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Frequency doubling and Mixing Circuit

This invention relates to a frequency doubling and mixing circuit effecting a frequency doubling operation and a frequency mixing operation.

In case of effecting a frequency doubling operation and mixing operation, conventionally, an input signal frequency (local oscillator frequency in this example) is doubled by a frequency doubler 51 and sent through a band-pass filter 52 to a mixer 53 as shown in Fig. 1. The signal thus sent to the mixer 53 is mixed with another signal inputted thereto. With the arrangement as shown above, however, it is required that the frequency doubler 51 and the frequency mixer 53 are structured independently of each other and the filter 52 is provided therebetween as shown in Fig. 1. Consequently, such a problem has been arisen that the frequency characteristic cannot be provided on a broad-band basis and at the same time, the whole circuit itself is difficult to be integrated. In addition, the reduction

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in power consumption is difficult due to the fact that the number of circuit elements to be used becomes large, and the number of terminals is unavoidably increased because the filter is to be provided outside.

In consideration of the above-mentioned problems, an object of this invention is to provide a frequency doubling and mixing circuit in which the frequency doubling and mixing operations can be achieved by one unit of circuit.

(1) According to a first aspect of this invention, a frequency doubling and mixing apparatus of this invention has as a first set of differential transistor-pair which has emitters connected in common and an emitter size ratio of $K:1$ (K is larger than one), and a second set of differential transistor-pair which similarly has emitters connected in common and an emitter size ratio of $K:1$. Said first set of differential transistor-pair is biased by a constant electric current from a first constant current source and said second set of differential transistor-pairs is biased by a constant electric current from a second



constant current source. In addition, it is equipped with a differential amplifying circuit which is driven by a differential current between a common collector output of the transistors each having a larger emitter size of said first and second sets of differential transistor-pairs and a common collector output of the transistors each having a smaller emitter size of said first and second differential transistor-pairs as its driving power source. A signal to be doubled is supplied across the common base contact points of the transistors which are different in emitter size from each other of said first and second sets of differential transistor-pairs, and a mixing signal is supplied from a differential input terminal of said differential amplifying circuit. A circuit output is taken out from an output terminal of said differential amplifying circuit.

The frequency doubling and mixing circuit according to the first aspect of this invention may be arranged such that a resistor is inserted in series into an emitter of each of the transistors forming said first and second sets of differential transistor-pairs. As values of such emitter resistors to be inserted, it is preferable that, for



example, when one transistor has an emitter resistor value of R_E , the other transistor has an emitter resistor value of R_E/K in order to satisfy an emitter size ratio of $K:1$.

In the frequency doubling and mixing circuit in the first aspect of this invention, the frequency doubling circuit and the mixing circuit are connected through a current mirror circuit directly to each other, so that the doubling operation of a signal to be multiplied and the mixing operation of thus doubled signal with a mixing signal can be carried out by one unit of circuit. Also, such a circuit can be practically realized with a small number of elements to be used, and reductions in circuit scale and power consumption can be realized.

(2) According to a second aspect of this invention, in a frequency doubling and mixing circuit of this invention, a signal to be doubled is supplied from a first input terminal-pair and a signal to be mixed therewith is supplied from a second input terminal. In addition, it has two sets of differential transistor-pairs in which only one transistor of each pair has an emitter resistor, and the collector of the transistors each having an emitter resistor and the collector of the transistors each not



having an emitter resistor of these two sets of differential transistor-pairs are connected in common, and one of the bases of the transistors each having an emitter resistor and one of the bases of the transistors each not having an emitter resistor are connected in common to one input terminal of said first input terminal pair, and the other bases thereof are connected in common to the other input terminal of said first input terminal-pair. In addition, two transistors are provided, the collectors of which are connected respectively to emitters of the transistors each not having an emitter resistor of said two sets of differential transistor-pairs, and the bases of which are connected in common to said second input terminal.

In the frequency doubling and mixing circuit according to the second aspect, two sets of differential transistor-pairs which are different in emitter size may be used instead of the two sets of differential transistor-pairs in which one transistor only of each pair has an emitter resistor. In addition, using two sets of differential transistor-pairs which are different in emitter size, an emitter resistor having a resistance value of RE/K may be inserted into an emitter of a transistor with



a larger emitter size, and an emitter resistor having a resistance value of R_E may be inserted into an emitter of a transistor with a smaller emitter size, thus being capable of advantageously improving the amplitude level of an input signal VLO.

