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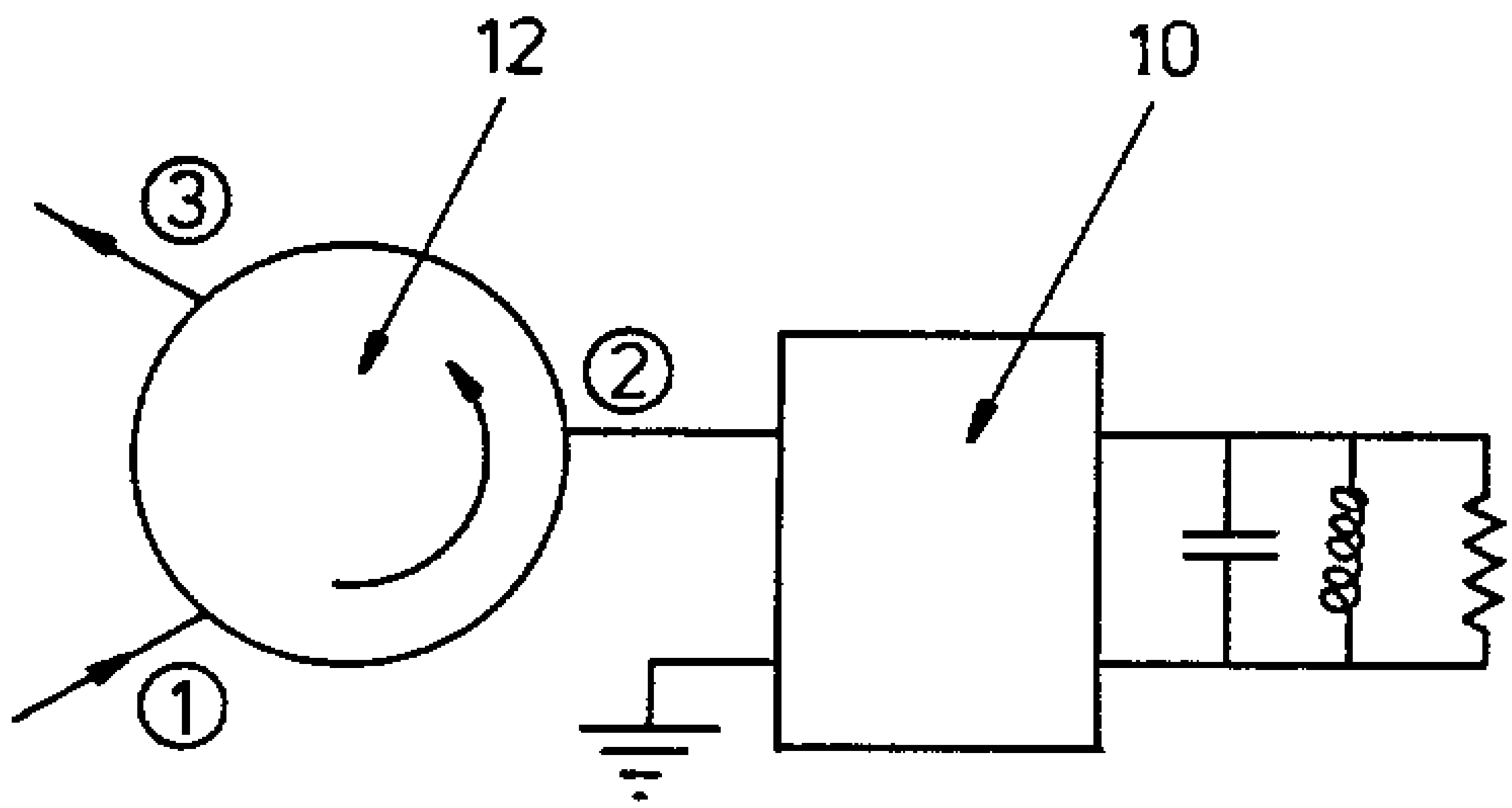

(57) Abrégé/Abstract:
A variable Q reflection mode filter having a three-port circulator device one port of which is terminated by a one-port filter. The filter increases the overall 3dB resonance point of the filter to reduce the unloaded Q factors required for particular applications. The filter may be arranged to provide a maximally flat response or a quasi equiripple response. Transmission zeros may be incorporated, to further enhance the effective unloaded Q.

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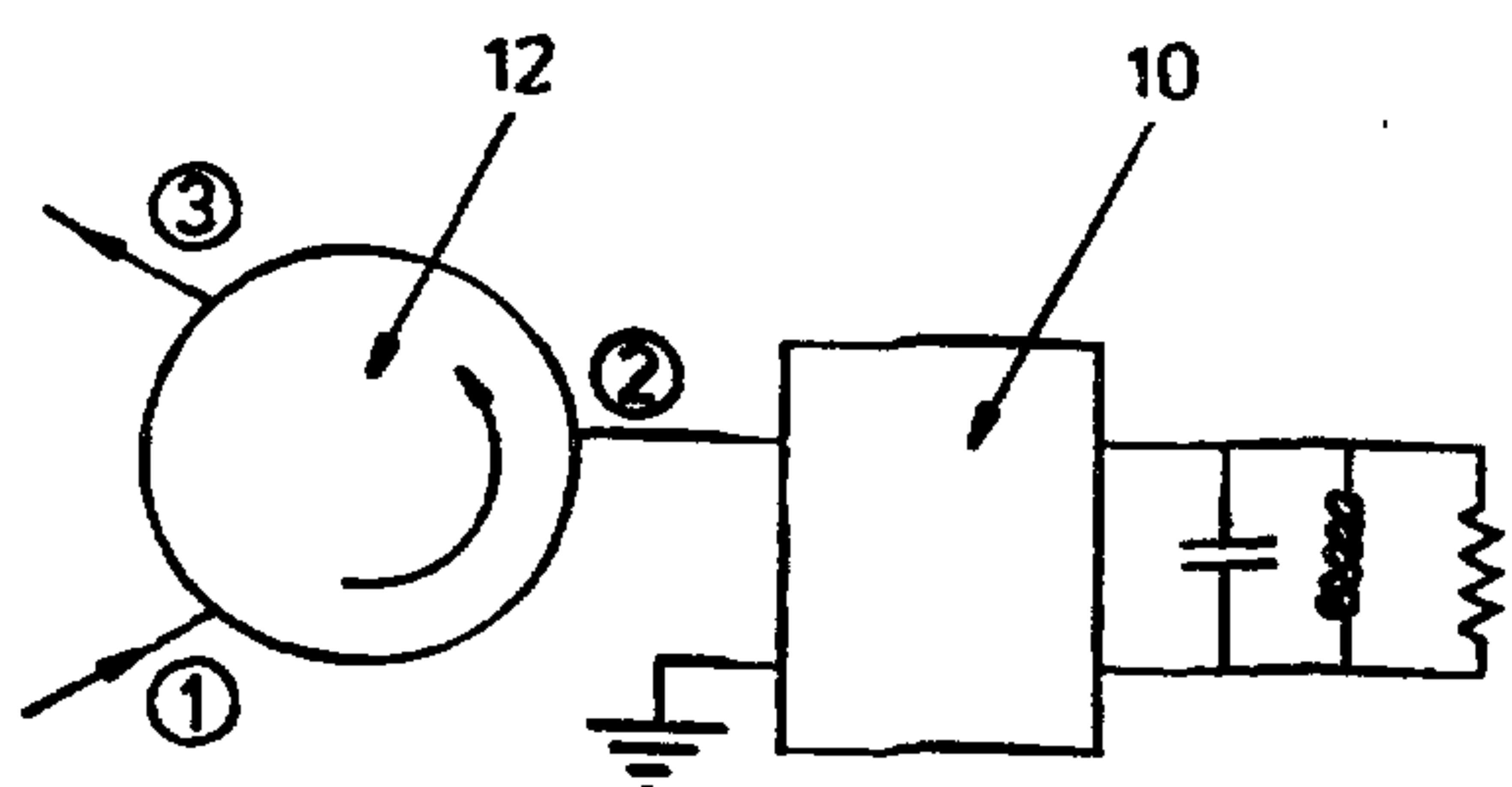
(57) Abstract

A variable Q reflection mode filter having a three-port circulator device one port of which is terminated by a one-port filter. The filter increases the overall 3dB resonance point of the filter to reduce the unloaded Q factors required for particular applications. The filter may be arranged to provide a maximally flat response or a quasi equiripple response. Transmission zeros may be incorporated, to further enhance the effective unloaded Q.



