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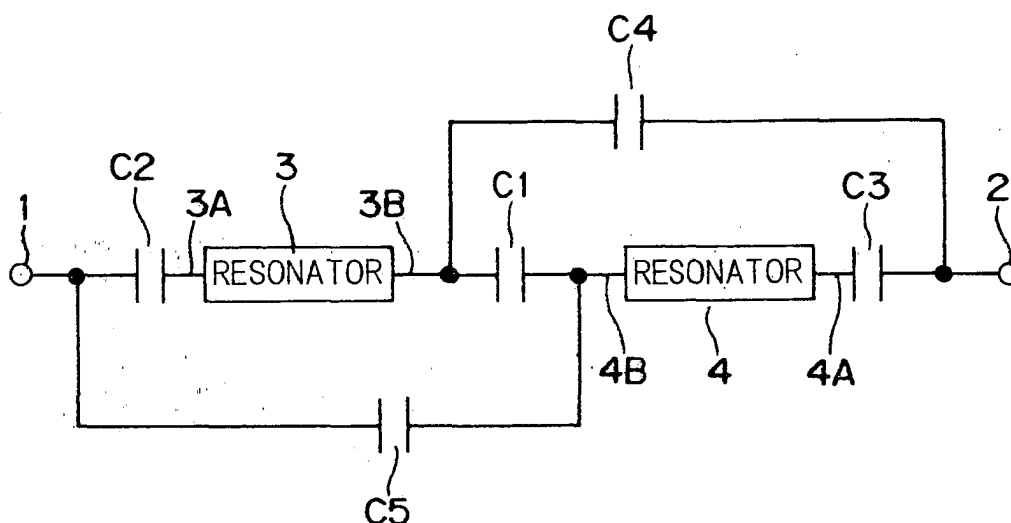
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(54) **High-frequency filter circuit, monolithic microwave integrated circuit and high-frequency communication apparatus**

(57) In this high-frequency filter circuit, a fourth capacitance C4 capacitively couples a first resonator 3 to an output port 2 bypassing a first capacitance C1, a second resonator 4 and a third capacitance C3, while a fifth capacitance C5 capacitively couples a second resonator 4 to an input port 1 bypassing the first capacitance C1, the first resonator 3 and a second capacitance C2. According to this filter circuit, the first resonator 3 is directly capacitively coupled to the output port 2 by the existence of this fourth capacitance C4, while the sec-

ond resonator 4 is directly capacitively coupled to the input port 1 by the existence of the fifth capacitance C5. According to the above-mentioned high-frequency filter circuit, there can easily be manufactured a high-frequency filter circuit, which is able to achieve downsizing and area saving and suitable for producing an MMIC and in which the higher region side of the passband is steeply sloped by a transmission zero, generating no parasitic passband at a frequency double that of the passband.

Fig.1



Description

BACKGROUND OF THE INVENTION

[0001] The present invention relates to a high-frequency filter circuit that constitutes, for example, a band-pass filter and relates, in particular, to a high-frequency filter circuit, a monolithic microwave integrated circuit and a high-frequency communication apparatus, which constitute a distributed element filter having a high-frequency transmission line such as a microstrip line and a coplanar line for use in a high-frequency band such as a microwave band and a millimeter wave band.

[0002] Fig. 14 is a schematic view of a typical distributed element filter circuit. This filter circuit has a microstrip line as a high-frequency transmission line on which the distributed element is based. In Fig. 14 are shown a dielectric substrate 105 that includes a dielectric 107 made of ceramic or the like, a GND pattern 108 on the lower surface of the dielectric substrate 107, an input port 101 and an output port 102. There are further provided resonators 103 and 104, which are so-called the $\lambda/2$ open-circuit line type resonators and in which the length of the microstrip line is designed to have a dimension of approximately $\lambda/2$ with respect to a wavelength λ of the center frequency of the filter

[0003] Both ends of these resonators 103 and 104 are open ends. A capacitance C101 shown in Fig. 14 is a capacitance component generated by a gap between the resonator 103 and the resonator 104. A capacitance C102 is a capacitance component generated by a gap between an input feed line 111 connected to an input port 101 and the $\lambda/2$ resonator 103. A capacitance C103 is a capacitance component generated by a gap between an output feed line 112 connected to the output port 102 and the $\lambda/2$ resonator 104.

[0004] Distributed constant filter circuits, as shown in this Fig. 14, which have been able to form a filter circuit of only one-layer printed wiring, easy to manufacture and able to restrain the cost low, have been frequently used in the frequency band of about 5 to 30 GHz. Further, the distributed element filter circuits as described above have recently been found here and there as examples used in the millimeter wave band of 30 to 60 GHz.

[0005] The filter circuit shown in this Fig. 14 has an equivalent circuit as shown in Fig. 15A. It is further known that its filter characteristics generally become as shown in Fig. 16. The equivalent circuit shown in Fig. 15B is obtained by folding double the $\lambda/2$ open-circuit line resonators 103 and 104 of the equivalent circuit shown in Fig. 15A taking advantage of the phenomenon that its center has a potential equal to that of the GND (ground) at and around the resonance frequency.

[0006] The characteristics shown in Fig. 16 are the results of designing the equivalent circuit shown in Fig. 15A by means of a circuit simulator on the market with the center frequency set at 12 GHz. The characteristics

shown in this Fig. 16 are the results of simulation when the capacitance C101 = 0.17145 pF, the capacitance C102 = C103 = 0.29200 pF, the characteristic impedance of the resonators 103 and 104 is $10\ \Omega$ and the electrical length at 12 GHz is 172 degrees.

[0007] As shown in Fig. 16, the characteristics of the high-frequency circuit are usually expressed by S parameters. That is, the characteristics are expressed by a parameter S11 that represents a reflection coefficient and a parameter S21 that represents a transmission coefficient.

[0008] The filter circuit of the prior art shown in Fig. 14 has the following three problems i), ii) and iii).

[0009] That is, i) the first problem has been the low degree of steepness of the filter characteristics. As is apparent with reference to Fig. 16, the parameter S21, which represents the transmission coefficient, has no transmission zero and has a gently sloping bandpass characteristic in this filter. In particular, it is more difficult to achieve a steep attenuation characteristic on the higher region side of the passband due to the second problem described next.

[0010] ii) As the second problem, there is the problem that a parasitic passband is disadvantageously generated at a frequency double that of the proper passband. This also becomes a cause of the aforementioned first problem. The characteristics of Fig. 16 are the results of designing with the center frequency set at 12 GHz. As is clear with reference to the transmission coefficient parameter S21, a second passband is generated at and around 24 GHz, or the frequency double this center frequency.

[0011] This second passband is a parasitic passband generated by the secondary resonance of the $\lambda/2$ open-circuit line type resonators 103 and 104. There is a tendency that the transmission coefficient parameter S21 in the frequency band of 12 GHz to 24 GHz increases as a whole due to this parasitic passband, and the attenuation value is degraded particularly at and around the frequencies of 12 GHz and 24 GHz.

[0012] iii) As the third problem, there has been the problem of a large circuit area. Particularly when producing a narrow-band bandpass filter, it is required to reduce the degree of coupling (capacitance C101) by expanding the gap between the resonators 103 and 104 in Fig. 14, and a dead space is generated in the direction indicated by arrow B in Fig. 14, causing an increased dimension.

[0013] On the other hand, it is known that reducing the number of components and the inter-circuit connecting portions by integrating a plurality of circuits in a one-chip form on an MMIC (monolithic microwave integrated circuit) is a very effective technique in terms of improving both the electrical performance and the manufacturing cost in an ultra-high-frequency band such as the millimeter wave band. This can also be said for the filter circuit, and there is an intense demand for integrating the filter circuit with an amplifier circuit and a mixer circuit, which

are connected before and behind the filter circuit, in a one-chip form on the MMIC. In particular, for the reduction in the chip cost of the MMIC, it is required to reduce the circuit area as far as possible by removing the dead space in the layout stage.

[0014] In contrast to this, it is difficult for the prior art filter shown in Fig. 14 to reduce the chip cost due to the obstacle of the large dead space in the direction of arrow B in Fig. 14.

[0015] Therefore, as shown in Fig. 17A, for example, the conventional millimeter wave band communication apparatus has been constructed of a local oscillator 201, a balance type mixer 202, an amplifier 203 and an antenna 204, and there has been often employed a balance type image rejection mixer 202 constructed of two mixers 205 and 206 instead of the filter for the purpose of image rejection. The reason for the above is the difficulties in achieving a degree of steepness required for the image rejection in the millimeter wave band in the prior art filter circuit that has no transmission zero as shown in Fig. 14. However, the balance type image rejection mixer has generally had a drawback that the bandwidth has been narrow, and it has been difficult to satisfy the demand of the system that has had a bandwidth of, for example, up to 2 to 3 GHz only by the balance type image rejection mixer. Moreover, when the balance type image rejection mixer is employed, it is usual that the chip area is enlarged double or more in comparison with the normal mixer circuit that is not the balance type. This has disadvantageously caused an increase in the chip unit price and difficulties in integrating any more other circuits (amplifier circuit and the like) on an identical chip.

[0016] In another case, a waveguide filter 303 has often been employed as a filter capable of obtaining high performance even in the millimeter wave band for the purpose of image rejection, as shown in Fig. 17B. However, in this case, there have been the drawbacks of difficulties in electrical connection between the waveguide that constitutes the waveguide filter 303 and the MMIC that constitutes the mixer 302 and the amplifier 304, as well as the expensiveness, large size and heavy weight of the waveguide filter 303 itself.