Further in addition, similar to the first aspect, in the second aspect, the frequency doubling operation of a signal to be doubled and the mixing operation of thus doubled signal with a mixing signal can be carried out by one unit of circuit.

(3) According to a third aspect of this invention, a frequency doubling and mixing circuit of this invention has a first input terminal-pair for receiving a signal to be doubled and a second input terminal-pair for receiving a mixing signal. In addition, it has a first differential amplifying circuit consisting of two sets of differential transistor-pairs in which only one transistor of each pair has an emitter resistor and the collector of the transistors each having an emitter resistor and the collector of the transistors each not having an emitter resistor of these two sets of differential transistor-pairs are connected in common, one of the bases of the



transistors each having an emitter resistor and one of the bases of the transistors each not having an emitter resistor are connected in common to one input terminal of said first input terminal pair and the other bases thereof are connected in common to the other input terminal thereof. In addition, the emitters of the transistors of each pair are respectively connected to constant current sources. Further in addition, it has a second differential amplifying circuit consisting of differential transistor-pair in which one base thereof is connected to one input terminal of said second input terminal-pair and the other base thereof is connected to the other input terminal of said second input terminal-pair. The second differential amplifying circuit is driven by a current mirror circuit in response to a differential current obtained from each output of said first differential amplifying circuit.

With the frequency doubling and mixing circuit according to the third aspect, it is preferable to have such a circuit that can control said current mirror circuit so as to drive said second differential amplifying circuit by an electric current obtained by subtracting a constant value of direct current from said differential current. Such control



circuit is preferable to be consisted of a constant current source which is inserted in parallel into one of the transistors forming said current mirror circuit.

In the circuit according to the third aspect, said first differential amplifying circuit constitutes a frequency doubling circuit and the second differential amplifying circuit constitutes a mixing circuit, which are directly connected through a current mirror circuit with each other. As a result, even in this case, the frequency doubling operation of a signal to be doubled and a mixing operation of a signal to be mixed therewith can be carried out by one unit of circuit.

In case of being equipped with a control circuit of said current mirror circuit, a constant value of direct current is subtracted from a differential current obtained from respective outputs of the first differential amplifying circuit as the frequency doubling circuit, and the second amplifying circuit as the mixing circuit is driven in response to the current thus obtained by the above subtraction, so that a double frequency component extremely superior in distortion factor becomes predominant as the driving current of the second differential amplifying



circuit, thus being effective to improve the conversion gain of the mixing circuit.

BRIEF DESCRIPTION OF THE DRAWINGS

Fig. 1 is a circuit diagram of a frequency doubling and mixing circuit of the prior art.

Fig. 2 is a circuit diagram of a frequency doubling and mixing circuit according to a first embodiment of this invention.

Fig. 3 is a diagram showing a characteristic of the circuit shown in Fig. 2.

Fig. 4 is a circuit diagram of a frequency doubling and mixing circuit according to a second embodiment of this invention.

Fig. 5 is a diagram showing a characteristic of the circuit shown in Fig. 4.

Fig. 6 is a circuit diagram of a frequency doubling and mixing circuit according to a third embodiment of this invention.

Fig. 7 is a diagram showing a characteristic of the circuit shown in Fig. 6.

Fig. 8 is a circuit diagram of a frequency doubling and



mixing circuit according to a fourth embodiment of this invention.

Fig. 9 is a circuit diagram of a frequency doubling and mixing circuit according to a fifth embodiment of this invention.

Fig. 10 is a circuit diagram of a frequency doubling and mixing circuit according to a sixth embodiment of this invention.

Fig. 11 shows a relation of a collector current and an input signal (voltage) V_{LO} of the circuit shown in Fig. 9.

Fig. 12 is a circuit diagram of a frequency doubling and mixing circuit according to a seventh embodiment of this invention.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

Preferred embodiments of this invention will be described below while referring to the drawings attached.