Microwave Filter

This invention relates to a microwave filter and more particularly to a variable Q reflection mode microwave filter.

All passive resonators have a finite unloaded Q factor. In narrow bandwidth applications, this resistive loss can lead to difficulties in the design process. In a bandpass application, designs which provide for both a good input and output match exhibit transfer characteristics with significant amplitude variation over the passband if mid band loss is minimised. This passband variation can only be reduced with given Q factors if the mid band loss is increased, possibly to an unacceptable level. Even in the case of a single resonator, filter problems occur due to the resistive loss which prevents a good input and output match being simultaneously achievable.

In the case of a narrow band bandstop application, the resistive loss of the resonators causes a roll-off of the insertion loss into the passband. A reduction in unloaded Q can quickly cause this loss to reach an unacceptable level, particularly where noise figure is important and the notch filter has been introduced to reject signals which would limit the dynamic range of the receiver. This requirement now exists in several countries where cellular telephone systems have multi-operator configurations.

In a conventional bandstop filter, resonators are coupled off from a main through-line with an electrical separation of an odd number of 90°. Each resonator couples loss into the system and this can be further increased by additional loss in the through-line. To meet a typical requirement at 900MHz, at least 20dB rejection has to be provided over a band in excess of 1MHz, whilst the loss at the 1.5MHz bandwidth is less than 2dB. To achieve this, unloaded Q's of greater than 20,000 are required, resulting in the necessity to use dielectric resonators for all of the cavities.

To meet rejection levels of 20 to 30dB, an alternative method may be used based upon the use of a bandpass filter connected to a 3dB hybrid, as set out in UK patent application No. 9324149.5 (2,284,311): this "hybrid notch filter" is more compact and provides a slightly lower loss, but still needs unloaded Q's of the order of 20,000 for all cavities. However, we have now devised a filter which acts as a true reflection mode filter and meets the above requirements with much lower unloaded Q's.

In accordance with this invention, there is provided a microwave reflection mode filter which comprises a three-port circulator device having one port terminated by a one-port filter.

The filter may be arranged, as explained herein, to provide a maximally flat response. Instead, the filter may be arranged, also as explained herein, to provide an equiripple (or quasi equiripple) response.

It will be appreciated that low loss circulators, suitable for use in the filter of this invention, are readily available on the market.

Embodiments of this invention will now be described by way of examples only and with reference to the accompanying drawings, in which:

FIGURE 1 is a schematic diagram of a filter in accordance with this invention;

FIGURE 2 is a diagram for use in explaining principles on which the filter of Figure 1 is based;

FIGURE 3 is a schematic diagram of a lossy ladder network for use in the filter of Figure 2;

FIGURE 4 shows the theoretical quasi equiripple response (with 20dB stopband level) of a filter having a 7th degree ladder network;

FIGURE 5 is a diagram of a ladder network for a filter having a quasi equiripple response;

FIGURE 6 shows the measured response of a 5th degree filter in accordance with the invention;

FIGURE 7 shows the measured response of a 6th degree filter in accordance with the invention;

FIGURE 8 shows the measured response of a 6th degree filter which includes cross-coupling; and

FIGURE 9 shows the passband loss of the 6th degree filter with cross-coupling.

Referring to Figure 1 of the drawings, there is diagrammatically shown a filter which comprises a resonant circuit 10 with loss, coupled to one of the ports of a circulator 12. The transmission characteristic from ports 1 to 3 of the circulator is the reflection characteristic from the network 10 connected to port 2. Assume that the coupling into the resonant circuit is adjusted such that the resistive part at resonance is matched to the impedance of the circulator, then at resonance all of the power supplied at port 1 emerges at port 2 and is absorbed in the resistive part of the resonator. Hence, there is no transmission to port 3. In this case the transmission characteristic from ports 1 to 3 of the circulator is of a single resonator with an infinite rejection at centre frequency, i.e. as if the resonance were from a resonator of infinite unloaded Q. If fo is the centre frequency and B the 3dB bandwidth of the resonance, then by a simple calculation the unload Q (Qu) of the resonator is given by:

$$Qu=\frac{2fo}{B} \qquad (1)$$

For example if B = 250KHz and fo = 1GHz, then Qu = 8,000. This shows that the type of specification previously considered can be met with cavities of much lower Qu if a design procedure is established for a multi-element filter.

The following discussion concerns the design for a variable Q reflection mode filter and in particular designs which provide a maximally flat response and a quasi equiripple response. The discussion provides solutions to the approximation problem and subsequent investigation of the synthesis leads to explicit design formulas for filters of arbitrary degree. Two examples of 5th and 6th degree filters will be given designed around the previous specification and the measured results are shown to be in good agreement with theory. Furthermore, it will be shown that by adding transmission zeros into the reflection mode filter, further enhancement in effective unloaded Q can be obtained: a 6th degree device will be discussed to demonstrate this point.

Firstly a reflection mode filter with a Maximally Flat response will be discussed with reference to Figure 2. Using a circulator, the reflection coefficient $S_{11}(p)$ of a reflection mode filter becomes the transmission coefficient of the overall device shown in Figure 2. Assuming that the circulator is normalised to 1Ω impedance, then in lowpass prototype form as shown in Figure 3 a Maximally Flat response will be achieved if:

$$|S_{11}(j\omega)|^2 = \frac{\omega^{2^n}}{D_n(\omega^2)} \qquad (2)$$

where n is the degree of the filter and $D_n(\omega^2)$ is a polynomial of degree n in ω^2 with:

$$\omega^{2^n} \leq D_n(\omega^2) \qquad (3)$$

For the normal maximally flat response:

$$|S_{11}(j\omega)|^2 = \frac{\omega^{2^n}}{1+\omega^{2^n}} \qquad (4)$$

and the resulting network is a lowpass ladder network with infinite Qu terminated in a 1Ω resistor.

Initially consider the case where each resonator will have the same Qu. Assume that the ratio of conductance to capacitance of each shunt element is normalised to ½ and using the transformation

$$z=2p+1 \tag{5}$$

then the input admittance of the network may be expressed as a reactance function Y(z). Hence:

$$S_{11}(p)=\frac{1-Y(z)}{1+Y(z)} \tag{6}$$

where p is the complex frequency variable and for a maximally flat response around p = 0:

$$\begin{aligned} S_{11}(p)&=\left(\frac{1-z}{1+z}\right)^{n} \\ &=\left[\frac{-p}{1+p}\right]^{n} \end{aligned} \tag{7}$$

where:

$$\begin{aligned} Y(z)&=\frac{(1+z)^{n}-(1-z)^{n}}{(1+z)^{n}+(1-z)^{n}} \\ &=\tanh[n\tanh^{-1}(z)] \end{aligned} \tag{8}$$

Now:

$$|S_{11}(j\omega)|^{2}=\frac{\omega^{2^{n}}}{(1+\omega^{2})^{n}} \tag{9}$$

and let the return loss $L_R = L_1$ (≈20dB) for $\omega<\omega_s$ and $L_R = L_2$ (≈ 1dB) for $\omega > \omega_p$, then:

$$10\log\left[1+\frac{1}{\omega_s^{2}}\right]^{n}=L_1 \tag{10}$$

and

$$10\log[1+\frac{1}{\omega_p^2}]^n = L_2 \qquad \textbf{(11)}$$

Therefore:

$$\begin{aligned}\frac{1}{\omega_s^2} &= 10^{\frac{L_1}{10n}}-1\\ \frac{1}{\omega_p^2} &= 10^{\frac{L_2}{10n}}-1\end{aligned} \qquad \textbf{(12)}$$

and for n large:

$$\frac{\omega_p}{\omega_s} \rightarrow \frac{L_1}{L_2} \qquad \textbf{(13)}$$

and hence there is a fundamental limit to the ratio of the passband frequency to the stopband frequency.