SUMMARY OF THE INVENTION

[0017] Accordingly, the object of this invention is to solve the aforementioned problems and provide a compact area-saving easily manufacturable high-frequency filter circuit, monolithic microwave integrated circuit and high-frequency communication apparatus, which is suitable for producing an MMIC and in which the higher region side of the passband is steeply sloped by a transmission zero, generating no parasitic passband at a frequency double that of the passband.

[0018] In order to achieve the above object, there is provided a high-frequency filter circuit wherein a first resonator and a second resonator are ca-

pacitively coupled to each other by a first capacitance, one terminal of the first resonator is capacitively coupled to an input port by a second capacitance, one terminal of the second resonator is capacitively coupled to an output port by a third capacitance, the other terminal of the first resonator is capacitively coupled to the output port by a fourth capacitance, and

the other terminal of the second resonator is capacitively coupled to the input port by a fifth capacitance.

[0019] In this invention, the other terminal of the first resonator is capacitively coupled to the output port by the fourth capacitance, while the other terminal of the second resonator is capacitively coupled to the input port by the fifth capacitance. That is, the fourth capacitance capacitively couples the first resonator to the output port bypassing the first capacitance, the second resonator and the third capacitance. The fifth capacitance capacitively couples the second resonator to the input port bypassing the first capacitance, the first resonator and the second capacitance.

[0020] As described above, according to this invention, the first resonator is directly capacitively coupled to the output port by the existence of this fourth capacitance, while the second resonator is directly capacitively coupled to the input port by the existence of the fifth capacitance. By the so-called jump coupling as described above, a transmission zero was able to be formed in the frequency characteristic curve of the transmission coefficient S₂₁ of the S parameters, and the degree of steepness of this frequency characteristic curve was able to be increased, allowing the filter characteristics to be improved.

[0021] In one embodiment of the present invention, the first resonator is comprised of a first line, the second resonator is comprised of a second line,

the first resonator and the second resonator are coupled to each other by the first capacitance with the first line facing the second line at a prescribed distance,

the input port is coupled to the first resonator by the second capacitance with a third line that includes the input port facing the first line,

the input port is coupled to the second resonator by the fifth capacitance with the third line facing the second line,

the output port is coupled to the second resonator by the third capacitance with a fourth line that includes the output port facing the second line, and

the output port is coupled to the first resonator by the fourth capacitance with the fourth line facing the first line.

[0022] In this embodiment, by making the fourth line that includes the output port face the first line that constitutes the first resonator, the output port is coupled to the first resonator by the fourth capacitance in a jump coupling manner. Moreover, by making the third line that includes the input port face the second line that constitutes the second resonator, the input port is coupled to

the second resonator by the fifth capacitance in a jump coupling manner.

[0023] As described above, in this embodiment, the fourth and fifth capacitances for effecting the jump coupling can be constructed of the third and fourth lines that constitute the input and output ports. Without newly adding any component, the embodiment is provided by devising the line arrangement in the vicinity of the resonators. Therefore, a bandpass filter circuit, in which the higher region side of the passband is steeply sloped by the transmission zero, can easily be provided by a simple circuit structure. Moreover, with the above-mentioned line arrangement, there can be provided a compact, area-saving and easily manufacturable high-frequency filter circuit, which can be minimized in size by removing the dead space and is also suitable for producing an MMIC.

[0024] In one embodiment of the present invention, a length of a confronting portion where the first line faces the second line is set within a range of 15% to 20% of a wavelength of a center frequency of a passband.

[0025] In the high-frequency filter circuit of this embodiment, by setting the length of the confronting portion where the first line that constitutes the first resonator face the second line that constitutes the second resonator within the range of 15% to 20% of the wavelength of the center frequency of the passband, the parasitic passband in the frequency band double that of the passband is suppressed. If the length of the confronting portion is out of the above-mentioned range, the parasitic passband cannot be suppressed.

[0026] In one embodiment of the present invention, the high-frequency filter circuit of the present invention is integrally formed on an identical semiconductor substrate together with an amplifier circuit or a mixer circuit.

[0027] According to the monolithic microwave integrated circuit of this embodiment, by integrally forming the high-frequency filter circuit on an identical semiconductor substrate together with an amplifier circuit or a mixer circuit, a one-chip up-converter MMIC can be provided. In addition to the cost reduction, downsizing and weight reduction of the single unit of the filter circuit, there can also be obtained the synergistic effects of achieving the remarkable simplification of the whole system, the reduction in the number of components and the simplification of the manufacturing processes.

[0028] In one embodiment of the present invention, a high-frequency radio communication apparatus, which has the high-frequency filter circuit of the present invention as a spurious rejection filter.

[0029] According to this embodiment, there can be provided a high-frequency radio communication apparatus having the spurious rejection filter, which can suppress the parasitic passband in the frequency band double that of the passband.

BRIEF DESCRIPTION OF THE DRAWINGS

[0030] The present invention will become more fully understood from the detailed description given hereinbelow and the accompanying drawings which are given by way of illustration only, and thus are not limitative of the present invention, and wherein:

Fig. 1 is an equivalent circuit diagram showing the principle of a transmission zero formation in an embodiment of the high-frequency filter circuit of this invention;

Fig. 2 is a graph showing one example of the result of simulating the frequency characteristic of the equivalent circuit of Fig. 1;

Fig. 3 is a layout diagram showing the construction of strip lines according to a first embodiment of the high-frequency filter circuit of this invention;

Fig. 4 is a graph showing one example of the result of simulating the frequency characteristic of the filter circuit of the first embodiment of the construction shown in Fig. 3;

Fig. 5 is a graph of a simulated characteristic showing a state in which the attenuation value in the frequency band at the double frequency depends on a distance L in the circuit of the construction of Fig. 3;

Fig. 6 is a layout diagram showing the construction of strip lines according to a modification example of the first embodiment;

Fig. 7 is a graph showing a frequency characteristic actually measured on the basis of the modification example of the construction of Fig. 6;

Fig. 8 is a layout diagram showing the construction of strip lines according to a second embodiment of this invention;

Fig. 9 is a graph of a simulated characteristic of the second embodiment of the construction of Fig. 8;

Fig. 10 is a layout diagram showing the construction of strip lines according to a third embodiment of this invention;

Fig. 11 is a graph of a simulated characteristic of the third embodiment of the construction of Fig. 10;

Fig. 12 is a graph showing one example of a simulated characteristic in the millimeter wave band of the circuit shown in Fig. 1;

Fig. 13 is a diagram showing a circuit of a millimeter wave radio communication apparatus according to a fourth embodiment of this invention;

Fig. 14 is a diagram showing the structure of a conventional distributed element high-frequency filter circuit;

Fig. 15A is an equivalent circuit of the filter circuit of Fig. 14, and Fig. 15B is another equivalent circuit;

Fig. 16 is a graph showing one example of a simulated characteristic of the equivalent circuit of Fig. 15A; and

Fig. 17A is a block diagram of one example of a con-

ventional millimeter wave radio communication apparatus, and Fig. 17B is a block diagram of another example of the circuit of a conventional millimeter wave radio communication apparatus.

DETAILED-DESCRIPTION OF THE PREFERRED EMBODIMENTS

[0031] The embodiments of this invention will be described more concretely in detail below with reference to the drawings.

(First Embodiment)

[0032] Fig. 1 shows an equivalent circuit diagram of the first embodiment of the high-frequency filter circuit of this invention. In this first embodiment, one example of the 12-GHz band high-frequency filter is shown. The principle of the generation of a transmission zero on the higher region side of the passband of this high-frequency filter will be described with reference to the equivalent circuit of this Fig. 1.

[0033] In the equivalent circuit of this first embodiment, a first resonator 3 and a second resonator 4 are capacitively coupled to each other by a first capacitance C1. One terminal 3A of the first resonator 3 is capacitively coupled to an input port 1 via a second capacitance C2. One terminal 4A of the second resonator 4 is capacitively coupled to an output port 2 via a third capacitance C3.

[0034] The other terminal 3B of the first resonator 3 is capacitively coupled to the output port 2 via a fourth capacitance C4. The other terminal 4B of the second resonator 4 is capacitively coupled to the input port 1 via a fifth capacitance C5.

[0035] Neither the fourth capacitance C4 nor the fifth capacitance C5 exists in the equivalent circuit of the prior art of Fig. 15A. That is, the equivalent circuit of this Fig. 1 has a circuit configuration in which jump coupling capacitances C4 and C5 are newly added to the equivalent circuit of Fig. 15A. This fourth capacitance C4 is the capacitance for coupling the other terminal 3B of the first resonator 3 to the output port 2 jumping over (bypassing) the first capacitance C1, the second resonator 4 and the third capacitance C3. The fifth capacitance C5 is the capacitance for coupling the other terminal 4B of the second resonator 4 to the input port 1 jumping over (bypassing) the first capacitance C1, the first resonator 3 and the second capacitance C2.

[0036] Next, Fig. 2 shows one example of the result of simulating the response of the filter circuit shown in Fig. 1 (transmission coefficient S21 and reflection coefficient S11) performed by a circuit simulator on the market. Referring to the response characteristic of this Fig. 2, dissimilarly to the response characteristic of the prior art shown in Fig. 16, it can be understood that transmission zeros p1 and p2 are generated on the higher region side of the passband (11.5 to 12 GHz) and the degree

of steepness is greatly improved. The simulation result shown in Fig. 2 was calculated on the following parameter conditions. The conditions were: C1 (first capacitance) = 0.22843 pF, C2 (second capacitance) = C3 (third capacitance) = 0.33561 pF, and C4 (fourth capacitance) = C5 (fifth capacitance) = 0.05821 pF. The first and second resonators 3 and 4 had a characteristic impedance of 10 Ω , and the electrical length at a frequency of 12 GHz was set at 167 degrees.