[First Embodiment]

A frequency doubling and mixing circuit according to a first embodiment of this invention is shown in Fig. 2. In Fig. 2, a first set of differential transistor-pair Q1 and



Q2 respectively has emitters which are connected in common and have an emitter size (emitter area) ratio of $k:1$ ($k>1$). To the common connection point of the emitters of the transistors Q1 and Q2 is connected a constant current source I0. A second set of differential transistor-pair Q3 and Q4 respectively has emitters which are connected in common and have the same emitter size ratio as of the transistors Q1 and Q2, or $k:1$ ($k>1$). The emitter common connection point of the transistors Q3 and Q4 also is connected to the constant current source I0. A signal (local oscillator signal in this embodiment) VL0 is applied through input terminal - pair 1 and 2 across the common base of the transistors Q1 and Q4 and that of the transistors Q2 and Q3. In addition, a differential current between a common collector output of the transistors Q1 and Q3 and that of the transistors Q2 and Q4 is derived by a current mirror circuit consisting of transistors Q5 to Q8. The differential current output derived by this current mirror circuit is taken out from another current mirror circuit consisting of transistors Q9 and Q10 to drive a differential amplifier consisting of transistors Q11 and Q12.

The transistors Q11 and Q12 is a differential



transistor-pair having emitters connected in common. Across the bases of the transistors Q11 and Q12, an input signal VIN to be mixed is applied through output terminals 3 and 4. A differential output of the differential amplifier consisting of the transistors Q11 and Q12 is derived by a current mirror load through the transistors Q13 and Q14, and a circuit output V0 is obtained by an emitter follower circuit consisting of a transistor Q15 and a resistance RL.

When a DC amplification factor of transistor is expressed by α , collector currents I_{c1} to I_{c4} of the transistors Q1 to Q4 become as follows; where $V_T = kT/q$, k is Boltzman's constant, T is absolute temperature, q is the charge of an electron, and K is an emitter size ratio.

$$I_{c1} = \frac{\alpha I_0}{\{1 + (1/K) \exp (-V_{L0}/V_T)\}} \quad \dots(1)$$

$$I_{c2} = \frac{\alpha I_0}{\{1 + K \exp (V_{L0}/V_T)\}} \quad \dots(2)$$

$$I_{c3} = \frac{\alpha I_0}{\{1 + (1/K) \exp (V_{L0}/V_T)\}} \quad \dots(3)$$

$$I_{c4} = \frac{\alpha I_0}{\{1 + K \exp (-V_{L0}/V_T)\}} \quad \dots(4)$$

From Eqs. (1) to (4), therefore, the sum I_p of the



collector currents I_{c1} and I_{c3} and the sum I_q of the collector currents I_{c2} and I_{c4} can be expressed as follows;

$$I_p = I_{c1} + I_{c3}$$

$$= \alpha \cdot I_0 \left[\frac{1}{1 + (1/K) \exp(-V_{L0}/V_T)} + \frac{1}{1 + (1/K) \exp(V_{L0}/V_T)} \right] \quad \dots\dots(5)$$

$$I_q = I_{c2} + I_{c4}$$

$$= \alpha \cdot I_0 \left[\frac{1}{1 + K \exp(-V_{L0}/V_T)} + \frac{1}{1 + K \exp(V_{L0}/V_T)} \right] \quad \dots\dots(6)$$

Hence, a difference $\Delta I_{p,q}$ of I_p and I_q becomes as follows; where $V_K = V_T \cdot (\ln K)$

$$\begin{aligned} \Delta I_{p,q} &= I_p - I_q = (I_{c1} - I_{c2}) - (I_{c4} - I_{c3}) \\ &= \alpha \cdot I_0 \cdot \tanh[(V_{L0} + V_K)/2V_T] \\ &\quad - \alpha \cdot I_0 \cdot \tanh[(V_{L0} - V_K)/2V_T] \\ &= \alpha \cdot I_0 \cdot \{ \tanh[(V_{L0} + V_K)/2V_T] \\ &\quad - \tanh[(V_{L0} - V_K)/2V_T] \} \quad \dots\dots(7) \end{aligned}$$

In this case, $\Delta I_{p,q}$ becomes an even function with respect to V_{L0} .