To achieve the objective, a response is required between the response in equation (4) and that in equation (9). One such response is:

$$\begin{aligned}|S_{11}(j\omega)|^2 &= \sum_{r=0}^{n} \frac{\omega^{2^n}}{\omega^{2^r}}\\ &= \frac{(1-\omega^2)\,\omega^{2^n}}{(1-\omega^{2^{n+2}})}\end{aligned} \qquad \textbf{(14)}$$

In this case:

$$10\log\left|\frac{1-\frac{1}{\omega^{2n+2}}}{1-\frac{1}{\omega_S^2}}\right| = L_1 \qquad \textbf{(15)}$$

and

$$\frac{\omega_p}{\omega_s} = \frac{10^{\frac{L1}{20n}}}{\sqrt{1-10^{\frac{-L_2}{10}}}} \qquad (16)$$

For L_1 = 20dB, L_2 = 1dB and n = 7:

$$\frac{\omega_p}{\omega_s} = 3.06$$

as compared to 1.53 from the normal maximally flat response given in equation (4).

From equation (14) the following bounded real reflection coefficient may be formed:

$$S_{11}(p) = \prod_{r=1}^{n} \frac{p^n}{(p - je^{j\theta r})} \qquad (17)$$

with

$$\theta_r = \frac{r\pi}{n+1} \qquad (18)$$

Forming the input admittance

$$Y(p) = \frac{1-S_{11}(p)}{1+S_{11}(p)} \qquad (19)$$

and synthesising as the lossy ladder network shown in Fig 3, explicit formulas are obtained for element values given by:

$$K_{r-1,r} = 1 \quad r = 1 \rightarrow n \qquad (20)$$

and if

$$10\log\left[\frac{1-\frac{1}{\omega_p^{2n+2}}}{1-\frac{1}{\omega_p^2}}\right]=L_2$$

hence:

$$\frac{1}{\omega_s^2}\approx 10^{\frac{L^1}{10n}}$$

$$\frac{1}{\omega_p^2}\approx 1-10^{\frac{-L_2}{10}}$$

and

$$\frac{\omega_p}{\omega_s}=\frac{10^{\frac{L1}{20n}}}{\sqrt{1-10^{\frac{-L_2}{10}}}} \qquad (16)$$

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where θ_r is a constant applicable to the r^{th} resonator. Forming the input admittance

$$Y(p) = \frac{1-S_{11}(p)}{1+S_{11}(p)} \qquad (19)$$

and synthesising as the lossy ladder network shown in Fig 3, explicit formulas are obtained for element values given by:

$$K_{r-1,r}=1 \quad r=1 \to n \qquad (20)$$

where $K_{r-1,r}$ is the admittance of the invertor coupling between the $(r-1)^{th}$ and r^{th} resonator, and if:

$$E_o=1$$

$$E_{r-1}E_r = \left[\frac{\cos\theta+\cos(r\theta)}{\cos\theta+\cos((r-1)\theta)}\right] \quad r=1 \to n \qquad (21)$$

with

$$\theta = \frac{\pi}{n+1}$$

then:

$$c_r = \frac{1}{\cos\theta}\left[\frac{\sin((r-1)\theta)}{E_{r-1}} + \sin(r\theta)E_r\right]$$

where E_r and E_{r-1} are intermediate variables and C_r is the capacitance of the r^{th} resonator, and

$$G_r = \frac{1}{E_{r-1}} - E_r \tag{22}$$

where G_r is the conductance of the r^{th} resonator.

It can also be shown that:

$$c_r > 0, G_r > 0 \quad r = 1 \rightarrow n$$

To illustrate the use of these explicit formulas, consider the case of n=5. Then:

$$\theta = \frac{\Pi}{6}, \cos\theta = \frac{\sqrt{3}}{2}$$

$$E_o = 1$$

$$E_o E_1 = \frac{\sqrt{3}}{1 + \frac{\sqrt{3}}{2}}, E_1 = 2\sqrt{3}(2 - \sqrt{3})$$

$$E_1 E_2 = \frac{\sqrt{3}+1}{2\sqrt{3}}, E_2 = \frac{5+3\sqrt{3}}{12} \tag{24}$$

$$E_2 E_3 = \frac{\sqrt{3}}{\sqrt{3}+1}, E_3 = 6\sqrt{3}(7 - 4\sqrt{3})$$

$$E_3 E_4 = \frac{\sqrt{3}-1}{\sqrt{3}}, E_4 = \frac{5+3\sqrt{3}}{18}$$

$$E_4 E_5 = 0, E_5 = 0$$

Therefore:

$$C_1=2(2-\sqrt{3}),\ G_1=7-4\sqrt{3}$$

$$C_2=\frac{(9+5\sqrt{3})}{12},\ G_2=\frac{1+\sqrt{3}}{12}$$

$$C_3=6(9-5\sqrt{3}),\ G_3=6(7-4\sqrt{3})$$

$$C_4=\frac{19+11\sqrt{3}}{18},\ G_4=\frac{7+4\sqrt{3}}{18}$$

$$C_5=9(3\sqrt{3}-5),\ G_5=9(3\sqrt{3}-5)$$

Also, defining the normalised Q of each resonator as

$$Q_r=\frac{C_r}{G_r} \tag{26}$$

then from equations (21) and (22):

$$Q_r=\frac{\cos(\frac{\theta}{2})+\cos(r-\frac{1}{2}\theta)}{\sin(\frac{\theta}{2})} \tag{27}$$

$$r=1\rightarrow n$$

or

$$Q_r=\frac{2\cos\frac{(r\theta)}{2}\cos\frac{(r-1)\theta}{2}}{\sin\frac{\theta}{2}} \tag{28}$$

$$r=1\rightarrow n$$

which directly assists in determining the relative Q's of each cavity.

Consideration will now be given to a quasi equiripple reflection mode filter. Extending the normally maximally flat response to the equiripple case gives:

$$|S_{11}(j\omega)|^2=\frac{\varepsilon^2 T_n^2(\omega)}{1+\varepsilon^2 T_n^2(\omega)} \tag{29}$$

where

$$T_n(\omega)=\cos\left[n\cos^{-1}(\omega)\right] \tag{30}$$

and $T_n(\omega)$ is the Chebyshev polynomial of n^{th} degree.
with the resulting lowpass ladder network having infinite Qu and terminated in a 1Ω resistor.

For the case of uniform Qu, it may readily be demonstrated that

$$|S_{11}(j\omega)|^2=\frac{(\omega-\omega_r)^2}{[1+(\omega-\omega_r)^2]} \tag{31}$$

and using optimisation the ω_r, $r = 1 \rightarrow n$ may be chosen to provide an equiripple response. For example, for $n = 7$, the ω_r are 0, ±1.95, ±3.85, ±5.51 to give a 20dB stopband level. However, as in the maximally flat case, as the degree of the network is increased, the selectivity tends to a limit and this prototype has restricted value.

It is not obvious how to extend the desirable maximally flat response given in equation (14) to an equiripple response. However, it is possible to readily generate the quasi equiripple function:

$$|S_{11}(j\omega)|^2 = \frac{\alpha^2(1-\omega^2)(1-T_{n+1}^2(\alpha\omega))}{(1-\alpha^2\omega^2)(T_{n+1}^2(\alpha)-T_{n+1}^2(\alpha\omega))} \qquad (32)$$

$$(\alpha>1)$$

where α is a constant which determines the ripple level.

This function is of degree n in ω^2 due to the cancellation of both the factors $(1-\omega^2)$ and $(1-\alpha^2\omega^2)$. The ripple level in the stopband is given by:

$$L_1 = \frac{\alpha^2}{T_{n+1}^2(\alpha)} \qquad (33)$$

For L_1 = 20dB, then α = 1.100, 1.077, 1.060 for n = 6,7 and 8 respectively. Thus, since α is of the order of unity, the factor

$$\frac{\alpha^2(1-\omega^2)}{(1-\alpha^2\omega^2)} \qquad (34)$$

causes very little deviation from the equiripple behaviour. This is illustrated in Fig 4 for the case of n = 7.

