\frac{\text{Im}(Z)}{\text{real}(Z) - Z_0} \right) - \arctan \left(\frac{\text{Im}(Z)}{\text{real}(Z) + Z_0} \right) \right)} \quad (14)$$

where n indicates the number of pairs of active elements in the circuit of the reflective load of figure 4 and figure 5, which is a schematic representation of a reflective load of the generic circuit of a multiple active element reflective load RTPS as shown in figure 4.

[0036] More generally, (14) can be written as:

$$S_{21} = -j\Gamma_{in} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} = \left| \frac{Z - qZ_0}{Z + qZ_0} \right|^n \cdot e^{jn \left(-\frac{\pi}{2} + \arctan \left(\frac{\text{Im}(Z)}{\text{real}(Z) - qZ_0} \right) - \arctan \left(\frac{\text{Im}(Z)}{\text{real}(Z) + qZ_0} \right) \right)} \quad (15)$$

where q indicates the position of the transmission zero on the resistance scale. According to an example, the value of q can be adjusted with a proper selection impedance transformers $k_{i,j}, i=2\dots n, j=1,2$. The phase shift provided by (15) can be written as:

$$\Delta\Phi = n \left\{ \arctan \left(\frac{X_{max}}{R - qZ_0} \right) - \arctan \left(\frac{X_{max}}{R + qZ_0} \right) \right\} - \left\{ \arctan \left(\frac{X_{min}}{R - qZ_0} \right) - \arctan \left(\frac{X_{min}}{R + qZ_0} \right) \right\} \quad (16)$$

[0037] For $q=1$, the phase shift of the proposed structure of figures 4 and 5 is increased n -times. Nevertheless, simply setting $q=1$, does not necessarily result in the optimal phase shift. The optimal phase shift is found by finding the roots of:

$$\frac{\partial \Delta\Phi}{\partial q} = 0 \quad (17)$$

yielding the following 6th order polynomial

$$P_1 q^6 + P_2 q^4 + P_3 q^2 + P_4 = 0 \quad (18)$$

Where

$$\begin{aligned} P_1 &= Z_0^6, \quad P_2 = -R^2 Z_0^4 - Z_0^4 X_{max} X_{min} + Z_0^4 X_{max}^2 + Z_0^4 X_{min}^2, \\ P_3 &= -R^4 Z_0^2 - 6R^2 Z_0^2 X_{min} X_{max} - Z_0^2 X_{max} X_{min}^3 + Z_0^2 X_{max}^2 X_{min}^2 - Z_0^2 X_{max}^3 X_{min} - 2R^2 Z_0^2 X_{max}^2 - 2R^2 Z_0^2 X_{min}^2, \\ P_4 &= R^6 - R^4 X_{min} X_{max} - R^2 X_{min}^3 X_{max} + R^2 X_{max}^2 X_{min}^2 - R^2 X_{min}^3 X_{max} + R^4 X_{max}^2 + R^4 X_{min}^2 - X_{max}^3 X_{min}^3 \end{aligned} \quad (19)$$

[0038] The first four roots of (18) are complex conjugate, while the remaining two roots are real with equal magnitude, but opposite signs. As such, there is always one solution to (18) that yields the optimum value of the parameter q .

[0039] The expression given by (19) can be simplified if it can be assumed that the parasitic resistance of the varactor diode can be neglected. This is a valid assumption in most cases, since this resistance is typically of the order of 1- 2 ohms. By setting $R=0$ in (18) the optimal value for q according to an example becomes:

$$q = \frac{\sqrt{X_{max} X_{min}}}{Z_0} \quad (20)$$

[0040] The insertion loss of the proposed RTPS (log scale) is:

$$|S_{21}| = 20n \log_{10} \left| \frac{Z - qZ_0}{Z + qZ_0} \right| + \text{insertion loss of 3-dB coupler} \quad (21)$$

[0041] Here, the first term on the right represents the insertion loss of the reflective circuit of the proposed RTPS. For $q=1$ the insertion loss of the proposed reflective load is n -times higher than the insertion loss of the first order reflective circuit given by (4), while if parameter q is set in accordance with (20), the insertion loss of the reflective loads is always lower than that achieved with $q=1$. The second term on the right is the insertion loss of a 3-dB coupler.

[0042] For comparison, a cascade connection of n first order circuits will yield the same phase shift as (16), however, its insertion loss will be:

$$|S_{21}| = 20n \log_{10} \left| \frac{Z - qZ_0}{Z + qZ_0} \right| + n \cdot (\text{insertion loss of 3-dB coupler}) + (n - 1) \cdot \text{loss due to interconnections} \quad (22)$$

[0043] In quantitative terms, the reduction in the overall insertion loss over the cascade connection is:

$$\Delta |S_{21}| = (n - 1) \cdot (\text{insertion loss of 3-dB coupler}) + (n - 1) \cdot \text{loss due to interconnections} \quad (23)$$

[0044] Thus, in an example, in view of (22) and (23), (16) and (21), the amount of phase shift of RTPS can be increased in a linear fashion with respect to the pairs of active elements, without increasing the insertion loss in the same linear fashion. For example, if the insertion loss of a 3-dB coupler is 0.3 dB ($2 \cdot 0.3$ dB in RTPS configuration), and for $n=2$, $q=1$ the reduction of the insertion loss using the circuit over according to an example compared to a conventional cascade connection is:

$$\Delta |S_{21}| = 0.6 \text{ dB} + \text{loss due to interconnections} \quad (24)$$

[0045] In the above equations the condition stipulated above relating to the retention of a minimum number of 3-dB couplers in the design of RTPS has been fulfilled.

[0046] According to an example, two RTPS - one with two active elements per reflective load and the other with a three active elements per reflective load - are described in more detail, although it will be appreciated that other variations are possible.

Two active element per reflective load

[0047] A two active element per reflective load RTPS 600 according to an example is schematically represented in figure 6. Setting $n = 2$ in (10) and substituting (10) into (13), the following expression for the transmission coefficient is obtained:

$$S_{21} = -j\Gamma = j \frac{-Zk_{11}^2 + Z_0k_{12}^2 + Z_0Z^2}{Zk_{11}^2 + Z_0k_{12}^2 + Z_0Z^2} \quad (25)$$

[0048] Equation (25) assumes that the 3-dB coupler 601 is ideal. The zeroes of (25) yield the following values for the transformers k_{11} and k_{12} :

$$k_{11} = Z_0 \sqrt{2q} \text{ and } k_{12} = Z_0 q \quad (26)$$

[0049] Upon which (26) becomes:

$$S_{21} = -j\Gamma_{in} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} = \frac{Z - qZ_0}{Z + qZ_0} \cdot e^{j2\left(-\frac{\pi}{2} + \arctan\left(\frac{\text{Im}(Z)}{\text{real}(Z) - qZ_0}\right) - \arctan\left(\frac{\text{Im}(Z)}{\text{real}(Z) + qZ_0}\right)\right)} \quad (27)$$

[0050] By setting $q=1$ it follows that $k_{11} = Z_0 \sqrt{2}$ and $k_{12} = Z_0$ which is identical to the result in [3]. As such, the result reported in [3] is a special case of the circuit of figure 6. However, in order to increase the amount of phase shift obtainable from the circuit of figure 6, parameter q is selected in line with (20) according to an example, with knowledge of the maximum and minimum reactance of the varactor diode.

[0051] For example, a varactor diode can have the following variation of capacitance: $C_{\min} = 0.4 \text{ pF}$ and $C_{\max} = 1.6 \text{ pF}$, with a parasitic junction resistance of $R_p = 1 \Omega$. The 3-dB coupler has a rated insertion loss of 0.3 dB and its operating frequency range is 2.3-2.7 GHz. The insertion loss and phase shift performance of a first order phase shifter according to an example are depicted in figures 7 and 8. The capacitance of the varactor diode varies between 0.4 pF to 1.6 pF in the steps of 0.1 pF. That is, figure 7 is graph of the insertion loss of a first order phase shifter according to an example in which capacitance varies from 0.4 pF to 1.6 pF in steps of 0.1 pF, and figure 8 is a graph of the insertion phase of the first order phase shifter.

[0052] Figures 9-12 represent the insertion loss and phase shift performance of second order phase shifters with $q=1$ and $q=1.6$, respectively. In the first case, the parameter $q=1$ and the second case corresponds to the case when parameter q attains the optimum value obtained by the use of (20), to yield the value of $q=1.6$.