Next, when $x \ll 1$, $\tanh x$ can be expanded in series as;



$$\tanh x = x - (x^3/3) + \dots \dots \dots (8)$$

From this,

$$\begin{aligned} \Delta I_{p,q} &= (\alpha \cdot I_0/4V_T) \cdot (\ln K) \cdot \\ &\quad \times \{1 - (\ln K)^2/12 - [1/4(V_T)^2] \cdot V_{L0}^2 \dots \dots \} \\ &\dots \dots (9) \end{aligned}$$

Here, Eq. (9) can be approximated as follows;

$$\begin{aligned} \Delta I_{p,q} &\approx (\alpha \cdot I_0/4V_T) \cdot (\ln K) \cdot \\ &\quad \times \{1 - (\ln K)^2/12 - [1/4(V_T)^2] \cdot V_{L0}^2\} \\ &\dots \dots (10) \end{aligned}$$

Accordingly, the $\Delta I_{p,q}$ can be approximately expressed in terms of an equation involving only the term of the square of input signal V_{L0} , and if the DC component is removed, it is found that almost of the frequency components contained in the differential current $I_{p,q}$ are of a frequency of $2 f_{L0}$ as compared with an input signal frequency of f_{L0} . The reason of this is that if the signal is a sine wave, then,

$$2 \sin^2 A = 1 - \cos 2A$$

thus being obtainable a frequency component double in frequency as compared with the DC component. As shown above, a circuit consisting of the transistors Q1 to Q8 becomes a frequency multiplying circuit.



In this case, however, if the $I_{p,q}$ when it is no signal ($V_{L0} = 0$) is expressed as $ID_{p,q}$, ID becomes a direct current, from Eq. (7), the following equation can be obtained;

$$ID_{p,q} = 2\alpha \cdot I_0 \cdot \tanh[(\ln K)/2] \quad \dots\dots\dots(11)$$

The value thus obtained is identical to the value obtained when $V_{L0} = 0$ in Eq. (10). This means that $\Delta I_{p,q}$ can be obtained by superposing an AC component with a frequency of $2 f_{L0}$ onto the DC component $ID_{p,q}$, the characteristics of which are shown in Fig. 3.

Referring to Fig. 3, for example, when it was simulated with such parameters that were set as $K = 9$, $V_{L0} = 60$ mV and $f_{L0} = 1$ kHz, the frequency component (f_{L0}) of an input signal became a level of -60 dB or less on the output side as compared with the twofold frequency component ($2 f_{L0}$) of the input signal, resulting in being obtainable as very good distortion factor property. As a result, no need to use filter results on the output side, so that the wide operating input frequency range can be obtained.

Next, the difference $\Delta I_{p,q}$ between the collector current sums I_p and I_q is substantially equal to a collector output current of the transistor Q_8 through the current



mirror circuit consisting of the transistors Q5, Q6, Q7 and Q8. This current becomes a driving current source of the differential amplifier consisting of the transistors Q11 and Q12 through the current mirror circuit consisting of the transistors Q9 and Q10. Accordingly, an output current I_{OUT} of this differential amplifier can be expressed as follows:

$$I_{OUT} = I_{C11} - I_{C12} = \alpha \Delta I_{p,q} \cdot \tanh(V_{IN}/2V_T) \dots\dots(12)$$

where I_{C11} and I_{C12} are collector currents of the transistor power amplifying circuits Q11 and Q12, respectively.

Here, when $x \ll 1$, $\tanh x$ can be expanded in series as:

$$\tanh x = x - (x^3/3) + \dots\dots\dots(13)$$

so that if V_{IN} is extremely smaller than $2V_T$, I_{OUT} can be approximated as;

$$\begin{aligned} I_{OUT} &= \alpha^2 \cdot I_0 \cdot \{ \tanh[(V_{L0} + V_K)/2V_T] - \tanh[(V_{L0} - V_K)/2V_T] \} \\ &\quad \times \tanh(V_{IN}/2V_T) \dots\dots(14) \\ &= \alpha^2 \cdot I_0 \cdot (\ln K)/4V_T \cdot \{ [1 - (\ln K)^2/12] - (V_{L0})^2/4(V_T)^2 \} \\ &\quad \times (1/2V_T) \cdot \{ V_{IN} - [1/12(V_T)^2] \cdot (V_{IN})^3 \} \dots\dots \\ &= \alpha^2 \cdot I_0 \cdot (\ln K)/8(V_T)^2 \\ &\quad \times \{ [1 - (\ln K)^2/12] \cdot V_{IN} - (V_{L0})^2 \cdot (V_{IN})/4(V_T)^2 \\ &\quad - [1 - (\ln K)^2/12] \cdot (V_{IN})^3/12(V_T)^2 \\ &\quad + (V_{L0})^2 \cdot (V_{IN})^3/48(V_T)^4 \} \dots\dots \\ &\quad \dots\dots(14) \end{aligned}$$