[0037] In this first embodiment, as is apparent from the response characteristic simulation result of Fig. 2, the problem that the degree of steepness on the higher region side is low, or the first problem out of the aforementioned three problems of the prior art, can be solved. This fact can be explained by the equivalent circuit shown in Fig. 1.

[0038] Next, the second and third problems of the prior art are closely related to the line layout that constitutes the high-frequency filter circuit. Therefore, it is difficult to provide an explanation with the abstract equivalent circuit.

[0039] Therefore, the explanation is continued by means of characteristic charts (Figs. 4 and 5) that show the calculation results of an electromagnetic field simulator on the market by the moment method as follows. A characteristic chart (Fig. 7) is not a simulation result but an actual measurement result.

[0040] Fig. 3 shows one example of the actual layout of the equivalent circuit of Fig. 1 as a microstrip line circuit on a 0.47-mm thick PPO (polypropylene oxide) substrate 10. The lower surface side of this substrate 10 is entirely a grounding layer (not shown). In Fig. 3, only the electrode pattern on the upper surface side of the substrate 10 is shown. As shown in Fig. 3, this microstrip line circuit is constructed of a third line 31 that includes the input port 1, a first line 32 that constitutes the first resonator 3, a second line 33 that constitutes the second resonator 4 and a fourth line 34 that includes the output port 2. These lines 31, 32, 33 and 34 are entirely strip-shaped lines that have the widths in the Y-direction of Fig. 3 and are extended in the X-direction (lengthwise direction).

[0041] As shown in Fig. 3, the line 32 is displaced in the X-direction and the Y-direction with respect to the line 31, whereas the line 32 and the line 31 have a section along which the lines face each other in the Y-direction. The line 33 is displaced in the X-direction and the reverse Y-direction with respect to the line 32, whereas the line 33 and the line 32 have a section along which the lines face each other in the Y-direction. The line 33 and the line 31 do not face each other in the Y-direction but face each other in the X-direction. The line 34 is displaced in the X-direction and the Y-direction with respect to the line 33, whereas the line 34 and the line 33 have a section along which the lines face each other in the Y-direction. The line 34 and the line 32 do not face each other in the Y-direction but face each other in the X-direction.

[0042] The strip-shaped line 31 is constructed of a portion 31a that constitutes the input port 1 and a portion 31b that constitutes an input feed line 11. The width in the Y-direction of the portion 31b of this line 31 is approximately a half smaller than the width in the Y-direction of the portion 31a.

[0043] The portion 31b of this line 31 faces a portion 32a of the line 32 at a prescribed distance in the Y-direction. This portion 31b and the portion 32a constitute the second capacitance C2. This line 32 constitutes the first resonator 3, which is a $\lambda/2$ (1/2 wavelengths) open-circuit line type resonator.

[0044] Moreover, the portion 32b of this line 32 faces a portion 33a of the line 33 at a prescribed distance. An overlap distance over which this portion 33a and the portion 32b face each other is assumed to be L. This line 33 constitute the second resonator 4, which is a $\lambda/2$ open-circuit line type resonator. Moreover, the portion 33b of this line 33 faces a portion 34a that constitutes the output feed line of the line 34 at a prescribed distance in the Y-direction. This portion 33b and the portion 34a constitute the third capacitance C3. Moreover, the portion 34b of this line 34 constitutes the output port 2. The width in the Y-direction of the portion 34a is approximately a half of the width in the Y-direction of the portion 34b. Moreover, the portions 31a, 32a, 32b, 33a, 33b and 34b have approximately same widths in the Y-direction.

[0045] Further, an end portion 34a-1 of the portion 34a of this line 34 is adjacent to an end portion 32b-1 of the portion 32b of the line 32 at prescribed distances in the X-direction and the Y-direction. This end portion 34a-1 and the end portion 32b-1 constitute the fourth capacitance C4. Further, the end portion 33a-1 of the portion 33a of the line 33 is adjacent to the end portion 31b-1 of the portion 31b of the line 31 at prescribed distances in the X-direction and the Y-direction. This end portion 33a-1 and the end portion 31b-1 constitute the fifth capacitance C5.

[0046] That is, the fourth capacitance C4 for jump coupling is provided by arranging the open end 32b-1 located on the side opposite to the input port 1 of the line 32 that constitutes the resonator 3 close to the open end 34a-1 of the line 34 that constitutes the output port 2. Moreover, the other capacitance C5 for jump coupling is provided by arranging the open end 33a-1 located on the side opposite to the output port 2 of the line 33 that constitutes the resonator 4 close to the open end 31b-1 of the line 31 that constitutes the input port 1.

[0047] In this one example, minimum Line/Space (line and space) was set at 200 μm , and the lengths in the X-direction of the $\lambda/2$ resonators 3 and 4 were set at about 4.6 mm. The dimension L shown in Fig. 3 is the overlap distance of the arrangement of the two $\lambda/2$ resonators 3 and 4, i.e., the dimension L is the length along which the lines 32 and 33 face each other in the Y-direction. In Fig. 3, the length L was a half of the resonator length (dimensions in the X-direction of the lines 32 and 33), i.e., about $\lambda/4$ (1/4 wavelengths) in Fig. 3.

[0048] As described above, in the filter circuit of this first embodiment, the two resonator lines 32 and 33 and the two input/output lines 31 and 34 are arranged closer to each other than in the conventional structure as shown in Fig. 14, generating almost no dead space. Therefore, area saving can be achieved, and this arrangement is also suitable for use in the case of monolithic integration on a millimeter wave MMIC (monolithic microwave integrated circuit).

[0049] Next, Fig. 4 shows the result of simulating the circuit of the line layout shown in Fig. 3 by a moment method electromagnetic field simulator on the market. Referring to Fig. 4, similarly to the equivalent circuit simulation result shown in Fig. 2, transmission zeros p11 and p12 are generated on the higher region side of the passband (12 GHz), and the degree of steepness is improved. Also, according to the characteristic of this Fig. 4, similarly to the equivalent circuit simulation result shown in Fig. 2, a parasitic passband P13 is generated at a frequency (24 GHz) double that of the passband (12 GHz), degrading the attenuation value.

[0050] Accordingly, in the filter circuit of this embodiment, the attenuation value in this band of the doubled frequency can be improved by the method described as follows, if necessary.

[0051] That is, by adjusting the overlap distance L between the two $\lambda/2$ resonators 3 and 4 (distance between the lines 32 and 33 facing each other in the Y-direction) in the circuit of the construction shown in Fig. 3, the attenuation value in the band of the doubled frequency of the filter characteristic shown in Fig. 4 (region P13 in Fig. 4) changes. Fig. 5 shows this behavior in the form of a graph based on an electromagnetic field simulation result. In this Fig. 5, the horizontal axis represents a value obtained by dividing the overlap distance L in Fig. 3 by a wavelength λ at a center frequency f_0 of the passband, while the vertical axis represents the worst value (maximum value) of the transmission coefficient S21 in the band of 21 to 24 GHz. It can be understood that the transmission coefficient S21 in the band of 21 to 24 GHz (at and around $2f_0$) can be reduced when the distance L is 15 to 20% of the wavelength λ at the center frequency f_0 of the filter.

[0052] Fig. 6 shows a line layout, in which the overlap distances L between the resonators is shortened to about $\lambda/6$ of the line layout shown in Fig. 3, on the basis of this principle.

[0053] In the line layout shown in this Fig. 6, the overlap distances L between the resonators (distance between the lines 32 and 33 facing each other in the Y-direction), is shortened to about $\lambda/6$ in comparison with the line layout of Fig. 3. However, the distance between the line 32 and the line 33 in the Y-direction is same. Moreover, according to the layout of Fig. 6, the positional relation between the line 31 and the line 33 and the positional relation between the line 32 and the line 34 are not changed from the layout of Fig. 3. Moreover, according to the layout of Fig. 6, the length along which the line

31 and the line 32 face each other in the Y-direction and the length along which the line 33 and the line 34 face each other in the Y-direction are made longer than those of the layout of Fig. 3 in accordance with the arrangement that the overlap distances L between the resonators is shortened to about $\lambda/6$.

[0054] Fig. 7 shows the actual measurement result of a sample of a microstrip line circuit that adopts the line layout of this Fig. 6, the sample being actually produced on an experimental basis. Referring to the region P encircled by the dashed line in Fig. 7, it can be understood that the transmission coefficient S21 that represents the attenuation value at the frequency (24-GHz band) double that of the passband (12-GHz band) is improved to about 20 dB at worst. According to the measuring method, a probing pad of GSG (ground signal ground (coplanar)) obtained by adding a via (VIA) hole to the terminal end portion of the input/output microstrip line of the substrate produced on an experimental basis was provided, and the S parameters were measured by a network analyzer with a coplanar high-frequency probe, which had undergone LRM (line reflect match) calibration, applied to the probing pad.

[0055] In this embodiment, the example of the microstrip line circuit of the distributed element type has been described. However, the equivalent circuit of Fig. 1 shows the principle of this invention, and this invention can also be applied to a circuit of another high-frequency strip line, such as a concentrated constant circuit and a coplanar line.

[0056] According to the layouts shown in Figs. 3 and 6, the line widths of the terminal end portions 31b and 34a of the input/output lines 31 and 34 were made thinner in comparison with the portions 31a and 34b. This configuration was adopted for impedance matching.

(Second Embodiment)

[0057] Next, Fig. 8 shows the second embodiment of the filter circuit of this invention. This second embodiment has linear strip lines 52 and 53 and bent strip lines 51 and 54. The strip lines 51 and 52 face each other at a prescribed distance in the Y-direction, while the strip lines 52 and 53 face each other at a prescribed distance in the Y-direction. Moreover, the strip lines 53 and 54 face each other at a prescribed distance in the Y-direction.