[0053] Figure 9 is a graph of insertion loss of a second order phase shifter with $q=1$ (capacitance varies from 0.4 pF to 1.6 pF in steps of 0.1 pF). Figure 10 is a graph of insertion phase of the second order phase shifter with $q=1$ (capacitance varies from 0.4 pF to 1.6 pF in steps of 0.1 pF). Figure 11 is a graph of insertion loss of second order phase shifter with $q=1.6$ (capacitance varies from 0.4 pF to 1.6 pF in steps of 0.1 pF), and figure 12 is a graph of insertion phase of second order phase shifter with $q=1.6$ (capacitance varies from 0.4 pF to 1.6 pF in steps of 0.1 pF).

[0054] As can be seen from figure 9, the phase shift obtained from the second order circuit with $q=1$ is approximately two times greater than that achieved using the first order circuit, however, the insertion loss is not doubled, but increased by only a small amount - 0.2 dB in this case.

[0055] The second order phase shift realization with $q=1.6$ offers an even greater phase shift than its second order counterpart with $q=1$. That is, the phase shift of the second order circuit with $q=1.6$ is increased by an average of 10° across the indicated frequency range compared to the phase shift of the second order circuit with $q=1$. The insertion loss performance of the second order phase shifter with $q=1.6$ is also superior compared to its second order counterpart with $q=1$, exhibiting a 0.1 dB smaller insertion loss. As such, the optimal choice of parameter q not only increases the amount of phase shift obtained from a particular circuit, but it also results in reduced insertion losses.

Three active element per reflective load

[0056] A three active element per reflective load RTPS 1300 is schematically represented in Figure 13. Setting $n = 3$ in (10) and substituting (10) into (13), the following expression for the transmission coefficient is obtained:

$$S_{21} = -j\Gamma = j \frac{aZ^3 + bZ^2 + cZ + d}{-aZ^3 + bZ^2 - cZ + d} \quad (27)$$

where, $a = -k_{11}^2$, $b = Z_0 k_{12}^2 + Z_0 k_{21}^2$, $c = -k_{11}^2 k_{22}^2$ and $d = Z_0 k_{12}^2 k_{22}^2$. The transmission zero condition is achieved by setting $S_{21} = 0$. In this case, a third order polynomial in z is obtained and is solved so that it has a multiple and real root. This

is accomplished by setting the discriminant, Δ , to be zero:

$$\Delta = 18abcd - 4b^3d + b^2c^2 - 4ac^3 - 27a^2d^2 = 0 \quad (28)$$

[0057] The condition that the discriminant of (28) is zero yields a triple zero at:

$$Z = -\frac{b}{3a} = \frac{Z_0k_{12}^2 + Z_0k_{21}^2}{3k_{11}^2} \quad (29)$$

[0058] Solving (28) one obtains a quart-quadratic equation in k_{11}^4 , given by:

$$Ak_{11}^8 + Bk_{11}^4 + C = 0 \quad (30)$$

where, $A = -4k_{22}^4$, $B = -8Z_0^2k_{12}^4k_{22}^2 + 20Z_0^2k_{12}^2k_{22}^2k_{21}^2 + Z_0^2k_{22}^2k_{21}^2$, and

$$C = -4Z_0^4k_{12}^8 - 4Z_0^4k_{12}^2k_{21}^4 - 12Z_0^4k_{12}^6k_{21}^2 - 12Z_0^4k_{12}^4k_{21}^4.$$

[0059] The double zero in k_{11}^4 is achieved at:

$$k_{11}^4 = -\frac{B}{2A} = \frac{-8Z_0^2k_{12}^4k_{22}^2 + 20Z_0^2k_{12}^2k_{22}^2k_{21}^2 + Z_0^2k_{22}^2k_{21}^4}{8k_{22}^4} \quad (31)$$

with a condition that the discriminant of (30), $\Delta_1 = B^2 - 4AC$, disappears. This condition yields a third order polynomial in k_{12}^2 , given by:

$$Dk_{12}^6 + Ek_{12}^4 + Fk_{12}^2 + G = 0 \quad (32)$$

where, $D = -512$, $E = 192k_{21}^2$, $F = -24k_{21}^2$ and $G = k_{21}^6$. The triple zero of (32) is achieved at:

$$k_{12}^2 = -\frac{E}{3D} = \frac{1}{8}k_{21}^2 \quad (33)$$

provided that the discriminant of (32) disappears. It can be shown that the discriminant of (32) is always equal to zero, regardless of the value assigned to k_{21}^2 . This infers that the triple and identical zero of the polynomial given by (32)

is always achieved and that k_{21}^2 can be used as a parameter. Substituting (33) into (31), one finds the expression for

k_{11}^2 where k_{22} and k_{21}^2 are used as parameters:

$$k_{11}^2 = \sqrt{\frac{27}{64}} \frac{Z_0 k_{21}^2}{k_{22}} \quad (34)$$

[0060] The relationship between k_{22} and the rest of impedance transformers is found from (29). Imposing that the triple zero of (27) occurs at $q \cdot Z_0$, where q is a parameter that dictates the position of the transmission zero on the resistance scale, one obtains the following relationship for k_{22} :

$$k_{22} = \frac{\sqrt{27}}{3} q \cdot Z_0 \quad (35)$$

[0061] Substituting (35) into (34), the expression for k_{11}^2 now becomes:

$$k_{11}^2 = \frac{3}{8 \cdot q} k_{21}^2 \quad (36)$$

[0062] The following conditions for the characteristic impedances, k_{12} , k_{11} and k_{22} can now be expressed as:

$$k_{12}^2 = \frac{1}{8} k_{21}^2 ; k_{11}^2 = \frac{3}{8 \cdot q} k_{21}^2 \text{ and } k_{22} = \frac{\sqrt{27}}{3} q \cdot Z_0 \quad (37)$$

where Z_0 , q and k_{21} are used as parameters. Since, Z_0 is usually, but not necessarily, 50Ω , k_{21} and q can be used in the adjustment of the rest of the impedances of the transformers, k_{12} , k_{11} and k_{22} .

[0063] Figure 14 is a schematic representation of the dependence of the characteristic impedances k_{11} , k_{12} and k_{22} on k_{21} for the case when $Z_0=50\Omega$ and $q=1$. That is, figure 14 is a graph depicting the variation of quarter-wave characteristic impedances k_{12} (triangles), k_{11} (squares), and k_{22} (circles) against k_{21} for $Z_0=50\Omega$ and $q=1$.

[0064] Even though k_{22} does not vary with k_{21} , it is displayed in figure 14 for comparison with k_{12} and k_{11} . It can be appreciated from this figure that for practical implementations using distributed quarter-wave transformers, higher values of k_{21} can be used, since this choice results in more realizable characteristic impedances for k_{12} , k_{11} and k_{22} .

[0065] Upon proper selection of k_{11} , k_{12} , k_{21} the transmission coefficient becomes:

$$S_{21} = -j\Gamma_{in} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} = \left| \frac{Z - qZ_0}{Z + qZ_0} \right|^3 \cdot e^{j3 \left(-\frac{\pi}{2} + \arctan \left(\frac{\text{Im}(Z)}{\text{real}(Z) - qZ_0} \right) - \arctan \left(\frac{\text{Im}(Z)}{\text{real}(Z) + qZ_0} \right) \right)} \quad (38)$$

EXAMPLE:

[0066] In an example in which the varactor diode has the following variation of capacitance $C_{\min}=0.4\text{pF}$ and $C_{\max}=1.6\text{pF}$, with a parasitic junction resistance of $R_p=1\Omega$, the 3-dB coupler has a rated insertion loss of 0.3 dB and its operating frequency range is 2.3-2.7 GHz the third order phase shifter can be designed by setting $k_{21}=80\Omega$, since this choice results in a more realisable values of the impedance transformers. In an example, two cases are distinguished,

one with $q=1$ and the second case where the value of q is obtained by the use of (20), to yield the value of $q=1.6$. Figures 15, 16, 17 and 18 represent the insertion loss and phase shift performance of the third order phase shifters with $q=1$ and $q=1.6$, respectively. That is, figure 15 is a graph of the insertion loss of a third order phase shifter according to an example with $q=1$ (capacitance varies from 0.4 pF to 1.6 pF in steps of 0.1 pF), figure 16 is a graph of insertion phase of third order phase shifter according to an example with $q=1$ (capacitance varies from 0.4 pF to 1.6 pF in steps of 0.1 pF), figure 17 is a graph of insertion loss of third order phase shifter according to an example with $q=1.6$ (capacitance varies from 0.4 pF to 1.6 pF in steps of 0.1 pF), and figure 18 is a graph of insertion phase of third order phase shifter according to an example with $q=1.6$ (capacitance varies from 0.4 pF to 1.6 pF in steps of 0.1 pF).