From Eq. (14), the product of $(VLO)^2$ and VIN , that is, $(VLO)^2 \cdot (VIN)$ can be obtained. By this product, the frequency component contained in the output current $IOUT$ is predominated by $(2 fLO + fIN)$ and $(2 fLO - fIN)$ or $(fIN - 2 fLO)$. This is because the product of $\sin^2 A$ and $\sin B$ becomes the product of $\cos 2A$ and $\sin B$, and the product of $\cos 2A$ and $\sin B$ becomes the sum of $\sin (2A+B)$ and $\sin (2A-B)$ or $\sin (B-2A)$. As a result, the circuit output $IOUT$ is outputted having mixed the double wave signal of input signal VLO with the signal of VIN .

As explained above, the circuit shown in Fig. 2 comprises a frequency doubling circuit and mixing circuit of an input signal VLO , which are connected directly with each other through a current mirror circuit. In addition, by applying an output current of the frequency doubling circuit to a driving current source of the mixing circuit, a bias circuit of the mixing circuit does not need.

[Second Embodiment] -

Fig. 4 shows a second embodiment of this invention. In Fig. 4, the members having the same functions as those shown in Fig. 2 are expressed as the same reference numerals and



letters. The circuit of this embodiment has resistors inserted directly into respective emitters of differential transistor-pairs Q1 to Q4. Referring to these emitter resistors, for example, in case of the transistors Q1 and Q2, the emitter ratio is K:1, and the emitter resistor of the transistor Q1 is made RE/K when the emitter resistor of the transistor Q2 is made RE . In the case of the transistors Q3 and Q4, the same selection manner as shown above is applied. Fig. 5 shows the characteristics obtained when the emitter resistors are inserted as above. As seen from Fig. 5, the amplitude level of the input signal V_{LO} changes depending on the value of an emitter resistor to be inserted which means that the circuit of this embodiment can be used at any input signal level by appropriately selecting the value of the emitter resistor to be inserted.

As explained above, in the first and second embodiments, the frequency doubling circuit is connected directly to the mixing circuit using a current mirror circuit, so that a filter or the like does not need to be used, and expansion of frequency band and integration of circuit can be made easy. In addition, the distortion



factor property of an output of the frequency doubling circuit is outstanding good, and at the same time, a bias circuit of the mixing circuit can be advantageously eliminated because these circuits are connected directly with each other by a current mirror circuit.

[Third Embodiment]

Fig. 6 is for explaining a third embodiment of this invention. In Fig. 6, 21 and 22 show a first input terminal-pair to be applied with an input signal V_{LO} to be multiplied, 23 is a second input terminal, and an input signal V_{IN} as a mixing signal is applied across the input terminal 23 and the earth. In addition, Q 21 and Q 22, and Q 23 and Q 24 are differential transistor-pairs which have respective emitters connected in common, in which each of the transistors Q 22 and Q 23 of respective differential transistor-pairs has an emitter resistor R_E . Between these two sets of differential transistor-pairs, the collector of the transistors Q 22 and Q 23 each having an emitter resistor and the collector of transistors Q 21 and Q 24 each not having an emitter resistor are connected in common, the base of the transistor Q 21 having an emitter resistor and



the base of the transistor Q23 not having an emitter resistor are connected in common to the one input terminal 21 of the first input terminal pair, and the base of the transistor Q22 having an emitter resistor and the base of the transistor Q24 not having an emitter resistor are connected in common to the other input terminal 22 thereof. The collector of the transistors Q21 and Q23 is connected directly to a power source VCC, the collector of the transistors Q22 and Q24 is connected to an output terminal 25 and at the same time, connected through a resistor RL to the power source VCC.

Transistors Q25 and Q26 respectively constitute known constant current sources, which are respectively connected to the emitters of the transistors Q21 and Q24 each not having an emitter resistor of said two sets of differential transistor-pair, and the bases of which are connected in common to the second input terminal 23.