Forming the bounded real reflection coefficient from equation (32) gives

$$S_{11}(p,\alpha) = \prod_{r=1}^{n}\left[\frac{(p\alpha-j\cos\theta_r)}{(p\alpha-j\cos(\cos^{-1}(\alpha)+\theta_r)}\right] \qquad (35)$$

where

$$\theta_r = \frac{r\pi}{(n+1)} \tag{36}$$

Forming the input impedance:

$$Z(p,\alpha) = \frac{1+S_{11}(p,\alpha)}{1-S_{11}(p,\alpha)} \tag{37}$$

then the ladder structure shown in Fig 5 may be synthesised to give the characteristic admittance of the inverters as:

$$K_{01} = \sqrt{\frac{\sqrt{\alpha^2-1}}{\alpha}}$$

$$K_{r,r+1} = \frac{\sqrt{\alpha^2-\cos^2(\theta r)}}{\alpha}\; r=1 \to n-1 \tag{38}$$

$$\theta_r = \frac{r\pi}{(n+1)}$$

and the admittance of the rth shunt element is

$$Y_r = C_r p + G_r \frac{\sqrt{\alpha^2-1}}{\alpha} \tag{39}$$

where C_r and G_r are as given in equation (22).

We have constructed and tested two experimental devices in accordance with the invention. The first was of degree 5 with a 20dB stopband of just less than 1 MHz at a centre frequency of 840MHz. The cavities were TEM resonators where the variable Q was obtained by varying the diameters of the cavities. The measured response is shown in Fig 6, showing good agreement with theory. The second device was of degree 6 with a 20dB bandwidth in excess of 1.1MHz and tuned to a similar centre frequency. In this case, energy was decoupled from the last cavity to form a transmission characteristic:

$$|S_{12}(j\omega)|^2=\frac{K^2(1-\omega^2)}{(T^2_{n+1}(\alpha)-T^2_{n+1}(\alpha\omega))} \tag{40}$$

and the measured results in Fig 7 again demonstrate good agreement with theory.

An additional coupling was introduced between the input and the third resonator to produce a pair of real frequency transmission zeros close to the band-edge and then the network was optimised. This had the effect of reducing the reflected loss at the points and measured results are shown in Figs 8 and 9, showing that a significant reduction in loss can be achieved at the passband edges. In this case, the first two cavities were dielectric resonators.

It will be appreciated that in the variable Q reflection mode filters which have been described, the particular choice of the maximally flat and quasi equiripple solutions to the approximation problem have been shown to lead to explicit formulas for the element values in the ladder network realisations. Experimental devices of degree 5 and 6 have been designed and constructed with the measured results showing good agreement with theory. It will also be appreciated that a reciprocal device can be produced by connecting two reflection mode filters to the output ports of a 3dB hybrid.

Claims

1) A microwave reflection mode filter comprising a circulator device (12) having first, second and third ports (1,2,3), the first port (1) forming a signal input port and the third port (3) forming a signal output port, and a one-port filter (10) connected to and terminating the second port (2) of the circulator device (12), characterised in that the filter exhibits a maximally flat response and the one-port filter (10) comprises a ladder network of n resonators and having a reflection coefficient given by:

$$S_{11}(p)=\prod_{r=1}^{n}\frac{p^{n}}{(p-je^{j\theta_r})}$$

where θ_r is a constant applicable to the rth resonator of the ladder and given by:

$$\theta_r=\frac{r\pi}{n+1}$$

and p is the complex frequency variable, the admittance $K_{r-1,r}$ of the inverter coupling between the $(r-1)^{th}$ and r^{th} resonator is given by:

$$K_{r-1,r}=1 \quad r=1\rightarrow n$$

and the Q factors of the resonators are given by:

$$Q_r=\frac{C_r}{G_r}$$

where Q_r is the Q factor of the rth resonator, C_r is the capacitance of the rth resonator, G_r is the conductance of the rth resonator, and C_r and G_r are defined by

$$C_r = \frac{1}{\cos\theta}\left[\frac{\sin((r-1)\theta)}{E_{r-1}} + \sin(r\theta)E_r\right]$$

$$G_r = \frac{1}{E_{r-1}} - E_r$$

where E_r is a variable applicable to the rth resonator and defined by:

$$E_{r-1}E_r = \left[\frac{\cos\theta + \cos(r\theta)}{\cos\theta + \cos((r-1)\theta)}\right]$$

and:

$$\theta = \frac{\pi}{n+1}$$

2) A microwave reflection mode filter according to claim 1, characterised in that transmission zeros are provided.

3) A microwave reflection mode filter according to claim 2, characterised in that said transmission zeros are formed by coupling provided between the input of the one-port filter (10) and one of its resonators.

4) A microwave reflection mode filter according to claim 3, characterised in that said coupling is provided between the input of the one-port filter (10) and its third resonator.

5) A microwave reflection mode filter comprising a

circulator device (12) having first, second and third ports (1,2,3), the first port (1) forming a signal input port and the third port (3) forming a signal output port, and a one-port filter (10) connected to and terminating the second port (2) of the circulator device (12), characterised in that the filter exhibits an equiripple or quasi-equiripple response and said one-port filter (10) comprises a ladder network of n resonators and having a reflection coefficient given by:

$$S_{11}(p,\alpha)=\prod_{r=1}^{n}\left[\frac{(p\alpha-j\cos\theta_r)}{(p\alpha-j\cos(\cos^{-1}(\alpha)+\theta_r)}\right]$$

where θ_r is a constant applicable to the rth resonator of the ladder and given by:

$$\theta_r=\frac{r\pi}{n+1}$$

and p is the complex frequency variable, α is a constant and the admittances of the inverter couplings are given by:

$$K_{01}=\sqrt{\frac{\sqrt{\alpha^2-1}}{\alpha}}$$

$$K_{r,r+1}=\frac{\sqrt{\alpha^2-\cos^2(\theta r)}}{\alpha}\; r=1\rightarrow n-1$$

the admittance Y_r of the r^{th} resonator is given by

$$Y_r=C_r p+G_r\frac{\sqrt{\alpha^2-1}}{\alpha}$$

and the Q factors of the resonators is given by:

$$Q_r = \frac{C_r}{G_r}$$

where Q_r is the Q factor of the rth resonator, C_r is the capacitance of the rth resonator, G_r is the conductance of the rth resonator, and C_r and G_r are defined by:

$$C_r = \frac{1}{\cos\theta}\left[\frac{\sin((r-1)\theta)}{E_{r-1}} + \sin(r\theta)E_r\right]$$

$$G_r = \frac{1}{E_{r-1}} - E_r$$

where E_r is a variable applicable to the rth resonator and defined by:

$$E_{r-1}E_r = \left[\frac{\cos\theta + \cos(r\theta)}{\cos\theta + \cos((r-1)\theta)}\right]$$

and:

$$\theta = \frac{\pi}{n+1}$$

6) A microwave reflection mode filter according to claim 5, characterised in that transmission zeros are provided.

7) A microwave reflection mode filter according to claim 6, characterised in that said transmission zeros are formed by coupling provided between the input of the one-port filter (10) and one of its resonators.

8) A microwave reflection mode filter according to claim 7, characterised in that said coupling is provided between the input of the one-port filter (10) and its third resonator.