[0067] As evident from figure 16, the phase shift obtained from the third order circuit with $q=1$ is approximately three times greater than that achieved using the first order circuit, however, the insertion loss is not tripled, but increased by a maximum of 0.4 dB. The third order phase shift realization with $q=1.6$ offers an even greater phase shift than its third order counterpart with $q=1$. To be precise, the phase shift of the second order circuit with $q=1.6$ is increased by an average of 20° across the indicated frequency range compared to the phase shift of the second order circuit with $q=1$. The insertion loss performance of the second order phase shifter with $q=1.6$ is also superior compared to its third order counterpart with $q=1$, exhibiting a 0.1 dB smaller insertion loss.

Claims

1. An RF reflection type phase shifter (RTPS) comprising:

a coupling device coupled to multiple reflective loads, n , respective ones of which include at least one variable reactance device with maximum reactance, $X_{\max}$, and, minimum reactance, $X_{\min}$, and at least two impedance transformers the characteristic impedances of which are selected according to predefined criteria to provide an increase in the value of a phase shift to be applied to a signal input to the RTPS proportional to the value of n .

2. An RTPS as claimed in claim 1, wherein the characteristic impedances are selected in accordance with a selected value of a parameter, q , determined, for a given reflective load, according to:

$$q = \frac{\sqrt{X_{\max} X_{\min}}}{Z_0}$$

where Z_0 is the characteristic impedance of the coupling device and interconnecting microstrip lines.

3. An RTPS as claimed in claim 1 or 2, wherein the impedance of the or each variable reactance device is altered by varying a DC voltage or current applied to the variable reactance device.

4. An RTPS as claimed in any preceding claim, wherein the or each variable reactance device is a varactor diode.

5. An RTPS as claimed in any preceding claim, wherein the impedance transformers are one quarter-wavelength long at a selected frequency of operation.

6. An RTPS as claimed in any preceding claim, where the impedance transformers use microstrips or stripline technology.

7. An RTPS as claimed in any preceding claim, wherein the impedance transformers use lumped circuit elements.

8. An RTPS as claimed in any preceding claim, wherein the coupling device is a circulator or a 3-dB coupler.

9. A method for selecting a phase shift value for an output signal of a phase shifter including a coupling device coupled to multiple reflective loads, n , respective ones of which include at least one variable reactance device with maximum reactance, $X_{\max}$, and, minimum reactance, $X_{\min}$, and at least two impedance transformers, the method comprising:

selecting the characteristic impedances of impedance transformers according to predefined criteria to provide an increase in the value of a phase shift to be applied to a signal input to the phase shifter proportional to the

value of n.

10. A method as claimed in claim 9, further including:

5 selecting the characteristic impedances in accordance with a selected value of a parameter, q, determined, for a given reflective load, according to:

$$q = \frac{\sqrt{X_{\max} X_{\min}}}{Z_0}$$

15 where Z_0 is the characteristic impedance of the coupling device and interconnecting microstrip lines.

11. A method as claimed in claim 9 or 10, further including:

20 modifying the impedance of the or each variable reactance device by varying a DC voltage or current applied to the variable reactance device.

12. A beamformer for use in a wireless telecommunication network to modify the transmission profile of an antenna array, the beamformer including an RF reflection type phase shifter (RTPS) comprising a coupling device coupled to multiple reflective loads, n, respective ones of which include at least one variable reactance device with maximum reactance, $X_{\max}$, and, minimum reactance, $X_{\min}$, and at least two impedance transformers the characteristic impedances of which are selected according to predefined criteria to provide an increase in the value of a phase shift to be applied to a signal input to the RTPS proportional to the value of n.

13. A beamformer as claimed in claim 12, wherein the antenna array is composed of multiple antennas, wherein the phase shifter is operable to modify the phase of respective signals input to the multiple antennas.

14. A beamformer as claimed in claim 13, wherein the phase shifter comprises multiple parallel phase shift stages to apply respective different phase shifts to the respective signals input to the multiple antennas.

15. An antenna array in a wireless telecommunication network, the antenna array including multiple antennas operable to receive input from a beamformer that is operable to modify the transmission profile of the antenna array, the beamformer including an RF reflection type phase shifter (RTPS) comprising a coupling device coupled to multiple reflective loads, n, respective ones of which include at least one variable reactance device with maximum reactance, $X_{\max}$, and, minimum reactance, $X_{\min}$, and at least two impedance transformers the characteristic impedances of which are selected according to predefined criteria to provide an increase in the value of a phase shift to be applied to a signal input to the RTPS proportional to the value of n.

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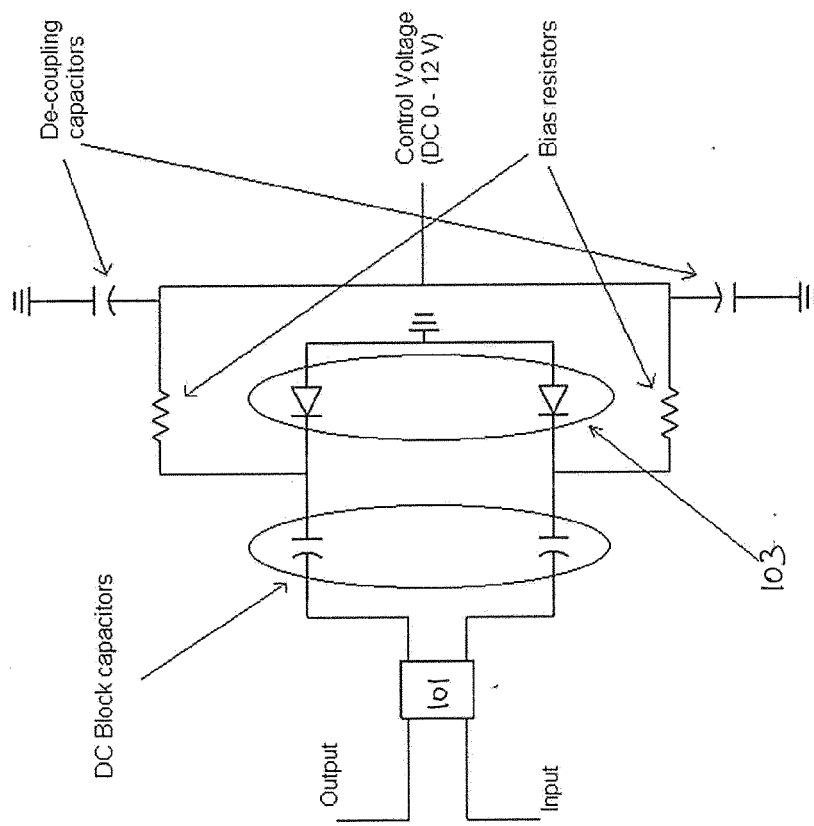


Figure 1

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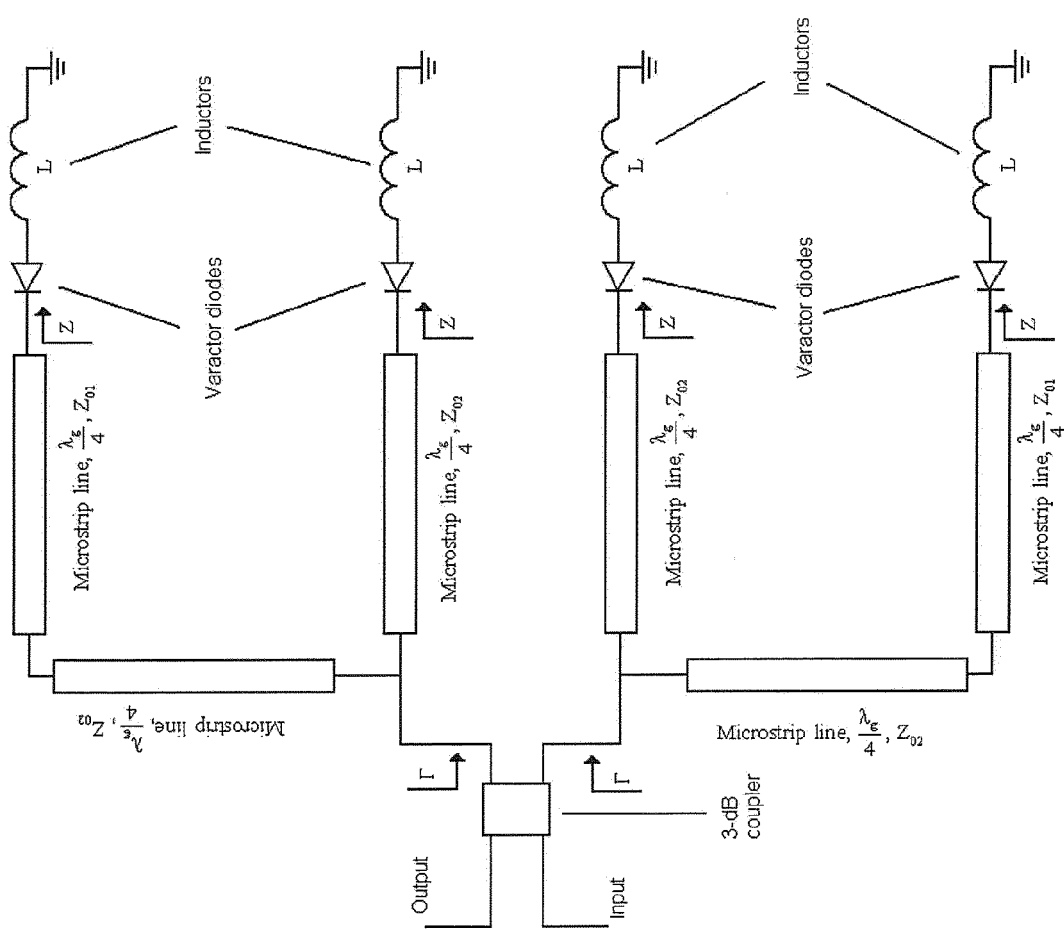