[0058] This strip line 52 constitutes a first resonator 63, while the strip line 53 constitutes a second resonator 64. This first resonator 63 and the second resonator 64 correspond to the first resonator 3 and the second resonator 4, respectively, of the equivalent circuit of Fig. 1.

[0059] In this second embodiment, the line 51 has a portion 51a that constitutes an input port 61, a portion 51b and a portion 51c that faces a portion 52a of the line 52. The portion 51a and the portion 51b extend in the reverse Y-direction from both ends of the portion 51c. Moreover, this portion 51b has its terminal end portion

51b-1 that constitutes an open end 65 bent in the X-direction. Moreover, a basal end 51a-1 of this portion 51a faces a terminal end portion 53a-1 of the portion 53a of the line 53 at a prescribed distance in the X-direction.

[0060] These ends 51a-1 and 53a-1 constitute a fifth capacitance C15 that serves as a jump coupling capacitance, while the portion 51c and the portion 52a constitute a second capacitance C12. This fifth capacitance C15 and the second capacitance C12 correspond to the capacitances C5 and C2, respectively, of the equivalent circuit of Fig. 1.

[0061] Moreover, the portion 52b of the line 52 faces the portion 53a of the line 53 at a prescribed distance in the reverse Y-direction, while this portion 52b and the portion 53a constitute a first capacitance C11. Moreover, a portion 53b of the line 53 faces a portion 54c of the lines 54 at a prescribed distance in the Y-direction, while this portion 53b and the portion 54c constitute a third capacitance C13. This first capacitance C11 and the third capacitance C13 correspond to the capacitance C1 and the capacitance C3, respectively, of the equivalent circuit of Fig. 1.

[0062] Moreover, this line 54 has portions 54a and 54b that extend in the Y-direction from both ends of the portion 54c. This portion 54a faces an end portion 52b-1 of the portion 52b of the line 52 at a prescribed distance in the reverse X-direction, while this portion 54a and the end portion 52b-1 constitute a fourth capacitance C14. This fourth capacitance C14 corresponds to the capacitance C4 of the equivalent circuit of Fig. 1. Moreover, the portion 54b of this line 54 constitutes an output port 62.

[0063] According to this second embodiment, the portion 51a of the line 51 that constitutes the input port 61 is bent approximately perpendicularly to the portion 51c, the portion 51b of the line 51 is bent approximately perpendicularly from the portion 51c, and the open end 65 is bent approximately perpendicularly from the portion 51b. According to this embodiment, the dimension in the X-direction of the occupation region on a substrate 70 can be reduced. Therefore, this arrangement enables the layout in a free space that has a short dimension in the X-direction on the substrate. Therefore, area saving and downsizing can be achieved by reducing the dead space on the surface of the substrate 70. Therefore, monolithic integration can also be achieved by utilizing a free space on the MMIC of another circuit.

[0064] Fig. 9 shows the electromagnetic field simulation result of the circuit of this second embodiment. Referring to Fig. 9, it can be understood that the attenuation value at the frequency (24-GHz band) double that of the passband (12-GHz band) is improved to about 20 dB at worst, similarly to the frequency characteristic (Fig. 7) of the S parameters of the modification example of the aforementioned first embodiment. It is to be noted that S11 represents the reflection coefficient.

(Third Embodiment)

[0065] Next, Fig. 10 shows the third embodiment, which is a modification example of the aforementioned second embodiment. This third embodiment differs from the second embodiment in that the strip line 54 of the second embodiment of Fig. 8 is arranged in a state in which it is rotated by 180° around the Y-direction axis that extends through its center of the X-direction. That is, this third embodiment has a strip line 81 obtained by changing the coupling direction of the strip line 54 that constitutes the output line in Fig. 8.

[0066] According to this third embodiment, the dimension in the Y-direction at the end in the X-direction can be made smaller than that of the second embodiment. On the other hand, the dimension in the Y-direction at the center portion in the X-direction becomes greater than that of the second embodiment. Therefore, according to this third embodiment, the lines can be arranged in a free space of which the dimension in the Y-direction at both ends in the X-direction is shorter than that of the second embodiment.

[0067] Fig. 11 shows the electromagnetic field simulation result of the circuit of this third embodiment. Referring to Fig. 11, the transmission coefficient S21 that represents the attenuation value at the frequency (24-GHz band) double that of the passband (12-GHz band) is about 10 dB at worst. It is to be noted that S11 represents the reflection coefficient.

(Fourth Embodiment)

[0068] As described in connection with the aforementioned first embodiment and its modification example as well as the second embodiment and the third embodiment, according to the filter technique of this invention, there is little dead space, and area saving and downsizing can be achieved. Therefore, if the filter circuit of the first through third embodiments is monolithically integrated on a millimeter wave band MMIC, then the resulting device is suitable for use in performing spurious rejection.

[0069] In this case, paying attention to the objective spurious signal to be rejected and to the features that the attenuation value at the frequency double that of the passband is improved as shown in the frequency characteristic of Fig. 7 and constructing a high-frequency radio communication apparatus that employs the high-frequency filter circuit of the aforementioned embodiment as a filter for rejecting the higher harmonic spurious, a low-noise communication apparatus can be provided.

[0070] Moreover, paying attention to the point that the degree of steepness is improved in the vicinity of the higher region side of the passband by the transmission zero, as shown in the frequency characteristic (Fig. 2) according to the circuit simulation of the equivalent circuit shown in Fig. 1, it is appropriate to employ the high-frequency filter circuit of the aforementioned embodi-

ment as a filter for rejecting, for example, the image signal and the local signal.

[0071] In this fourth embodiment, the advantage exerted on the entire communication system when the filter circuit of the aforementioned embodiment of this invention is monolithically integrated on a millimeter wave band MMIC as an image rejection filter will be described.

[0072] First of all, in the equivalent circuit shown in Fig. 1, the passband was set at 60 GHz, and in order to achieve a frequency characteristic as shown in Fig. 12, there were the settings of first capacitance C1 = 0.02411 pF and second capacitance C2 = third capacitance C3 = 0.03949 pF. There were the further settings of fourth capacitance C4 = fifth capacitance C5 = 8.40370 pF. The characteristic impedance of the first resonator 3 and the second resonator 4 were set at 12 Ω, and the electrical lengths at a frequency of 60 GHz was set at 162 degrees.

[0073] According to the aforementioned circuit construction, as shown in the characteristic of Fig. 12, a bandwidth F1 of about 2.5 GHz was able to be secured in the 60-GHz band, or the passband. At the same time, there was successful formation of a steeply attenuated region in the image band F2, which is separated from the bandwidth F1 about 2.5 GHz, by virtue of the effect of the transmission zero P of the transmission coefficient S21. That is, according to the filter of this invention, a bandwidth of 2 to 3 GHz in the 60-GHz band can easily be secured. Moreover, since the filter circuit itself is small, it is easy to further integrate another circuit (amplifier circuit or the like) on an identical chip. Furthermore, the filter of this invention can easily be monolithically integrated with the amplifier circuit and the mixer circuit located before and behind it on an MMIC, and the filter itself is low-cost, microminiature and ultralight.

[0074] The above-mentioned characteristics are suitable for a multi-channel TV signal transmission system as reported by, for example, the reference document of K. Hamaguchi et al., "A Wireless Video Home-Link Using 60GHz Band: A Concept of Developed System", Proc. of EuMC, vol.1, pp.293-296, 2000.

[0075] Fig. 13 shows the construction of a monolithic microwave integrated circuit of the fourth embodiment of this invention. This fourth embodiment is provided with the filter circuit of the aforementioned modification example (Fig. 6) of the first embodiment as an image rejection filter circuit 84. This fourth embodiment constitutes the aforementioned multi-channel TV signal transmission system.

[0076] In this TV signal transmission system, a TV signal inputted to a mixer circuit 82 is mixed with a local oscillation signal from a local oscillator 83, and a signal from this mixer circuit 82 is inputted to an amplifier circuit 85 via the filter circuit 84, amplified and transmitted from an antenna 86.

[0077] As shown in Fig. 13, this fourth embodiment is the so-called "one-chip up-converter MMIC" in which the mixer circuit 82, the filter circuit 84, the amplifier circuit

85 and the local oscillator circuit 83 are all formed on an identical chip 81. It is to be noted that the chip 81 may be divided into about two MMIC chips according to the convenience of manufacturing and design. As described above, by adopting the filter circuit of the aforementioned first, second or third embodiment, the system can be entirely formed of an MMIC. Accordingly, in addition to the cost reduction, downsizing and weight reduction of the single unit of the filter circuit, there can also be obtained the synergistic effects of achieving the remarkable simplification of the whole system, the reduction in the number of components and the simplification of the manufacturing processes.

[0078] In this embodiment, the example of the micro-strip line circuit of the distributed element type has been described. However, the equivalent circuit of Fig. 1 shows the principle of this invention, and this invention can also be applied to a circuit of other high-frequency lines, such as a concentrated constant circuit and a coplanar line.

[0079] The invention being thus described, it will be obvious that the same may be varied in many ways. Such variations are not to be regarded as a departure from the spirit and scope of the invention, and all such modifications as would be obvious to one skilled in the art are intended to be included within the scope of the following claims.