With the circuit structured as above, if the base voltages of the differential transistor-pair Q21 and Q22 are respectively expressed as $V_{BE\ 21}$ and V_{BE22} , and the current amplification factor is expressed as α , the following equation is established as;



$$V_{L0} + V_{BE21} + (I_{C22}/\alpha)R_E - V_{BE22} = 0 \quad \dots\dots(15)$$

Here, if $V_T = kT/q$, then, $V_{BE21} \gg V_T$ and $V_{BE22} \gg V_T$, and if the collector currents of the transistors Q21 and Q22 are respectively expressed as I_{C21} and I_{C22} , and the saturation currents of the transistors Q21 and Q22 are respectively expressed as I_{S21} and I_{S22} , then, there exist the following relations between V_{BE21} , V_{BE22} and V_T as;

$$V_{BE21} = V_T \cdot \ln (I_{C21}/I_{S21}) \quad \dots\dots(16)$$

$$V_{BE22} = V_T \cdot \ln (I_{C22}/I_{S22}) \quad \dots\dots(17)$$

In Eqs. (16) and (17), supposing that $I_{S21} = I_{S22}$, Eq. (15) may be expressed as follows;

$$V_{L0} + V_T \cdot \ln (I_{C21}/I_{C22}) + (I_{C22}/\alpha) R_E = 0 \quad \dots\dots(18)$$

Between the collector current I_{C25} of the transistor Q25, collector current I_{C21} of the transistor Q21 and collector current I_{C22} of the transistor Q22, there exists the following relation as;

$$\alpha \cdot I_{C25} = I_{C21} + I_{C22} \quad \dots\dots(19)$$

Hence, Eq. (18) becomes as follows;

$$V_{L0} + V_T \cdot \ln [(\alpha \cdot I_{C25}/I_{C22}) + 1] + (I_{C22}/\alpha) R_E = 0 \quad \dots\dots(20)$$

Here, in order to obtain the gradient of the collector current I_{C22} to the input voltage V_{L0} as a signal to be



multiplied, by differentiating I_{c22} with respect to V_{L0} , the following can be obtained as;

$$\frac{d I_{c22}}{d V_{L0}} = - \frac{1}{V_T \frac{I_{c22}(\alpha \cdot I_{c25} - I_{c22})}{\alpha \cdot I_{c25}} + \frac{R_E}{\alpha}} \quad \dots\dots(21)$$

Here, the point where the differential value (absolute value) is maximized satisfies the following;

$$I_{c22} = I_{c21} = (\alpha \cdot I_{c25})/2 \quad \dots\dots(22)$$

Hence, the maximum value is expressed by the following equation as;

$$\left| \frac{d I_{c22}}{d V_{L0}} \right| I_{c22} = (1/2) \alpha \cdot I_{c25} = \frac{\alpha \cdot I_{c25}}{4V_T + R_E \cdot I_{c25}} \quad \dots\dots(23)$$

In this case, the input voltage V_{L0} becomes as follows;

$$V_{L0} = -(1/2)R_E \cdot I_{c25} \quad \dots\dots(24)$$

In this case of referring to the differential transistor-pair consisting of the transistors Q23 and Q24, it can be determined in the same manner as above. That is, the derivation may be achieved by inverting the polarity of the input voltage V_{L0} in the above-mentioned method of derivation, and it is omitted to be shown here.



Next, if the transistors Q25 and Q26 each has a saturation current of I_s , the collector currents I_{c25} and I_{c26} thereof are given as:

$$\begin{aligned} I_{c25} = I_{c26} &\approx I_s [\exp (V_{IN}/V_T) - 1] \\ &\approx I_s \cdot \exp (V_{IN}/V_T) \end{aligned} \quad \dots\dots\dots(25)$$

Then, if V_F is a constant voltage value, V_{RF} is an AC signal value and an input signal voltage V_{IN} as a mixing signal is expressed as

$$V_{IN} = V_F + V_{RF} \quad \dots\dots\dots(26)$$

Eq. (25) becomes as follows;

$$\begin{aligned} I_{c25} = I_{c26} &= I_s \cdot \exp [(V_F + V_{RF})/V_T] \\ &\approx I_0 \cdot \exp (V_{RF}/V_T) \end{aligned} \quad \dots\dots\dots(27)$$

$$\text{where, } I_0 = I_s \cdot \exp (V_F/V_T) \quad \dots\dots\dots(28)$$

As a result, the DC values of the collector currents I_{c25} and I_{c26} become as follows;

$$I_{c25} = I_{c26} = I_0 \quad \dots\dots\dots(29)$$

Fig. 7 shows a relation of the input voltage V_{I0} and each of the collector currents I_{c21} , I_{c22} , I_{c23} and I_{c24} of respective transistors Q21, Q22, Q23 and Q24 when $V_{RF} = 0$.