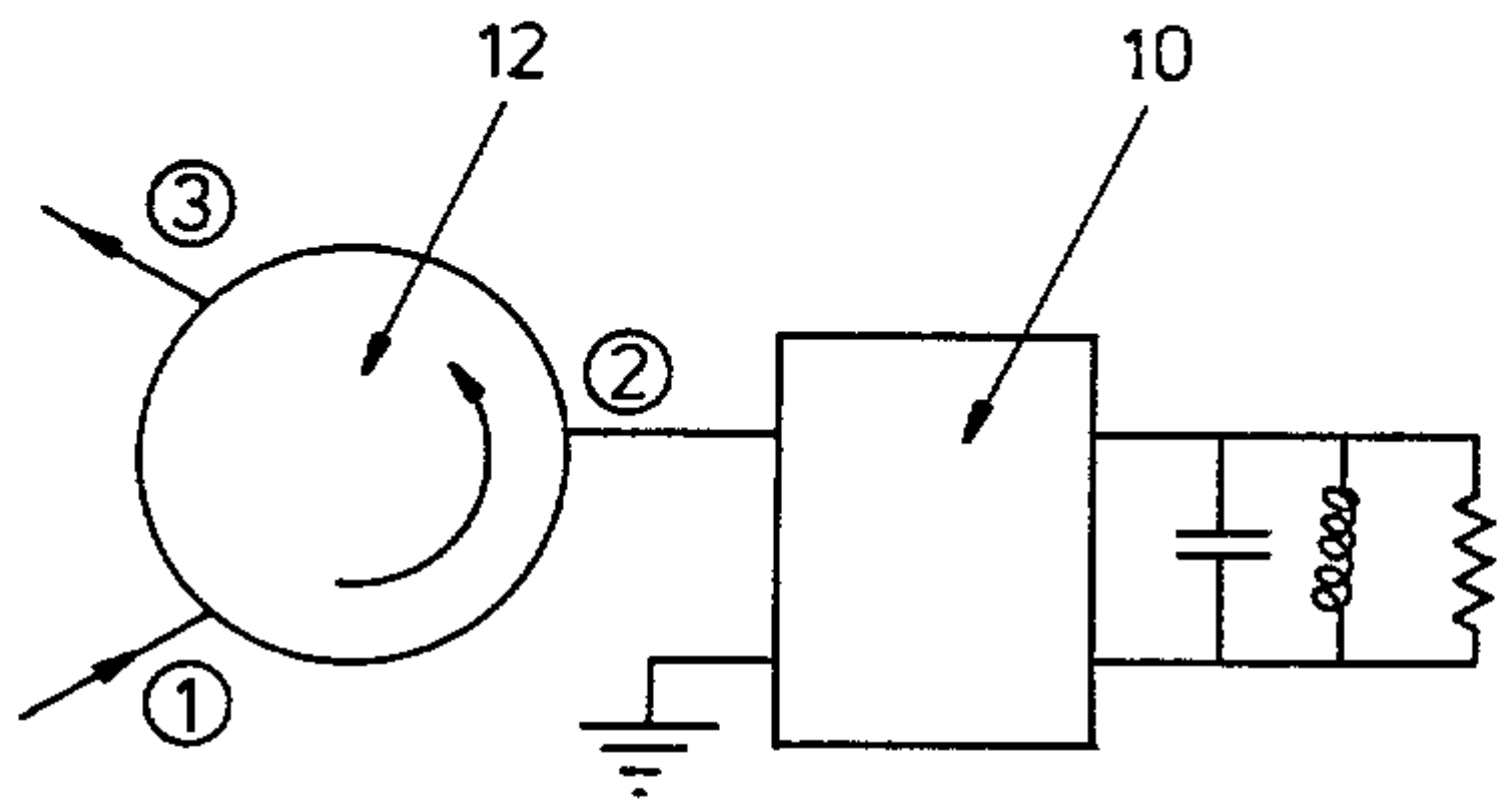

FIG. 1

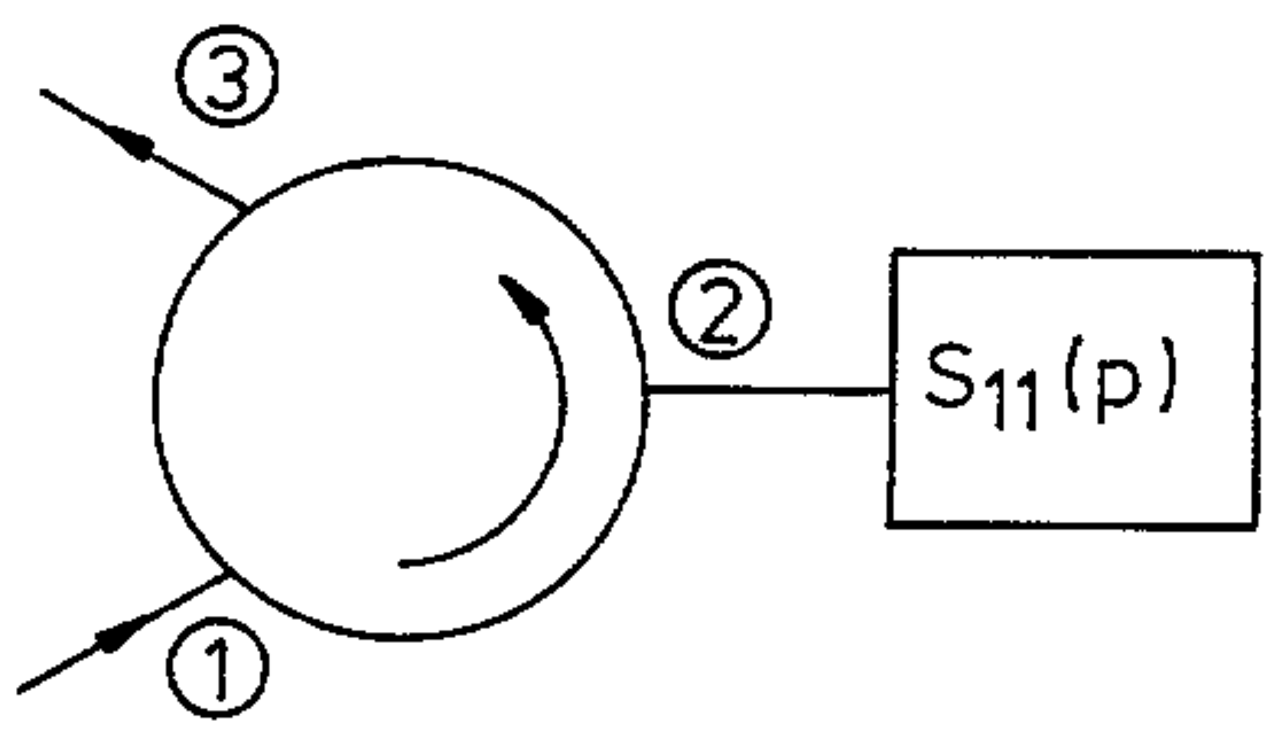

FIG. 2

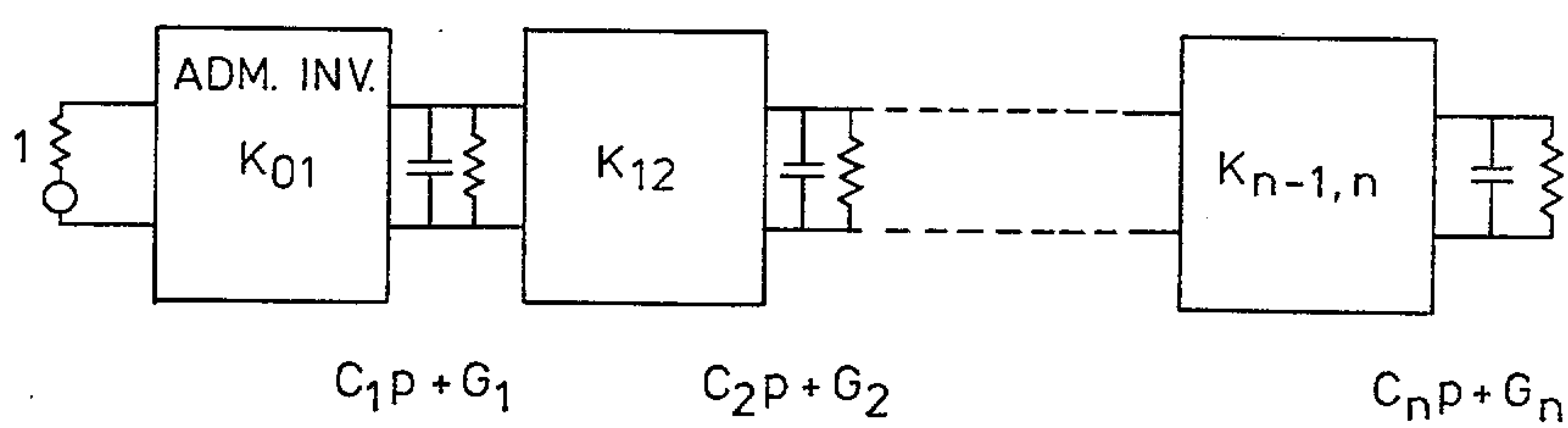

FIG. 3

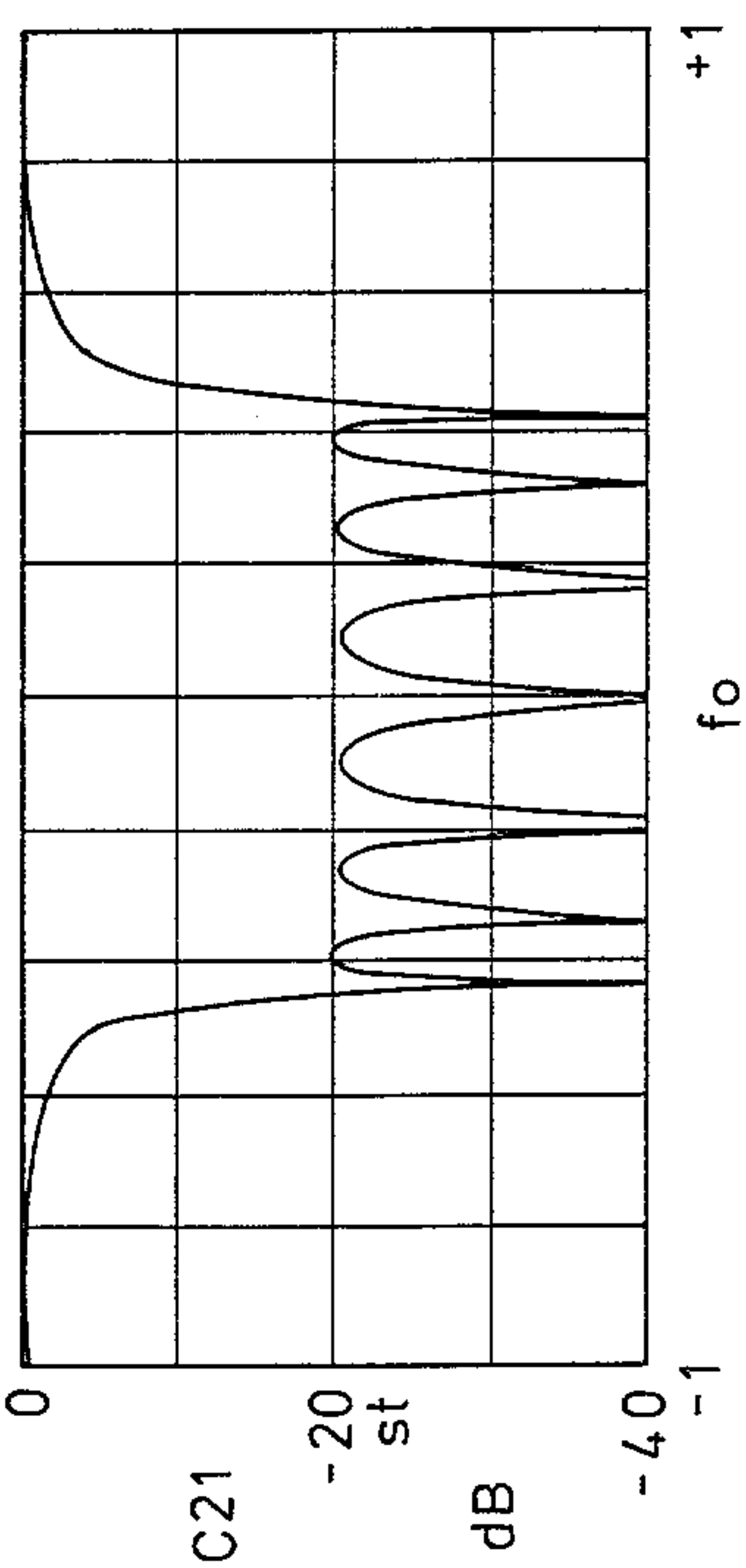