Claims

1. A high-frequency filter circuit wherein
 - a first resonator (3) and a second resonator (4) are capacitively coupled to each other by a first capacitance (C1),
 - one terminal (3A) of the first resonator (3) is capacitively coupled to an input port (2) by a second capacitance (C2),
 - one terminal (4A) of the second resonator (4) is capacitively coupled to an output port (2) by a third capacitance (C3),
 - the other terminal (3B) of the first resonator (3) is capacitively coupled to the output port (2) by a fourth capacitance (C4), and
 - the other terminal (4B) of the second resonator (4) is capacitively coupled to the input port (1) by a fifth capacitance (C5).
2. The high-frequency filter circuit as claimed in claim 1, wherein
 - the first resonator (3) is comprised of a first line (32), the second resonator (4) is comprised of a second line (33),
 - the first resonator (3) and the second resonator (4) are coupled to each other by the first capacitance (C1) with the first line (32) facing the second line (33) at a prescribed distance,
 - the input port (1) is coupled to the first reso-

nator (3) by the second capacitance (C2) with a third line (31) that includes the input port (1) facing the first line (32),

the input port (1) is coupled to the second resonator (4) by the fifth capacitance (C5) with the third line (31) facing the second line (32),

the output port (2) is coupled to the second resonator (4) by the third capacitance (C3) with a fourth line (34) that includes the output port (2) facing the second line (33), and

the output port (2) is coupled to the first resonator (3) by the fourth capacitance (C4) with the fourth line (34) facing the first line (32).

3. The high-frequency filter circuit as claimed in claim 2, wherein
 - a length (L) of a confronting portion where the first line (32) faces the second line (33) is set within a range of 15% to 20% of a wavelength of a center frequency of a passband.
4. A monolithic microwave integrated circuit in which the high-frequency filter circuit (84) claimed in claim 1 is integrally formed on an identical semiconductor substrate (81) together with an amplifier circuit (85) or a mixer circuit (82).
5. A high-frequency radio communication apparatus, which has the high-frequency filter circuit claimed in claim 3 as a spurious rejection filter.