In addition, $R_E \cdot I_{c25} = 16 V_T$ is supposed. In Fig. 7, I_1 is the sum of the collector current I_{c21} of the transistor Q21 and the collector current I_{c24} of the transistor Q24, and I_2



is the sum of the collector current I_{c22} of the transistor Q22 and the collector current I_{c23} of the transistor Q23, that is,

$$I_1 = I_{c21} + I_{c24} \quad \dots\dots\dots(30)$$

$$I_2 = I_{c22} + I_{c23} \quad \dots\dots\dots(31)$$

In this case, it is found that each of the currents I_1 and I_2 becomes a differential current and has a both wave rectification property for the input voltage V_{L0} . As a result, by appropriately selecting the emitter resistor R_E , such a characteristic can be obtained for each of the currents I_1 and I_2 that is considerably approximated to the square characteristic of the input voltage V_{L0} , which means that a double frequency $2 f_{L0}$ is resulted as compared with the input frequency f_{L0} , and it becomes predominant as the frequency component thereof. Also, when each of the currents I_1 and I_2 has the square characteristic closely with respect to the input voltage V_{L0} , all the frequency components thereof become the $2 f_{L0}$ component only. In addition, from Fig. 7, it is found that each of the currents I_1 and I_2 is proportional to the collector current I_{c25} of the transistor Q25.

Next, explanations will be made below on the mixing



operation. Supposing that a , b and c are respectively constants and the difference $I_{1,2}$ between the currents I_1 and I_2 is approximated as:

$$\begin{aligned}\Delta I_{1,2} &= I_1 - I_2 \\ &\approx \alpha \cdot I_{c25} \cdot (a - b \cdot V_{L0}^2 - c \cdot V_{L0}^4) \dots\dots\dots(32),\end{aligned}$$

the following equation can be obtained as;

$$\Delta I_{1,2} = \alpha \cdot I_0 \cdot \exp(V_{RF}/V_T) \times (a - b \cdot V_{L0}^2 - c \cdot V_{L0}^4) \dots\dots(33)$$

By expanding the logarithmic function of Eq. (33) in series, the following equation can be obtained as;

$$\begin{aligned}\Delta I_{1,2} &\approx \alpha \cdot [1 + (V_{RF}/V_T) + (1/2)(V_{RF}/V_T)^2 + \dots\dots] \\ &\quad \times (a - b \cdot V_{L0}^2 - c \cdot V_{L0}^4) \\ &\approx \alpha \cdot [1 + (V_{RF}/V_T)] \times (a - b \cdot V_{L0}^2 - c \cdot V_{L0}^4) \\ &= \alpha \cdot b \cdot (I_0/V_T) \cdot V_{RF} \cdot V_{L0}^2 + \alpha \cdot a \cdot I_0 + \alpha \cdot b \cdot I_0 \cdot V_{L0}^2 \\ &\quad + \alpha \cdot c \cdot I_0 \cdot V_{L0}^2 + (\alpha \cdot a \cdot I_0/V_T) \cdot V_{RF} \\ &\quad + (\alpha \cdot c \cdot I_0/V_T) \cdot V_{RF} \cdot V_{L0}^4 \\ &\dots\dots\dots(34)\end{aligned}$$

Referring to Eq. (34), the product of $(V_{L0})^2$ and V_{RF} , or $(V_{L0})^2 \cdot V_{RF}$ is included. Accordingly, the frequency component contained in the differential current $\Delta I_{1,2}$ is $(2f_{L0} + f_{RF})$ and $(2f_{L0} - f_{RF})$ or $(f_{RF} \pm 2f_{L0})$. The current $\Delta I_{1,2}$ is a differential output between the currents I_1 and I_2 , so that the same frequency component as shown



above is contained into each of the currents I1 and I2 as well.

As explained above, according to the circuit shown in Fig. 6, a double frequency signal of the input voltage VLO and a signal of the AC signal VRF are mixed to be outputted, so that it is found that the circuit shown in Fig. 6 constitutes a frequency doubling and mixing circuit.