FIG. 4

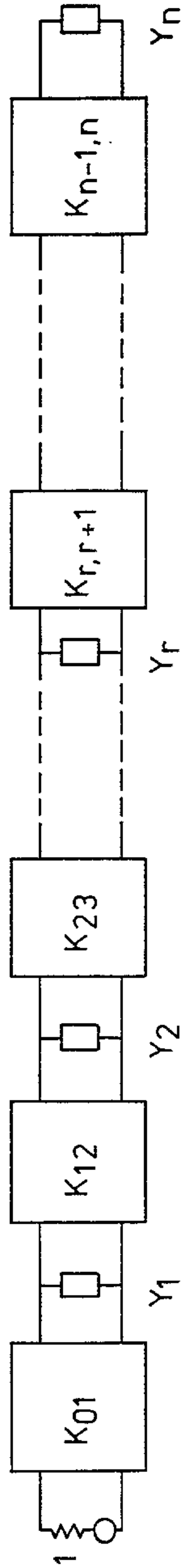

FIG. 5

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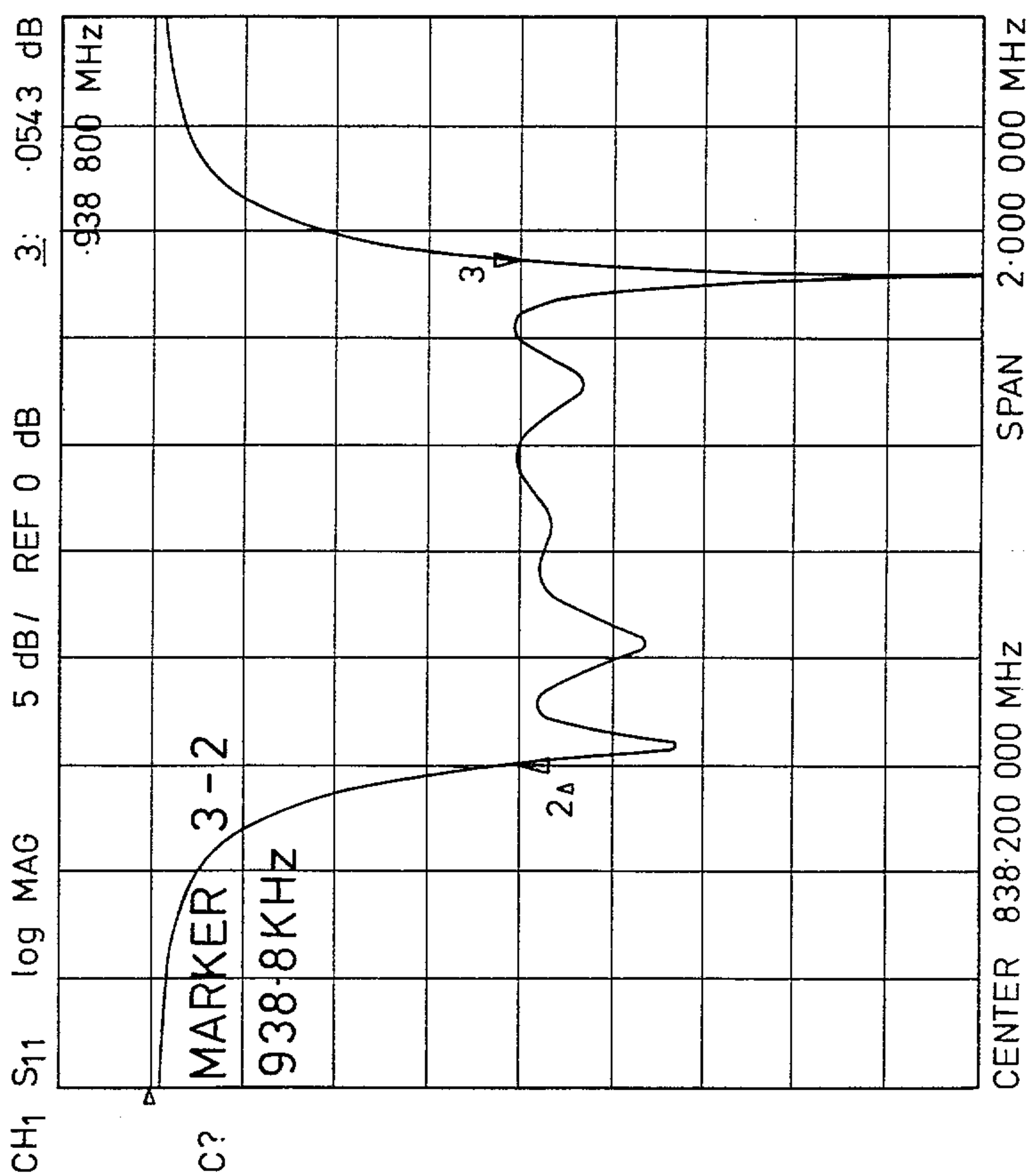