Figure 2

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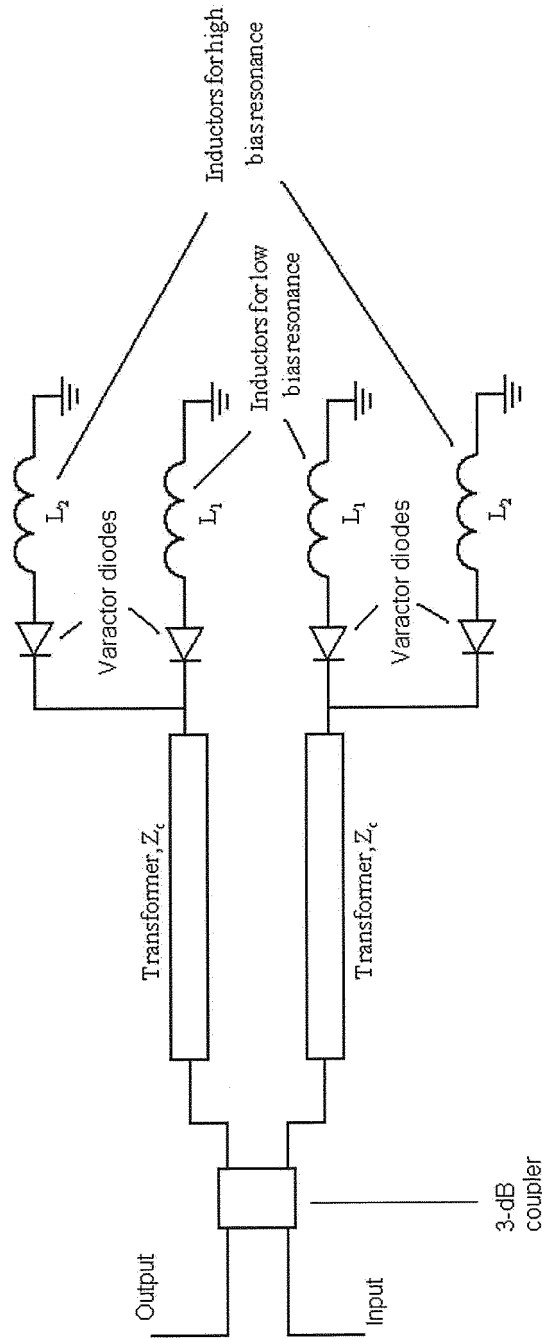