Fig. 1

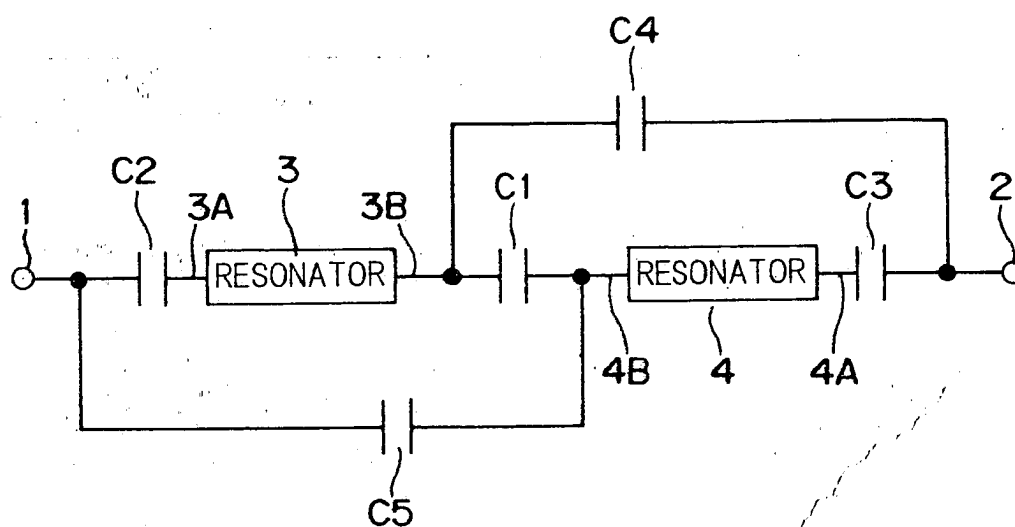


Fig. 2

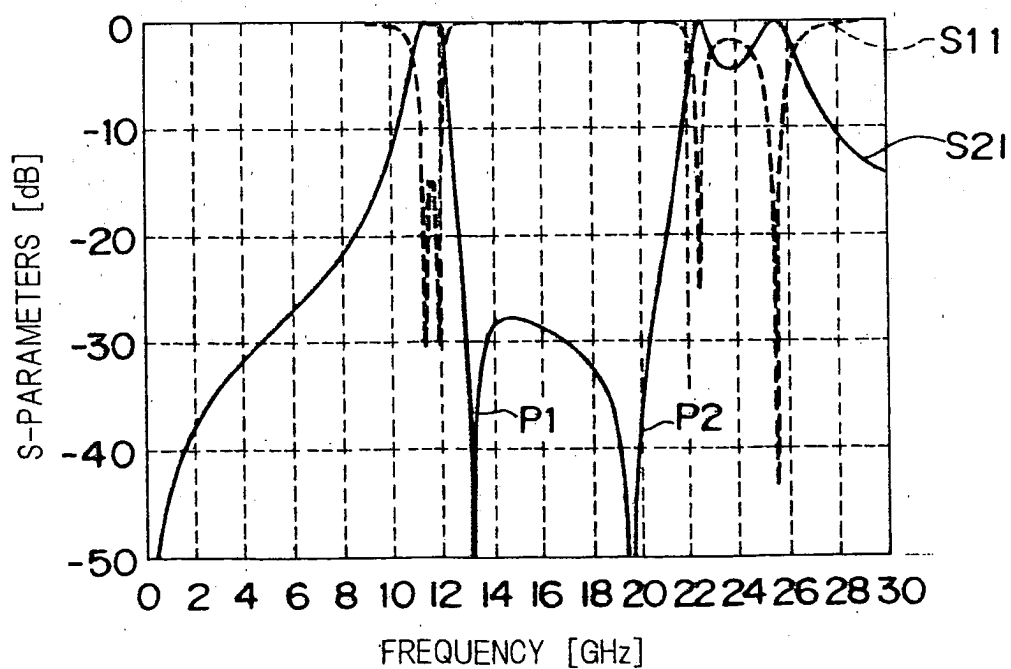


Fig.3

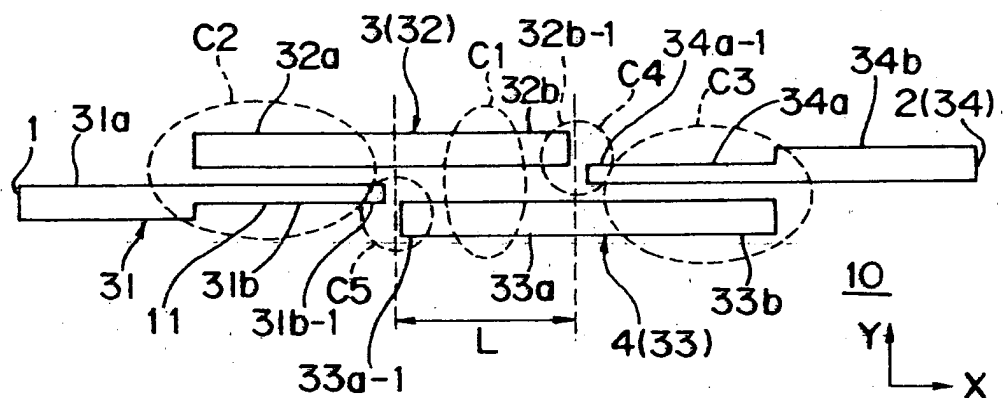


Fig.4

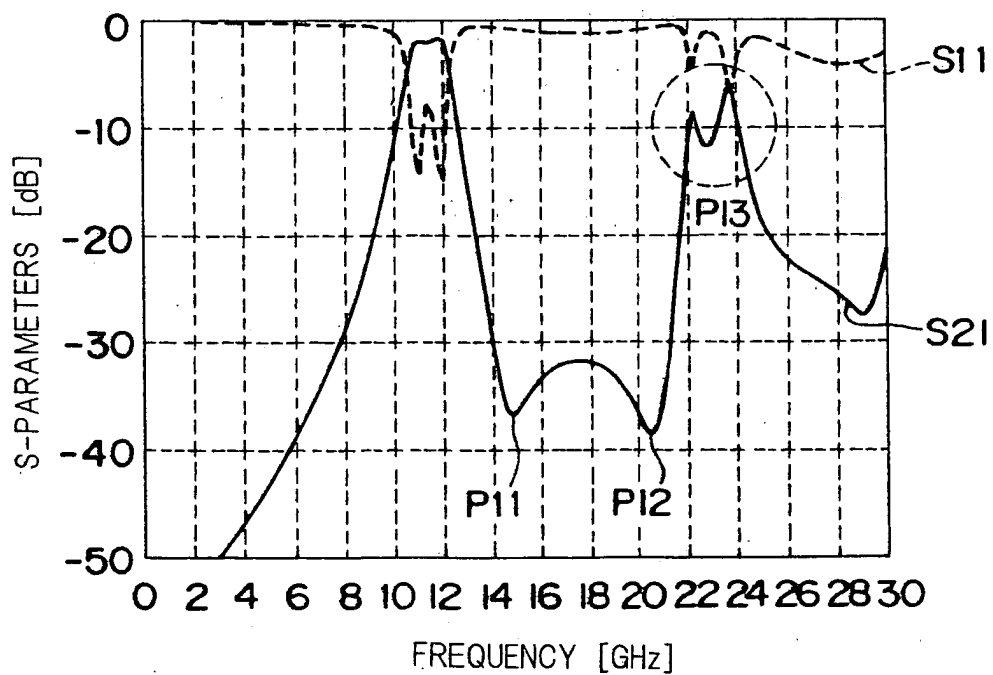


Fig.5

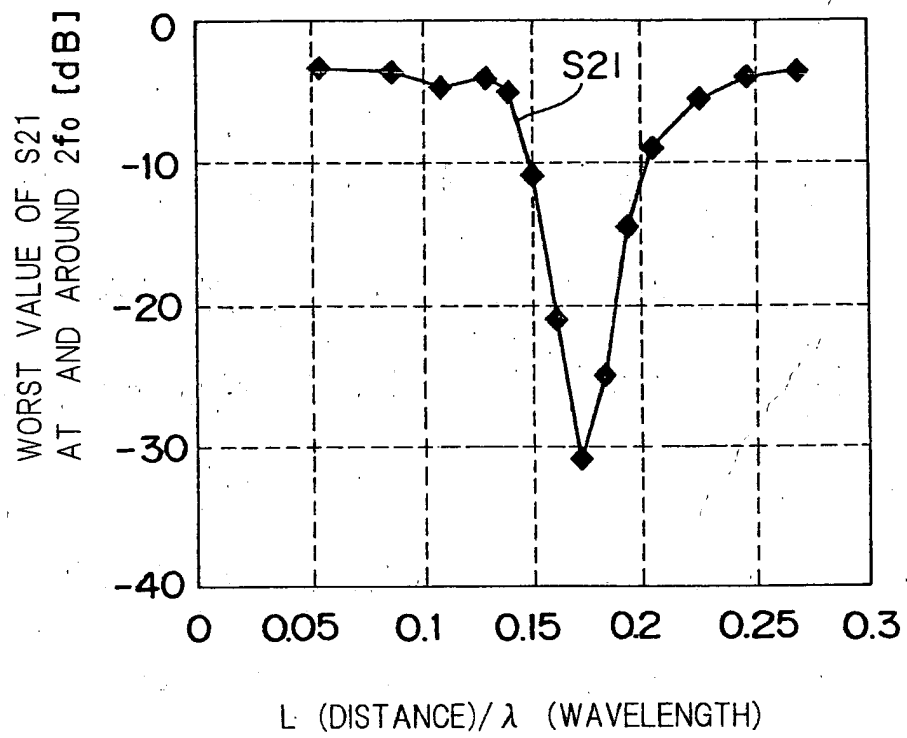


Fig.6

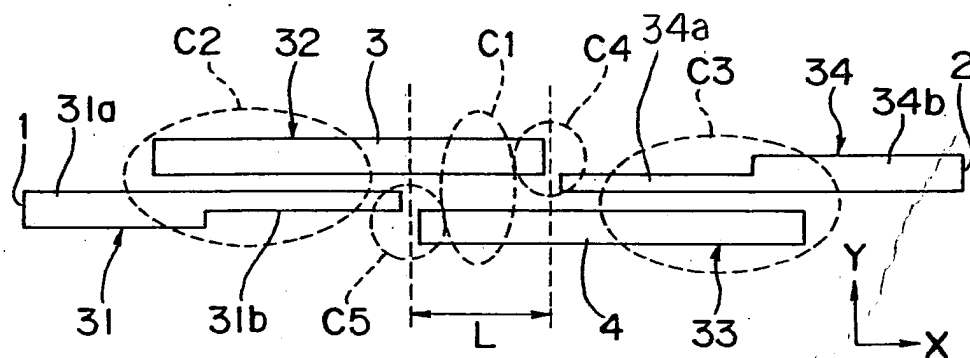


Fig.7

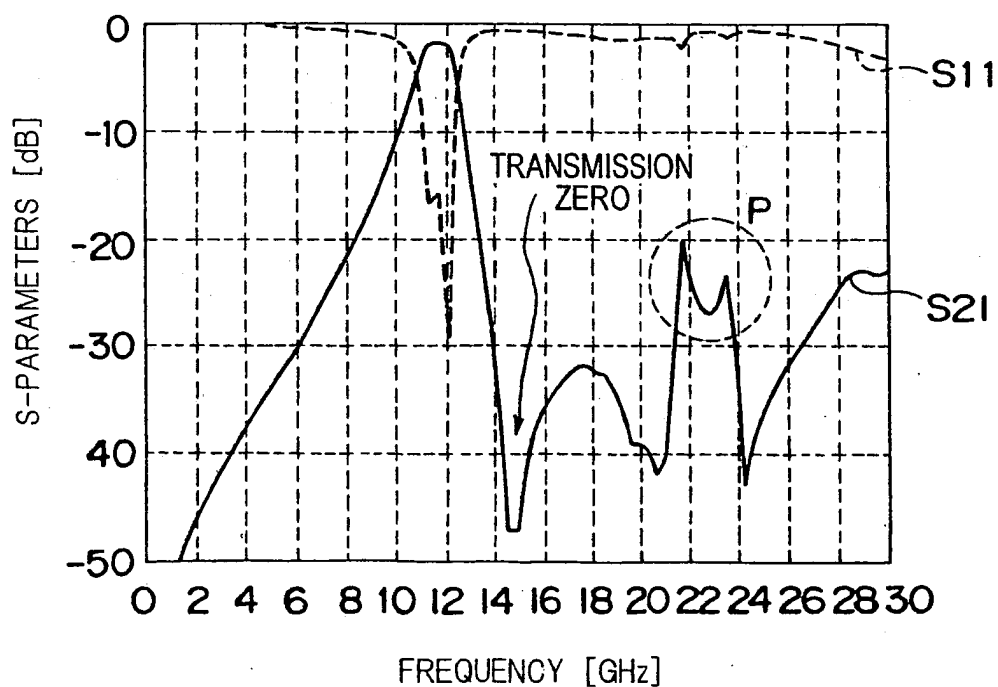


Fig.8

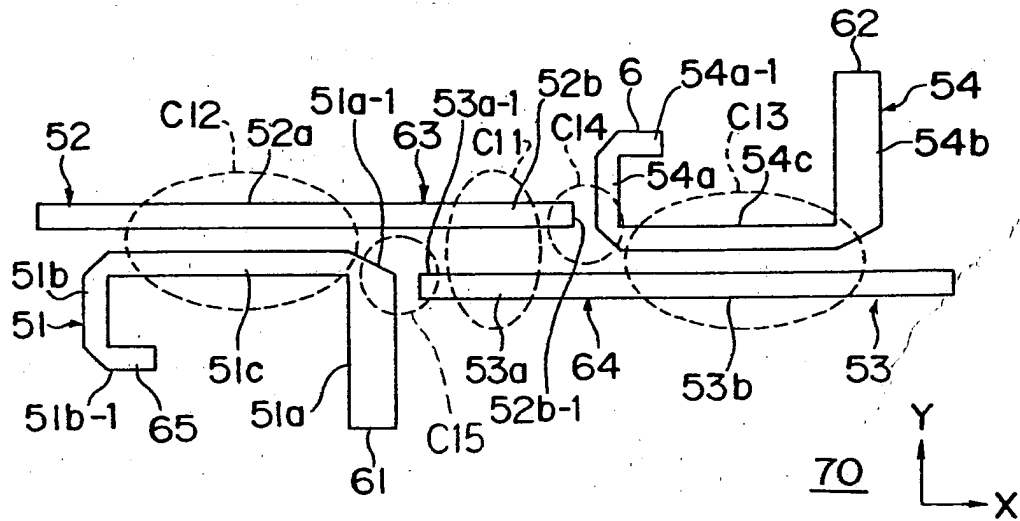


Fig.9

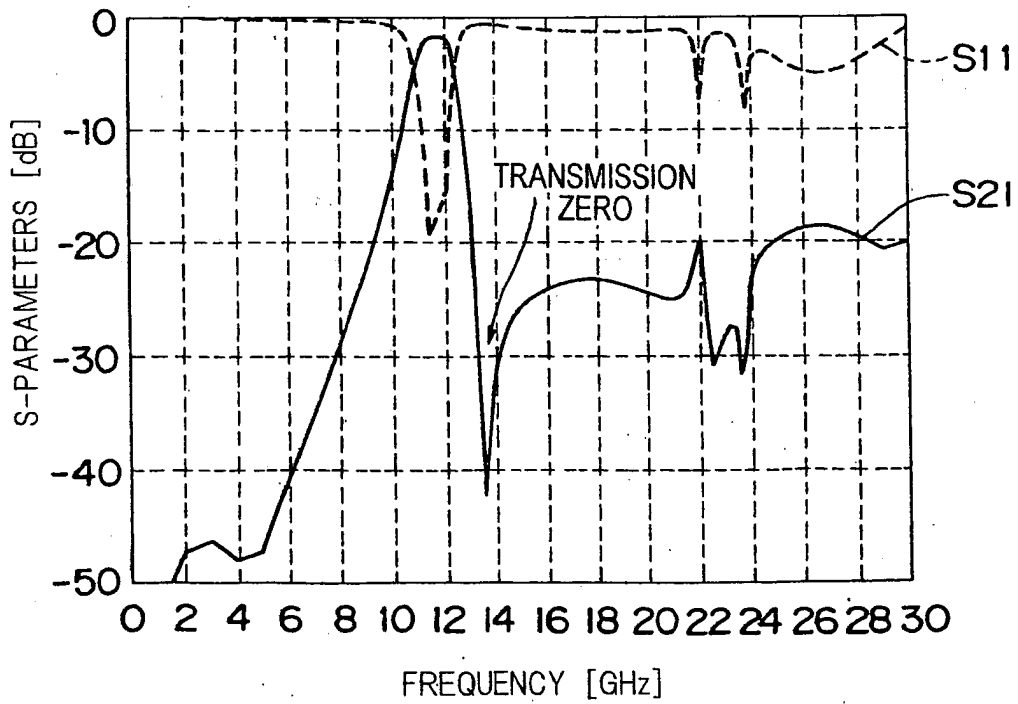


Fig. 10

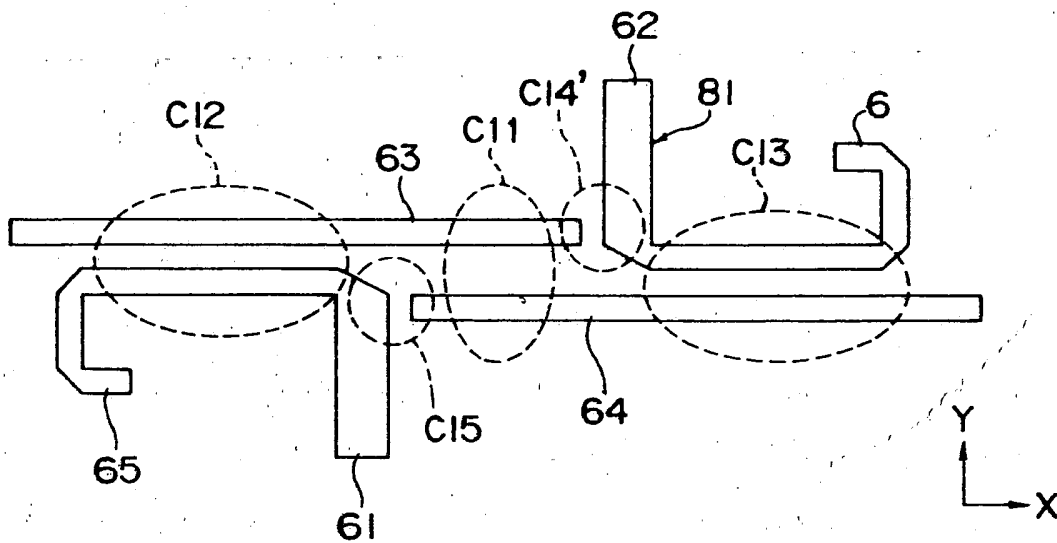