FIG. 6

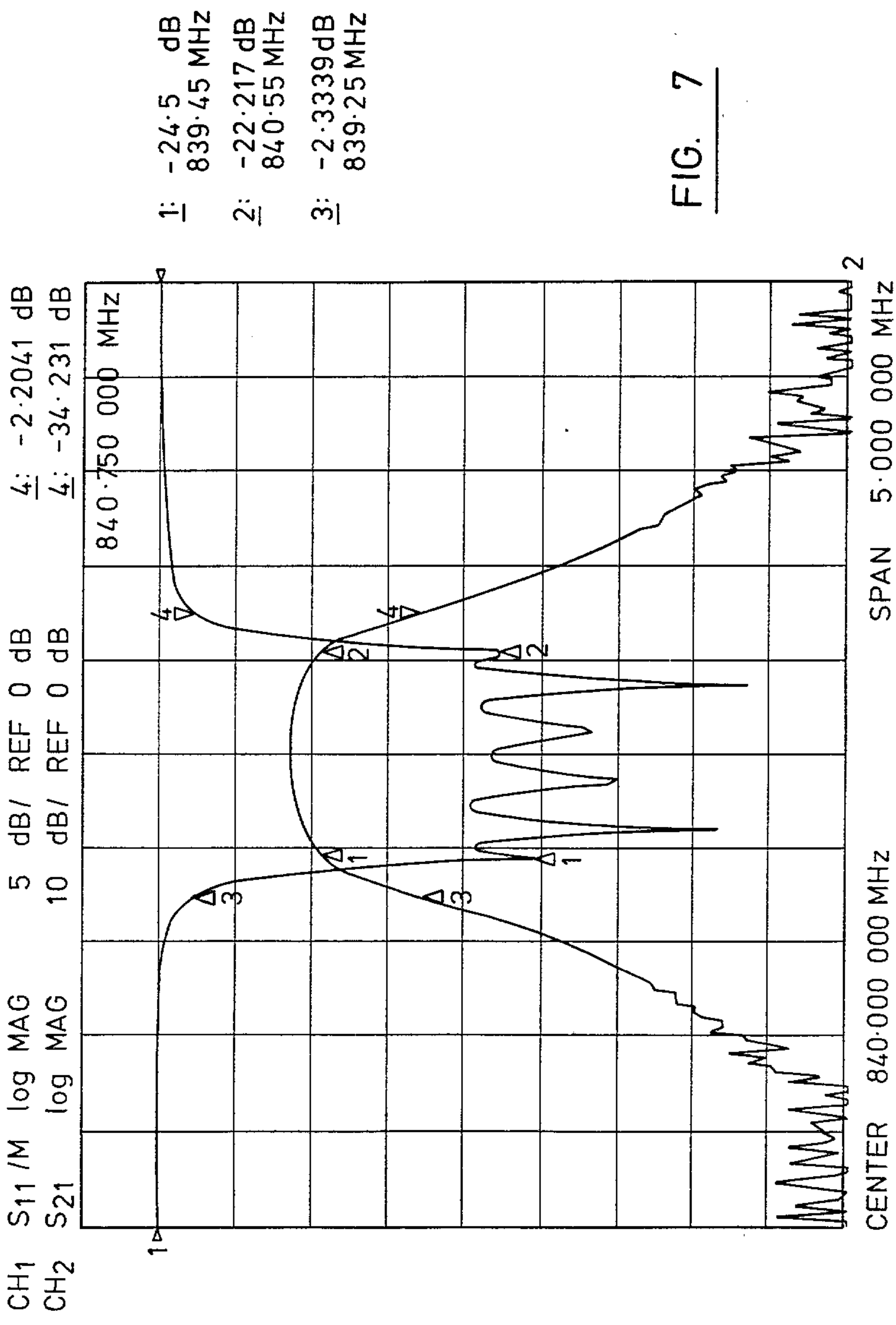

1: -24·5 dB
839·45 MHz
2: -22·217 dB
840·55 MHz
3: -2·3339 dB
839·25 MHz
CH1 S11 /M log MAG 5 dB/ REF 0 dB 4: -2·2041 dB
CH2 S21 log MAG 10 dB/ REF 0 dB 4: -34·231 dB
840·750 000 MHz
CENTER 840·000 000 MHz SPAN 5·000 000 MHz

FIG. 7

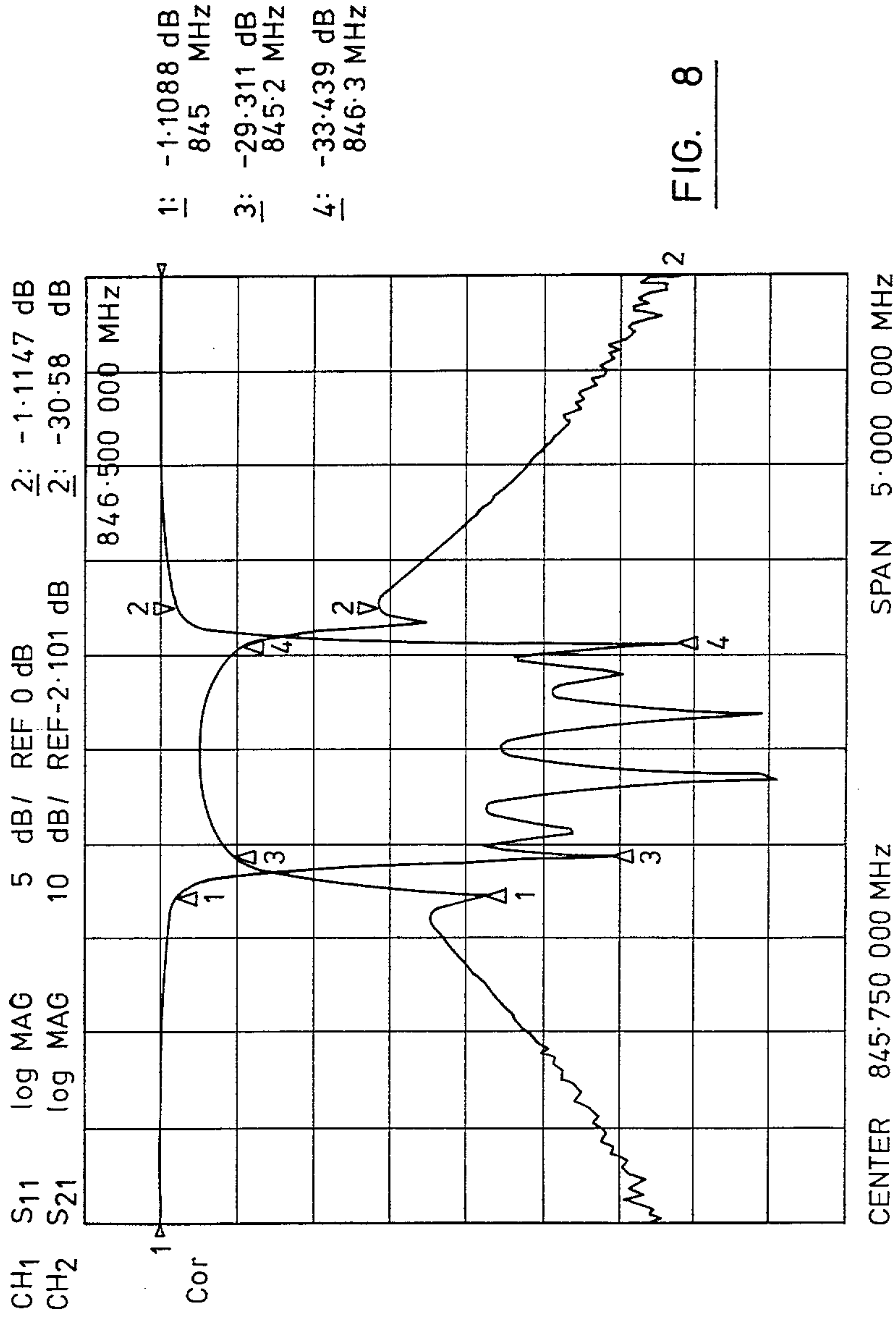

CH1 S11 log MAG 5 dB/ REF 0 dB 2: -1·1147 dB
CH2 S21 log MAG 10 dB/ REF-2·101 dB 2: -30·58 dB
846·500 000 MHz
1: -1·1088 dB
845 MHz
3: -29·311 dB
845·2 MHz
4: -33·439 dB
846·3 MHz
Cor
CENTER 845·750 000 MHz SPAN 5·000 000 MHz