Figure 3

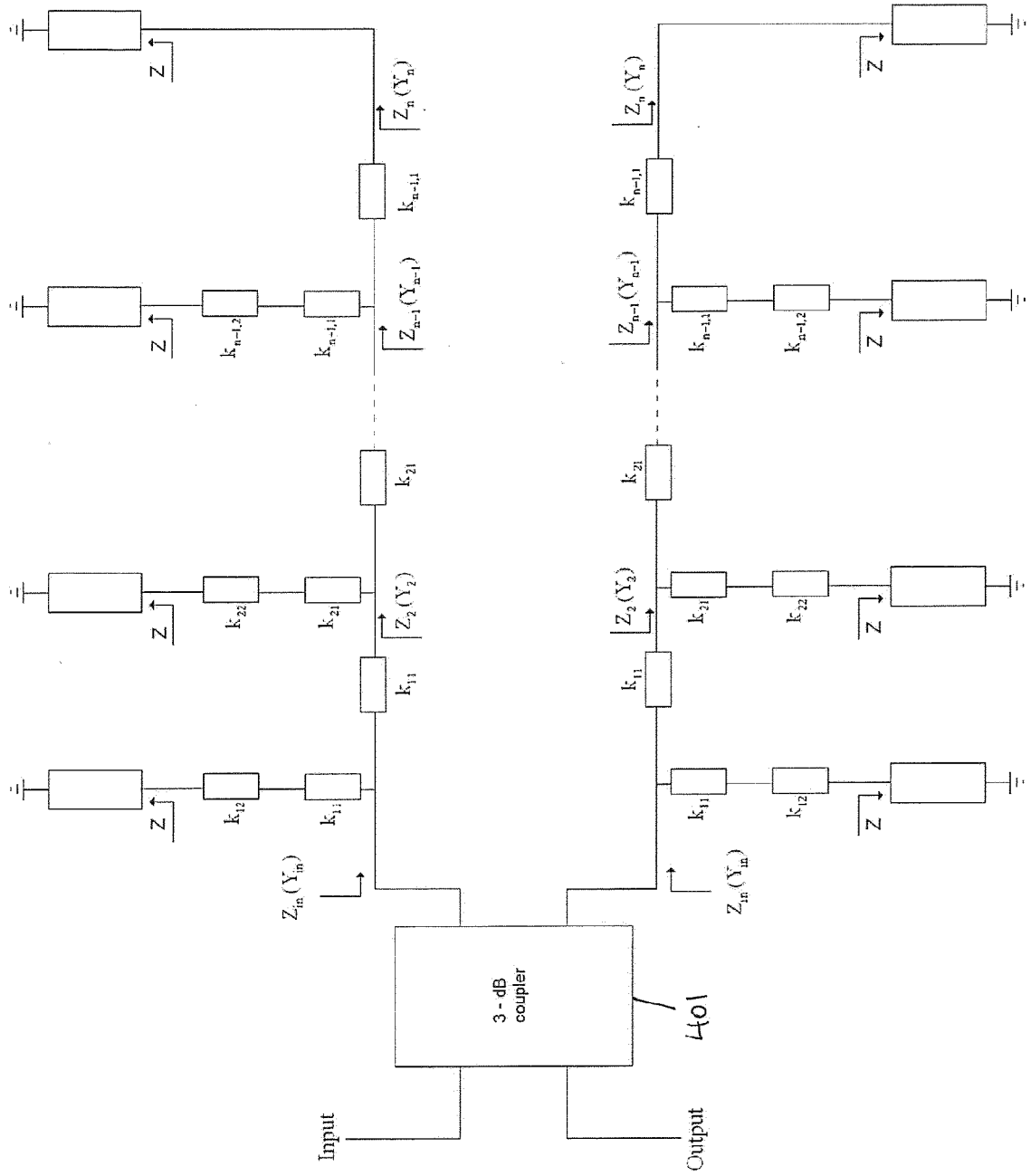


Figure 4

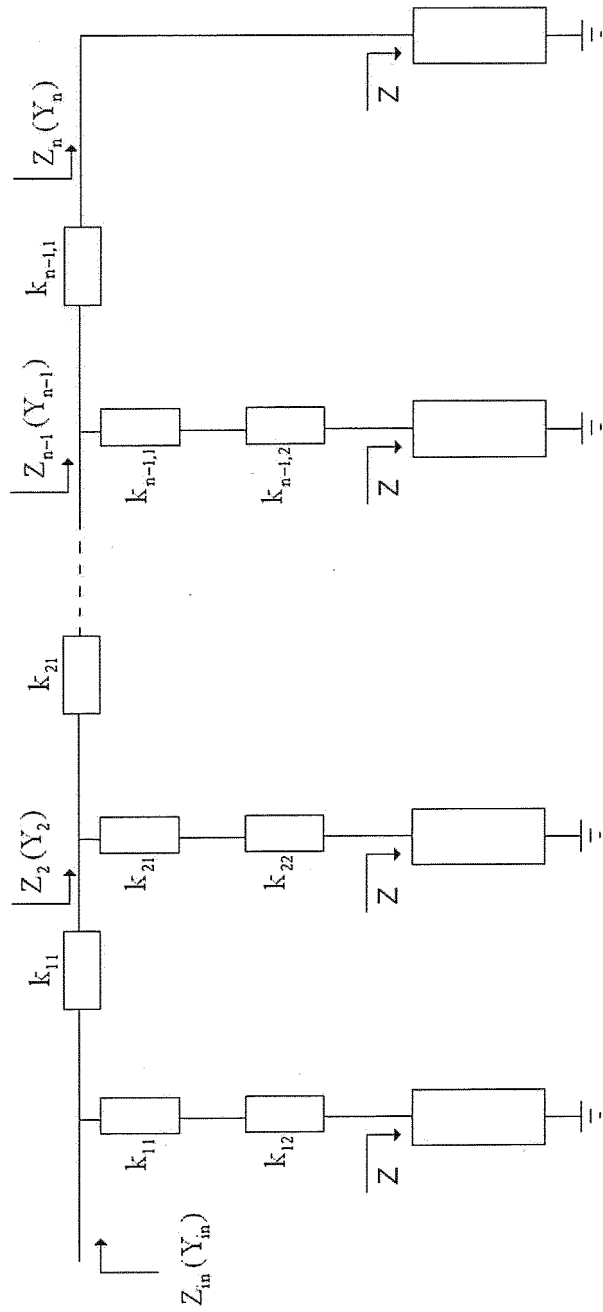


Figure 5

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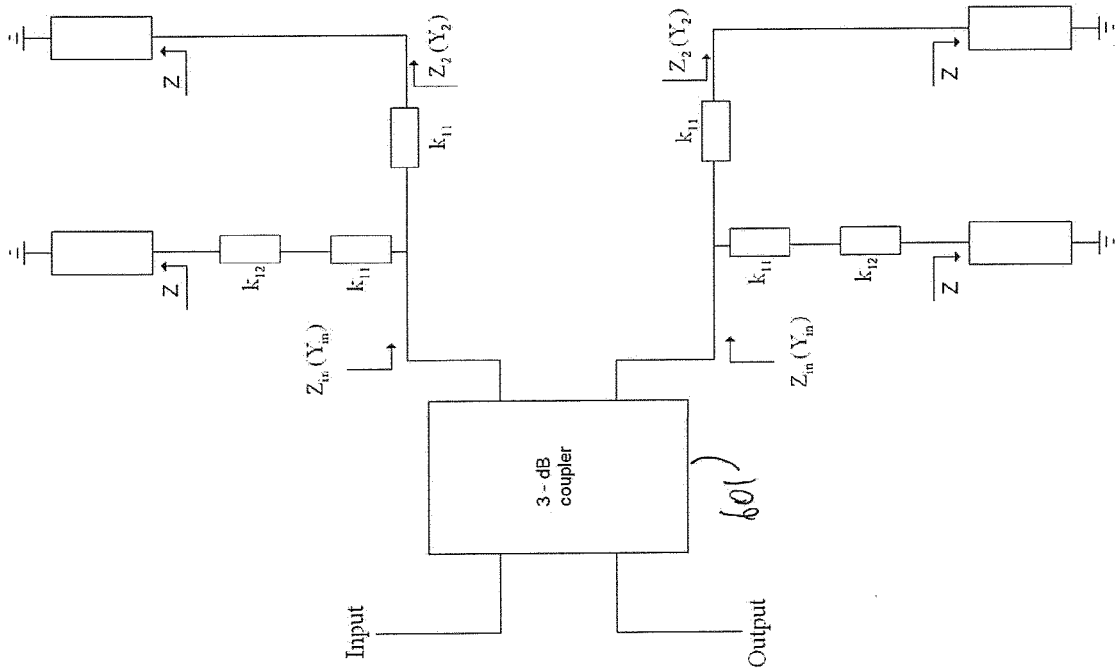