Fig. 11

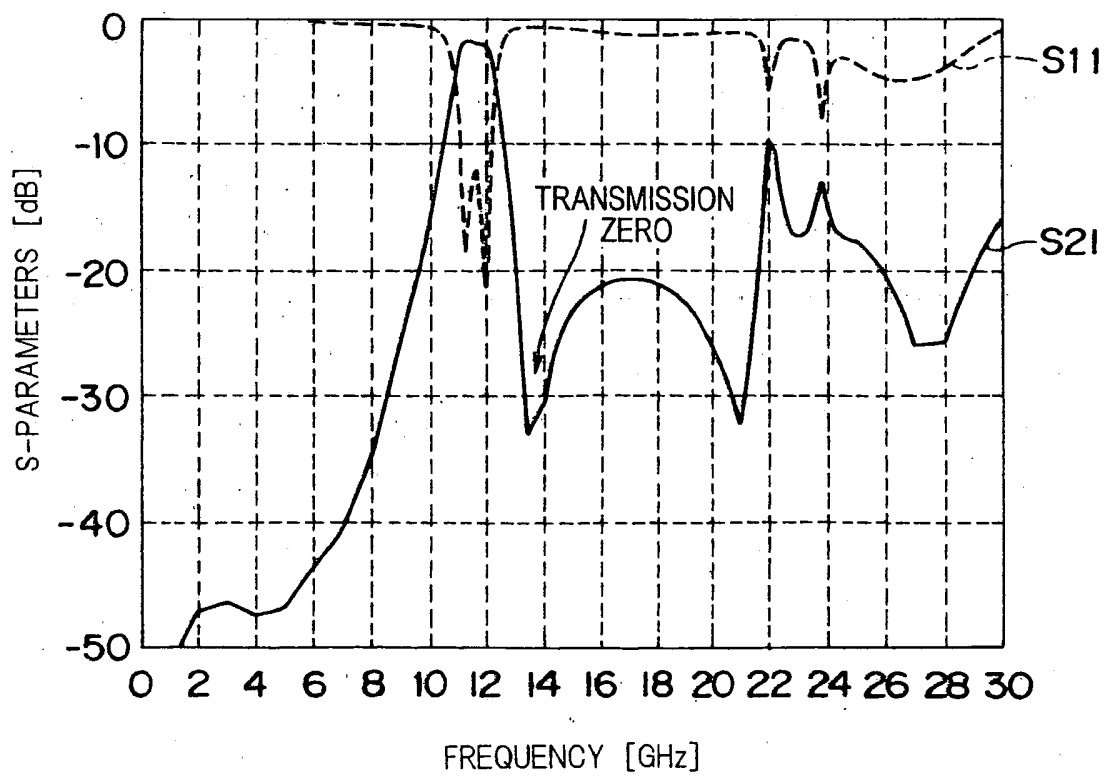


Fig. 12

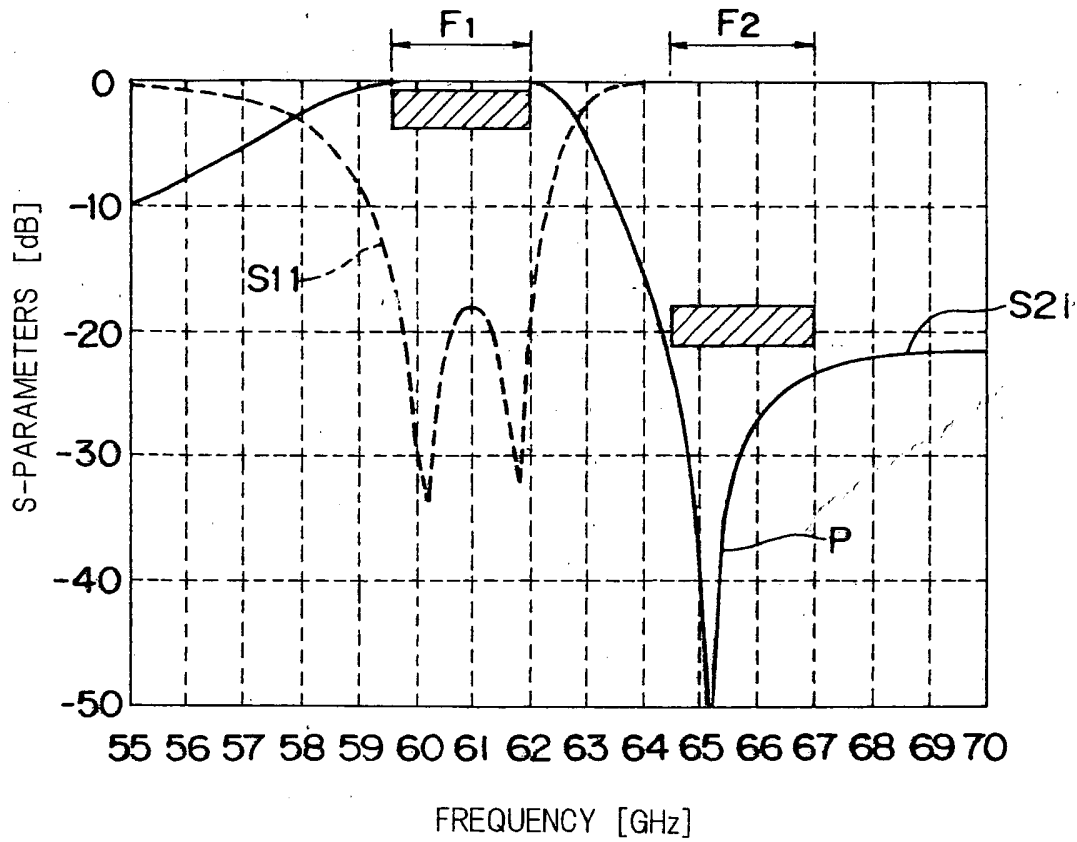


Fig. 13

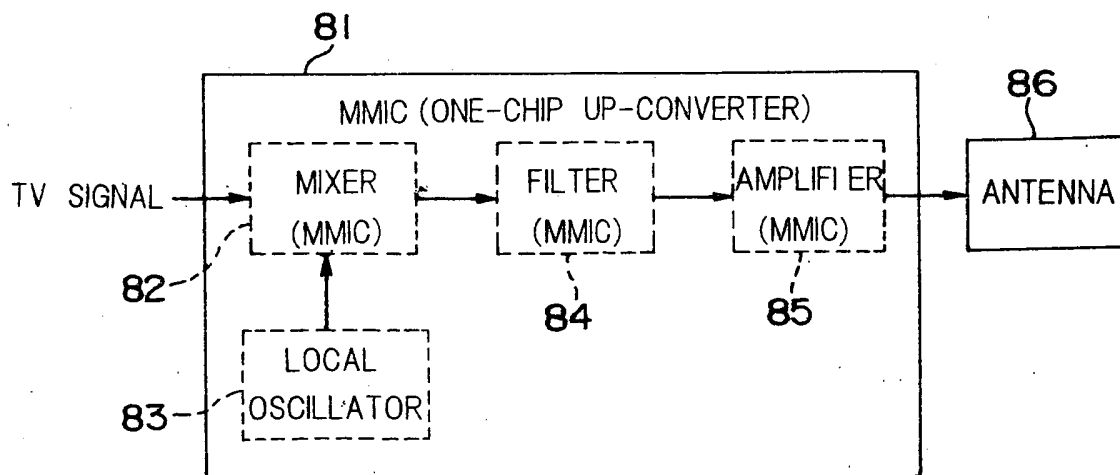


Fig. 14

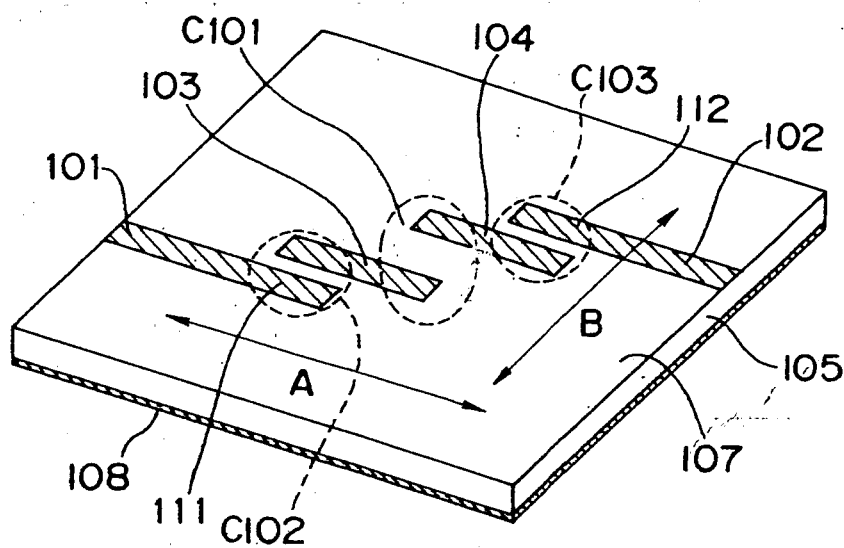


Fig. 15A

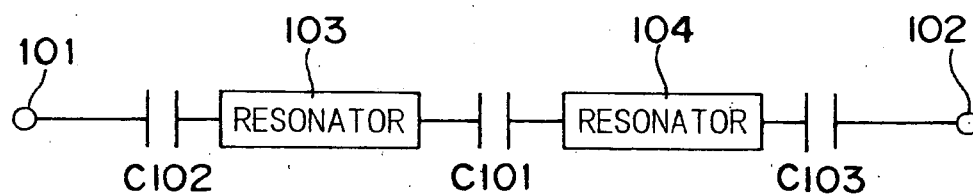


Fig. 15B

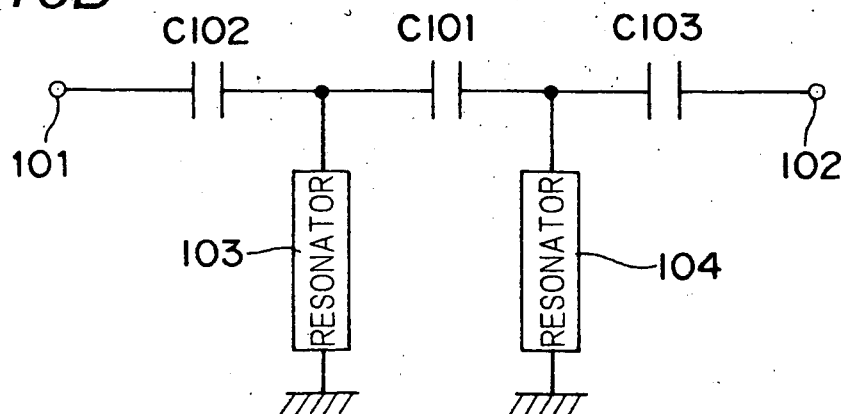


Fig.16

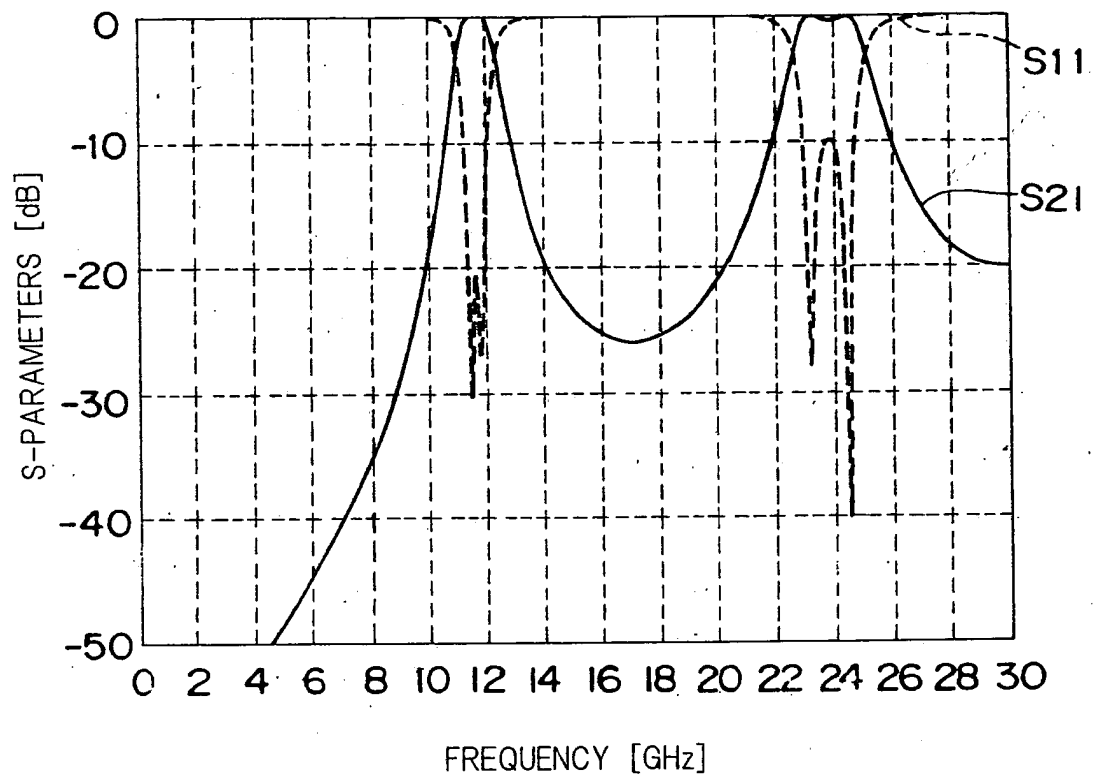


Fig. 17A

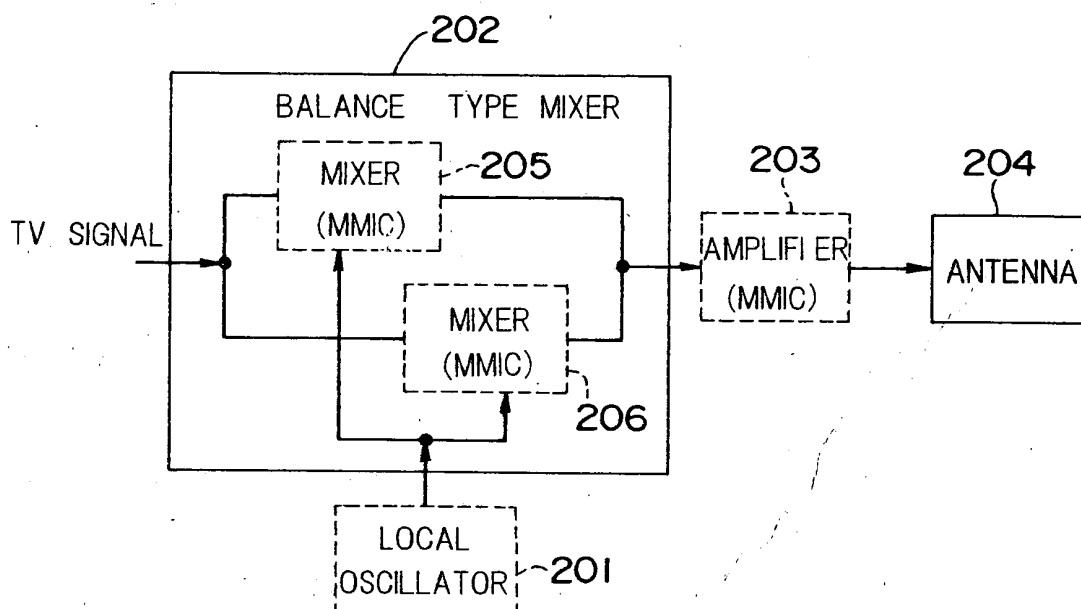
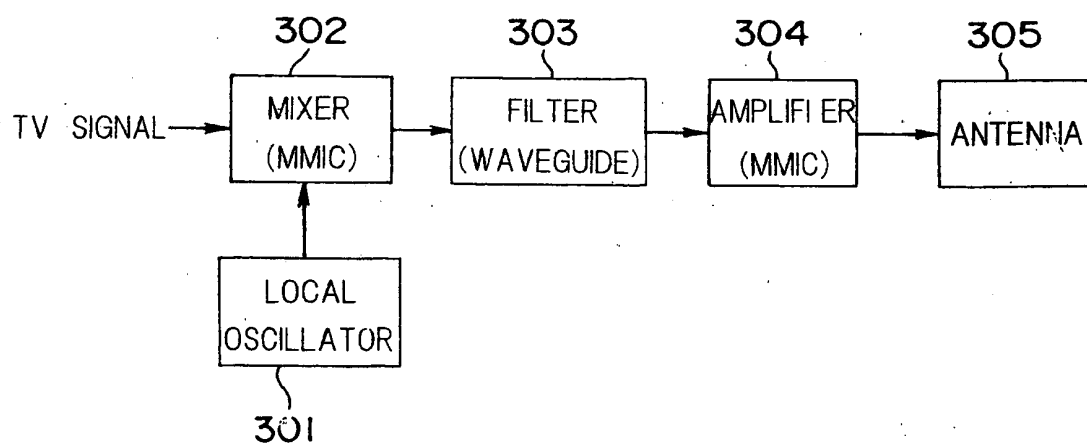


Fig. 17B





European Patent
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EUROPEAN SEARCH REPORT

Application Number
EP 03 00 2990

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A	US 3 605 045 A (RAMSBOTHAM ALAN J JR) 14 September 1971 (1971-09-14) * column 2, line 71 - column 3, line 33; figure 2 *	1	
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Place of search THE HAGUE		Date of completion of the search 29 April 2003	Examiner Den Otter, A
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