FIG. 8

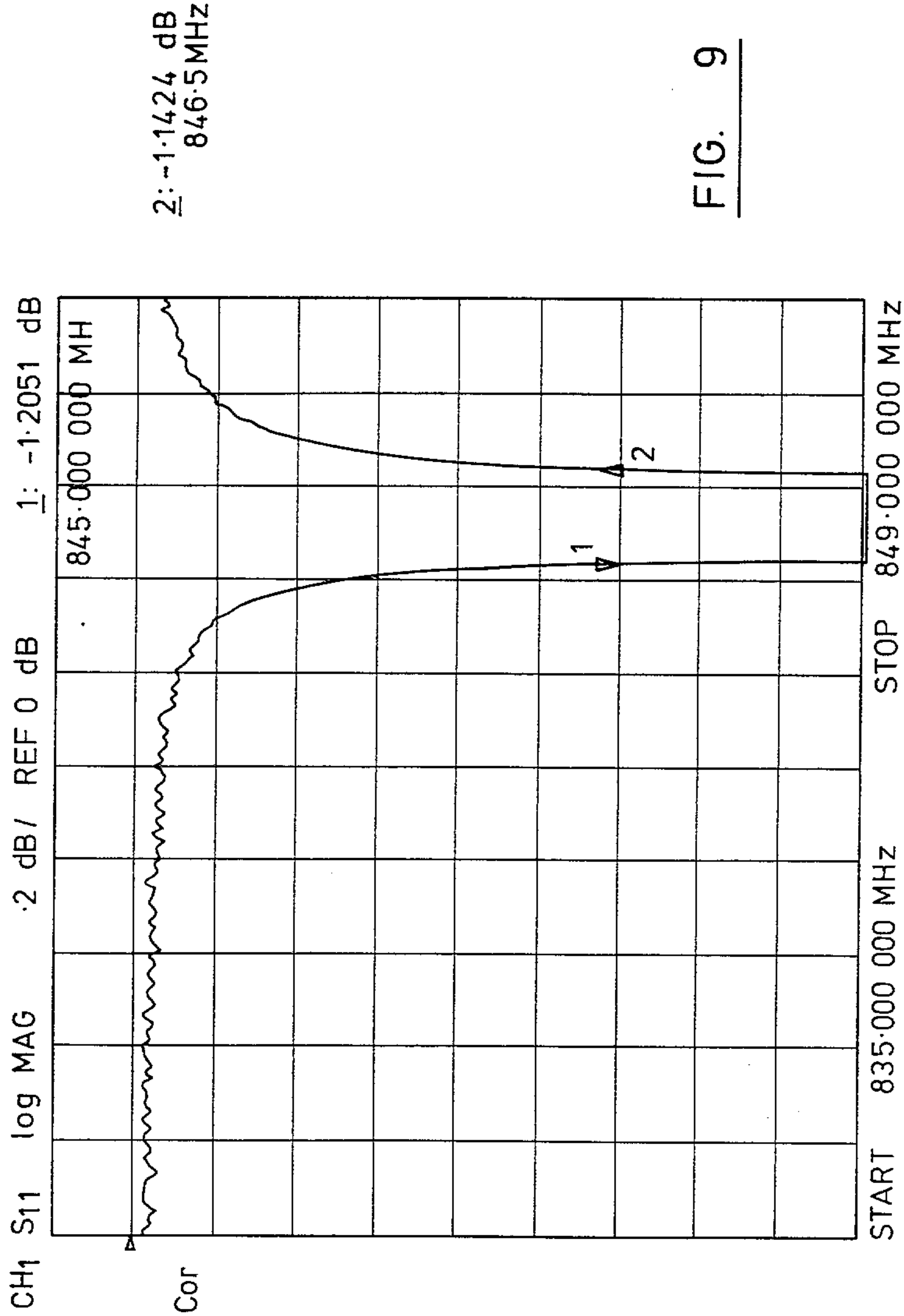
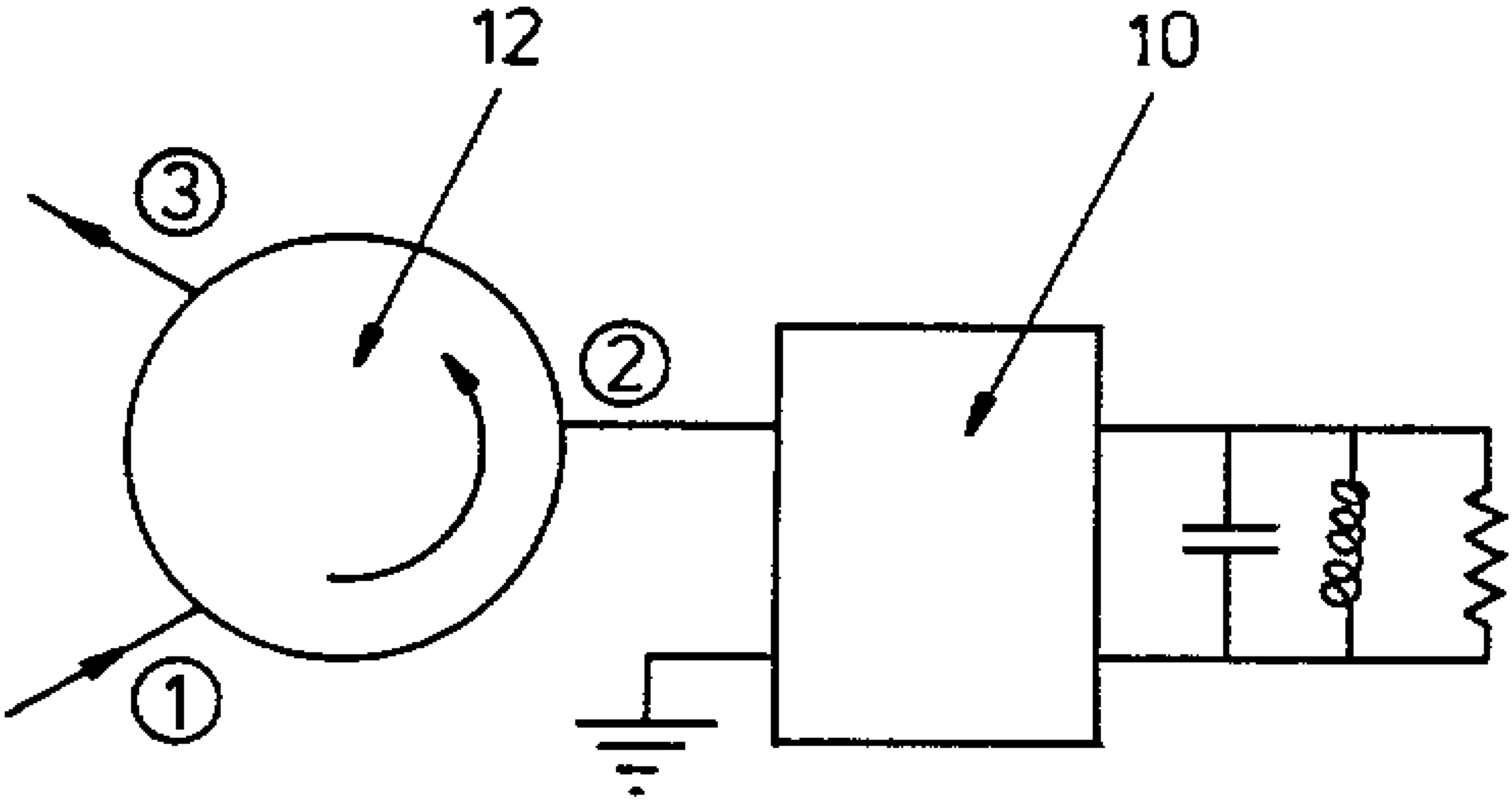

2: -1·1424 dB
846·5MHz
CH1 S11 log MAG ·2 dB/ REF 0 dB 1: -1·2051 dB
845·000 000 MH
Cor
1
2
START 835·000 000 MHz STOP 849·000 000 MHz

FIG. 9


12
10
3
2
1