Figure 6

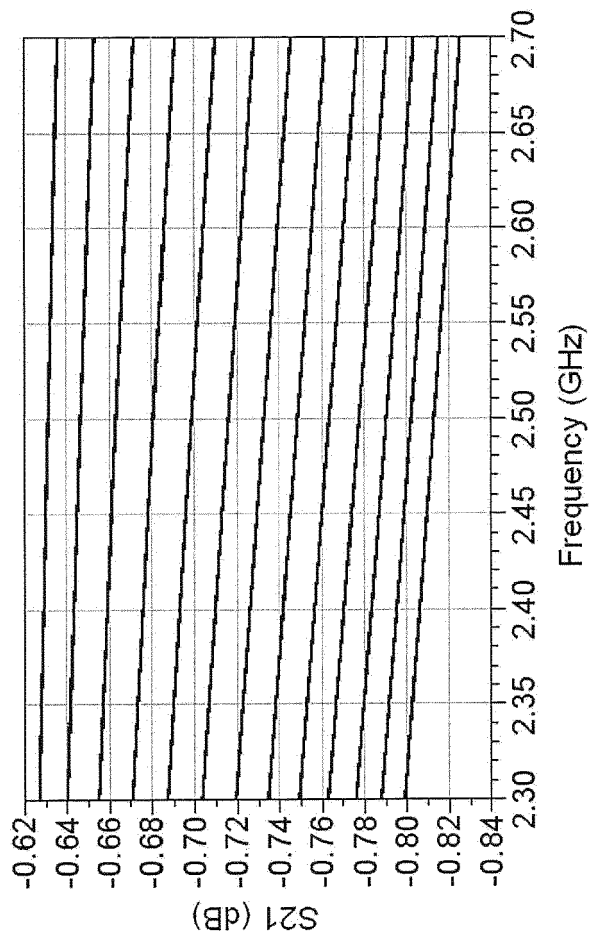


Figure 7

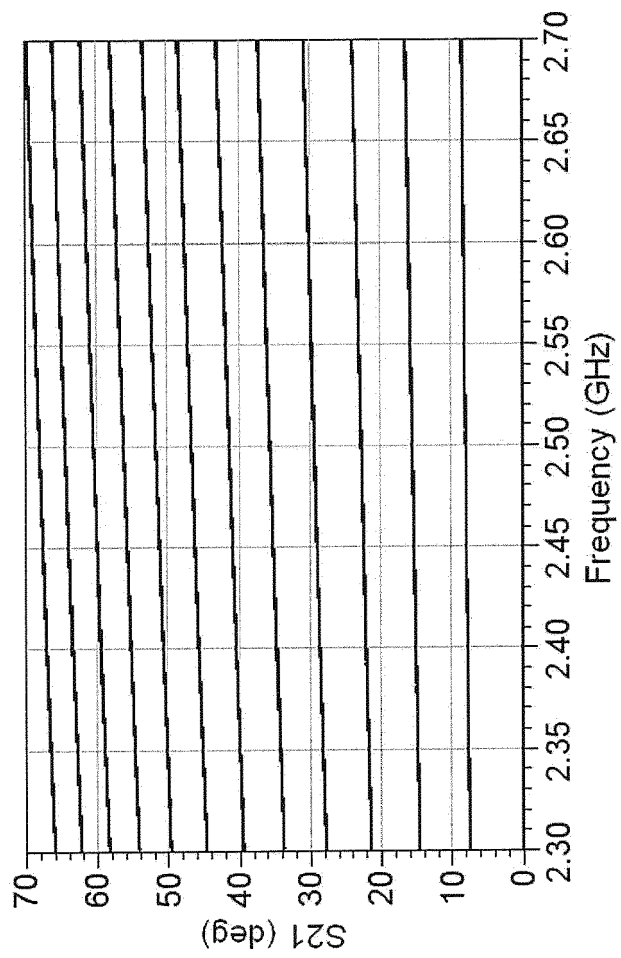


Figure 8

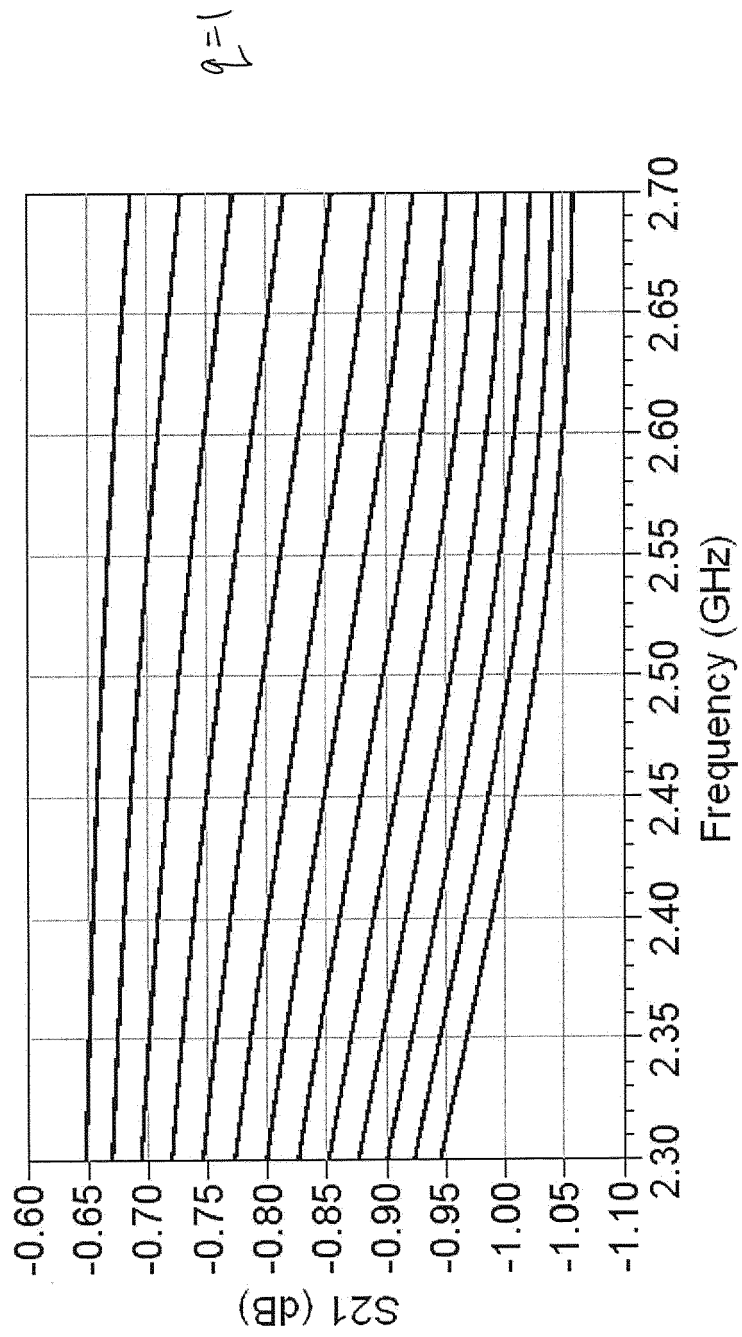


Figure 9

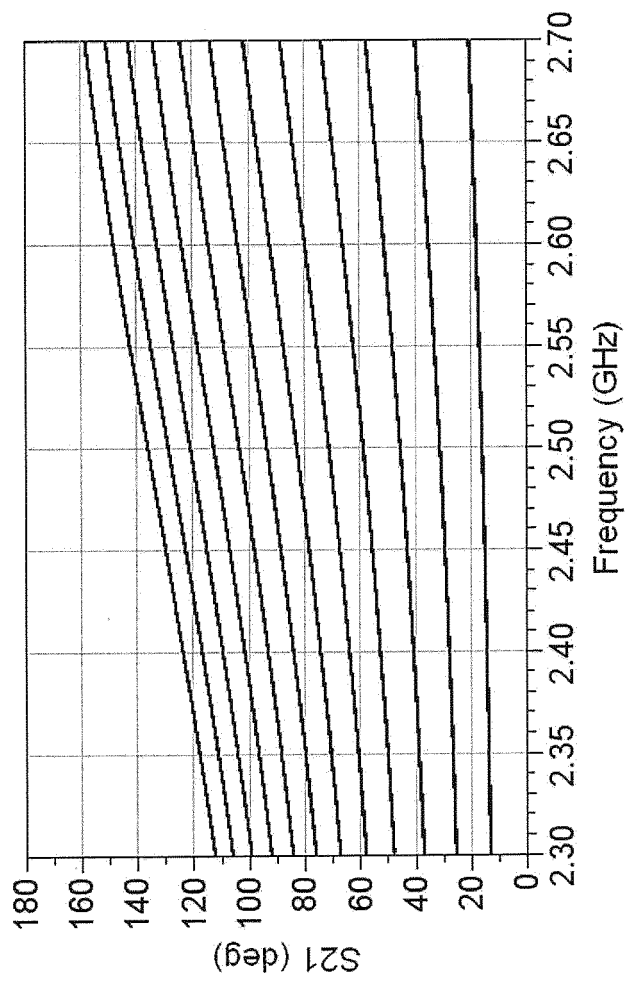


Figure 10

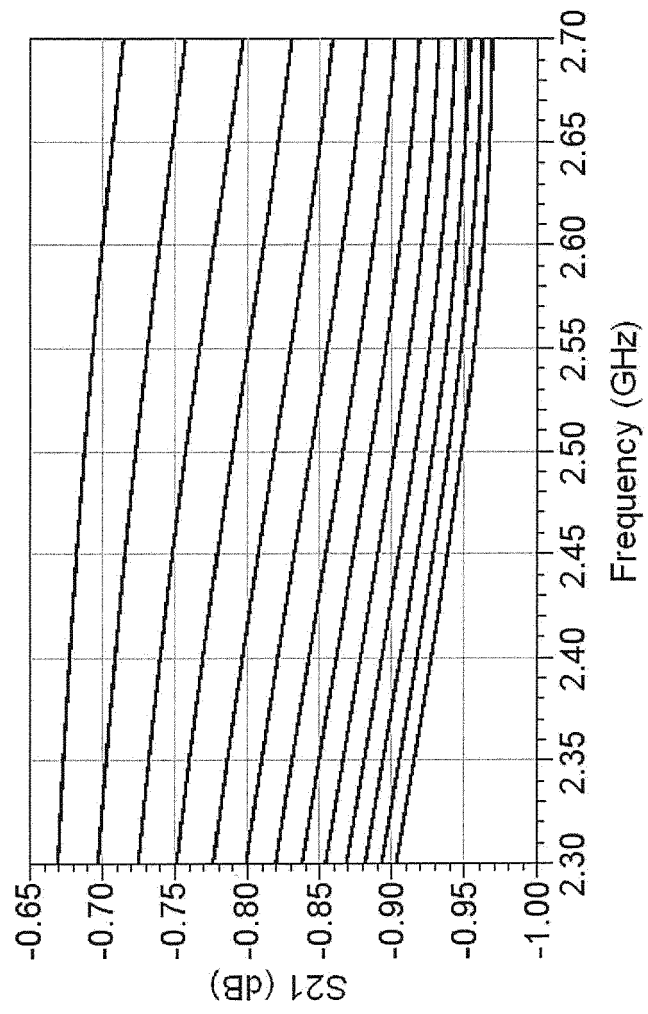
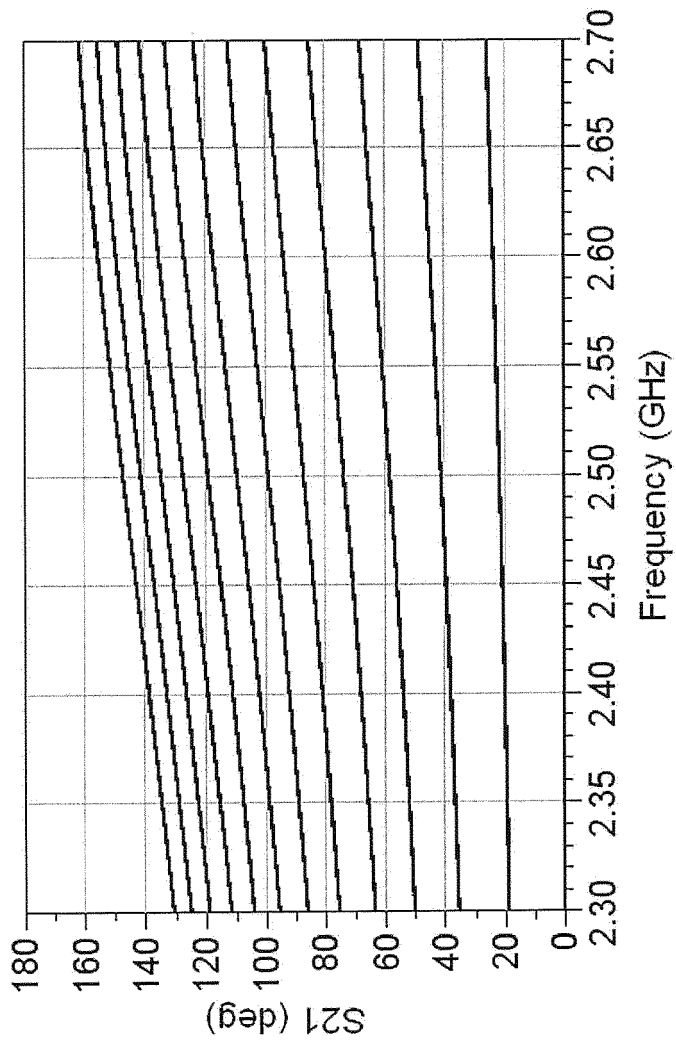


Figure 11



$q=1.6$

Figure 12

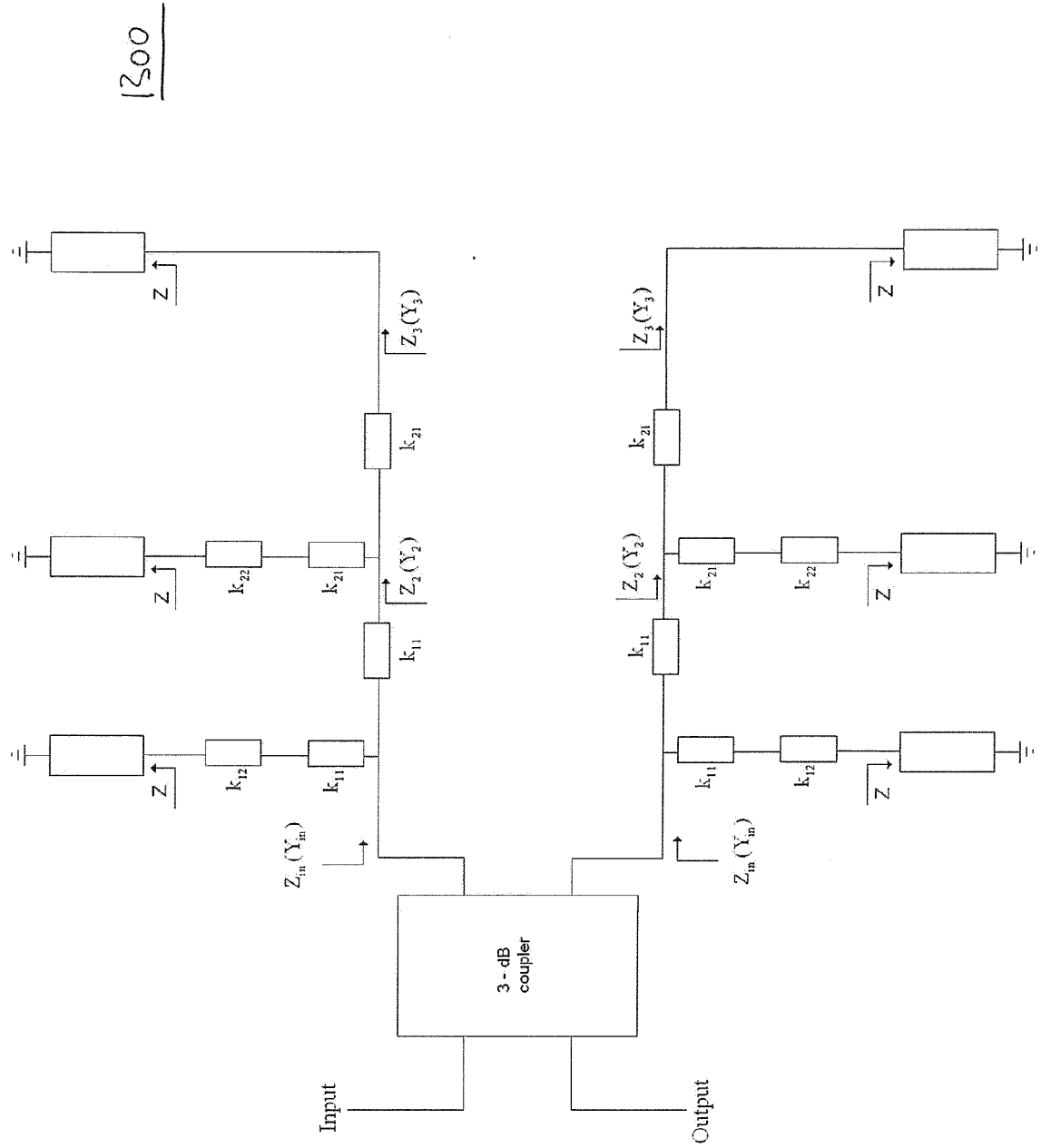


Figure 13

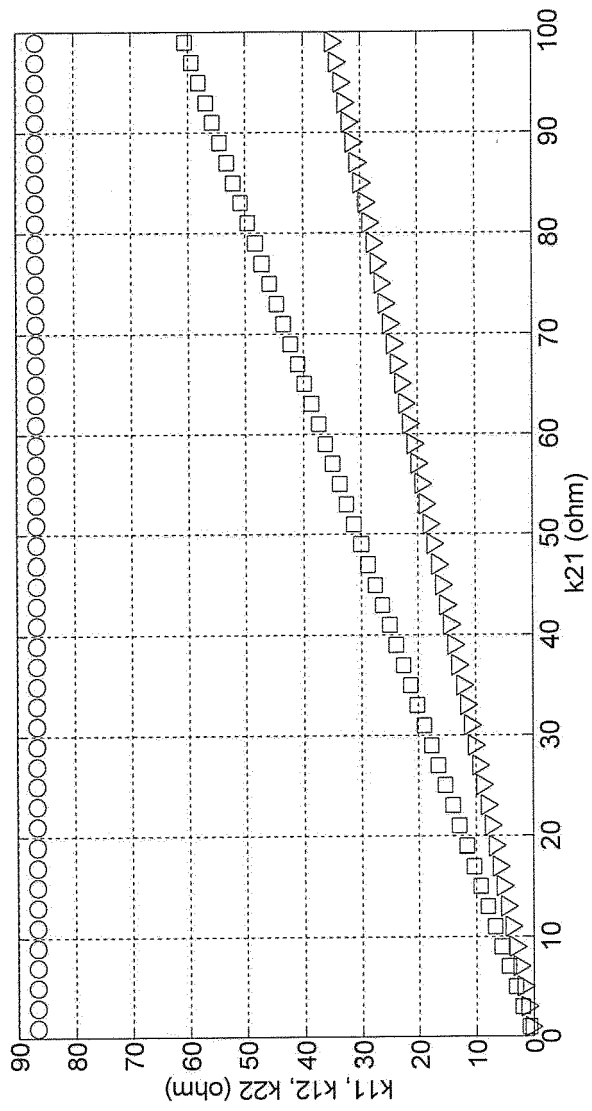


Figure 14

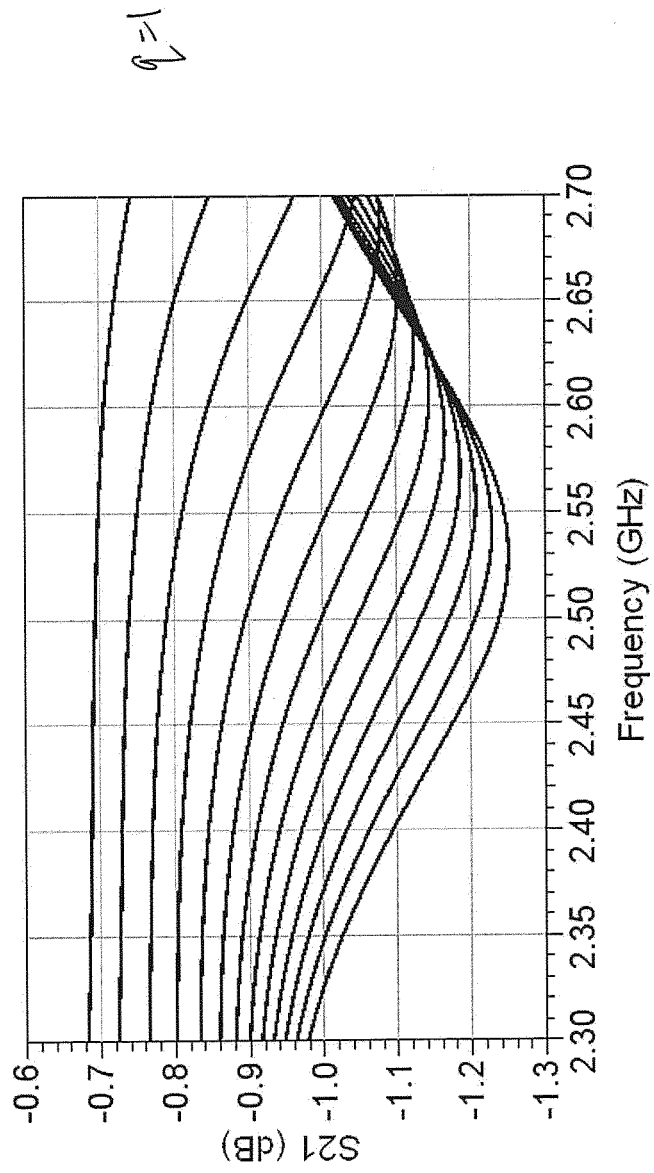


Figure 15

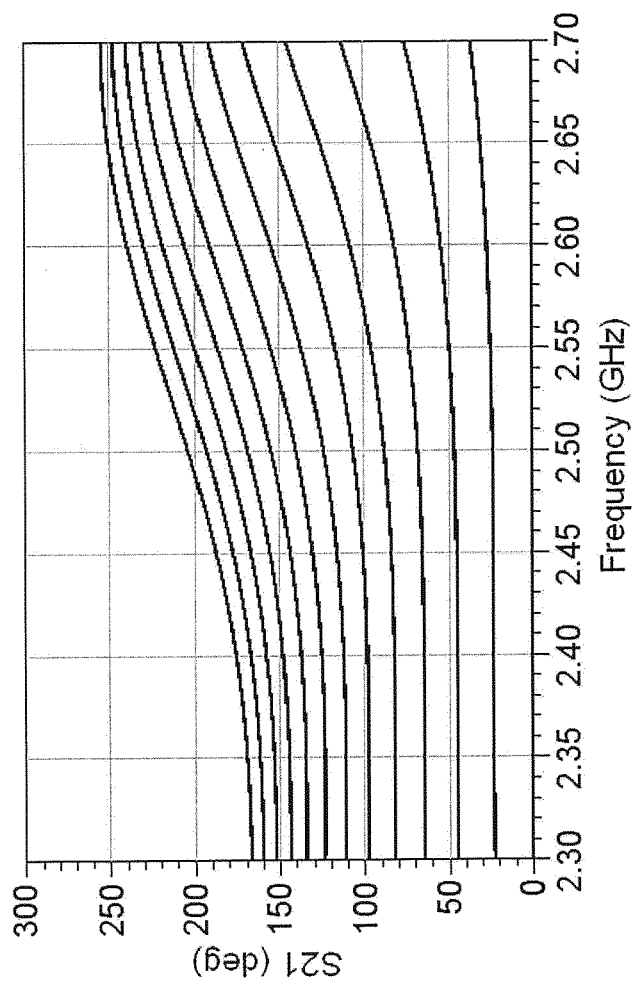


Figure 16

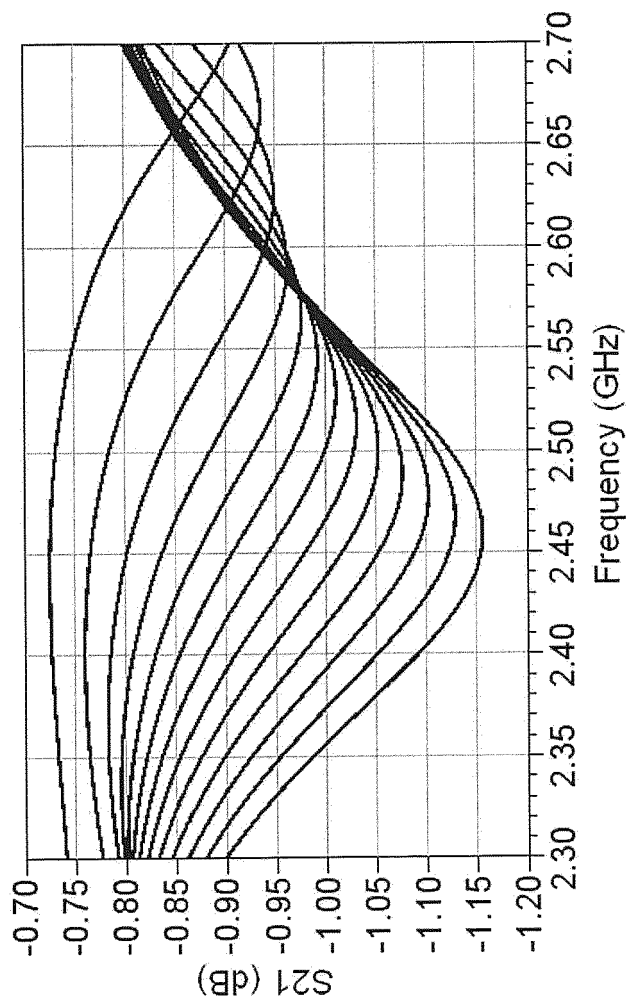


Figure 17

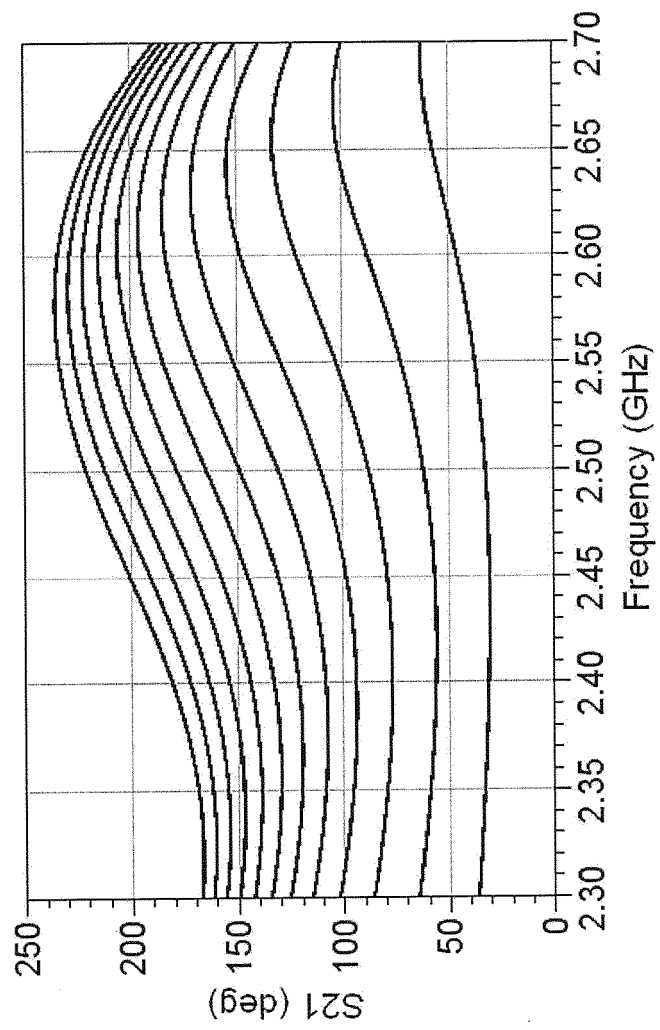


Figure 18



EUROPEAN SEARCH REPORT

Application Number
EP 14 30 6380

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X	EP 2 544 369 A1 (ALCATEL LUCENT [FR]) 9 January 2013 (2013-01-09) * paragraph [0047] - paragraph [0054]; figure 2 * * paragraph [0070] - paragraph [0078]; figure 6 *	1-15	
X	WON-TAE KANG ET AL: "Reflection-Type Low-Phase-Shift Attenuator", IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES, IEEE SERVICE CENTER, PISCATAWAY, NJ, US, vol. 46, no. 7, 1 July 1998 (1998-07-01), XP011037211, ISSN: 0018-9480 * the whole document *	1,9,12,15	
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X	US 5 276 411 A (WOODIN JR CHARLES E [US] ET AL) 4 January 1994 (1994-01-04) * column 2, line 58 - line 68; figures 1,2 * * column 3, line 56 - column 4, line 23; figures 3,4 *	1,9,12,15 2-8,10,11,13,14	
A			
The present search report has been drawn up for all claims			
Place of search The Hague		Date of completion of the search 13 February 2015	Examiner Pastor Jiménez, J
CATEGORY OF CITED DOCUMENTS X : particularly relevant if taken alone Y : particularly relevant if combined with another document of the same category A : technological background O : non-written disclosure P : intermediate document		T : theory or principle underlying the invention E : earlier patent document, but published on, or after the filing date D : document cited in the application L : document cited for other reasons & : member of the same patent family, corresponding document	

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			TECHNICAL FIELDS SEARCHED (IPC)
The present search report has been drawn up for all claims			
Place of search The Hague		Date of completion of the search 13 February 2015	Examiner Pastor Jiménez, J
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13-02-2015

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For more details about this annex : see Official Journal of the European Patent Office, No. 12/82