[Fourth Embodiment]

Fig. 8 shows a fourth embodiment of this invention, in which transistor-pair Q27 and Q28 and transistor-pair Q29 and Q30 are differential transistor-pairs which respectively have emitters connected in common. The emitter size of the transistor-pair Q28 and Q29 is made 1, the emitter size of the corresponding transistor-pair Q27 and Q30 is made K ($K > 1$). Between these two sets of differential transistor-pairs, the collector of the transistors Q27 and Q30, and that of the transistors Q28 and Q29, which are respectively equal in emitter size, are connected in common, the base of the transistor Q27 and that of the transistor Q29, which are different in emitter size, are connected in common in one input terminal 21, and the base of the transistor Q28 and



that of the transistor Q30, which are different in emitter size, are connected in common to the other input terminal 22. The collectors of the transistors Q28 and Q29 are connected directly to a power source VCC, and the collectors of the transistors Q27 and Q30 are connected to an output terminal 25 and at the same time, to the power source VCC through a resistor RL. Transistors Q25 and Q26 respectively constitute known constant current sources similar to the third embodiment.

With the structure as above, the collector currents Ic27, Ic28, Ic29, Ic30, Ic25 and Ic26 of respective transistors Q27, Q28, Q29, Q30, Q25 and Q26 become as follows;

$$I_{c27} = \frac{\alpha \cdot I_{c25}}{1 + (1/K) \cdot \exp(-V_{L0}/V_T)} \quad \dots\dots\dots(35)$$

$$I_{c28} = \frac{\alpha \cdot I_{c25}}{1 + K \cdot \exp(-V_{L0}/V_T)} \quad \dots\dots\dots(36)$$

$$I_{c29} = \frac{\alpha \cdot I_{c26}}{1 + (1/K) \cdot \exp(-V_{L0}/V_T)} \quad \dots\dots\dots(37)$$

$$I_{c30} = \frac{\alpha \cdot I_{c26}}{1 + K \cdot \exp(-V_{L0}/V_T)} \quad \dots\dots\dots(38)$$

$$I_{c25} = I_{c26} = I_s \cdot \exp(-V_{IN}/V_T) \quad \dots\dots\dots(39)$$



In Eqs. (35) to (39), supposing that $\exp(VL0/VT) \gg 1$ and $(VIN/VT) \gg 1$, approximations are made as follows;

$$\exp(VL0/VT) - 1 \approx \exp(VL0/VT) \quad \dots\dots\dots(40)$$

$$\exp(VIN/VT) - 1 \approx \exp(VIN/VT) \quad \dots\dots\dots(41)$$

As a result, the sum I_{p1} of the collector currents I_{c27} and I_{c30} and the sum I_{q1} of the collector currents I_{c28} and I_{c29} can be expressed as follows;

$$\begin{aligned} I_{p1} &= I_{c27} + I_{c30} \\ &= \alpha \cdot I_s \cdot \exp(-VIN/VT) \\ &\quad \times \frac{1}{1+(1/K)\exp(-VL0/VT)} + \frac{1}{1+(1/K)\exp(VL0/VT)} \\ &\quad \dots\dots\dots(42) \end{aligned}$$

$$\begin{aligned} I_{q1} &= I_{c28} + I_{c29} \\ &= \alpha \cdot I_s \cdot \exp(VIN/VT) \\ &\quad \times \frac{1}{1+K \cdot \exp(-VL0/VT)} + \frac{1}{1+K \cdot \exp(VL0/VT)} \\ &\quad \dots\dots\dots(43) \end{aligned}$$

Accordingly, the difference $\Delta I_{p1,q1}$ between the sums I_{p1} and I_{q1} can be obtained as;

$$\begin{aligned} \Delta I_{p1,q1} &= I_{p1} - I_{q1} \\ &= \alpha \cdot I_s \cdot \exp(VIN/VT) \\ &\quad \times \{ \tanh[(VL0+VK)/2VT] - \tanh[(VL0-VK)/2VT] \} \end{aligned}$$



.....(44)

Here, K is constant, and the current difference $\Delta I_{p1,q1}$ is an even function with respect to V_{L0} . In addition, $\exp(\pm x)$ can be shown as;

$$\exp(\pm x) = \sum_{n=0}^{\infty} (\pm 1)^n \frac{x^n}{n!} \quad \text{.....(45)}$$

When $x \ll 1$, $\tanh x$ can be expanded in series as;

$$\tanh x = x - (x^3/3) + \text{.....} \quad \text{.....(46)}$$

the current difference $\Delta I_{p1,q1}$ can be transformed as;

$$\begin{aligned} \Delta I_{p1,q1} &= \frac{2\alpha \cdot I_s \cdot \exp(V_{IN}/V_T) \cdot [k - (1/k)]}{[k + (1/k)] + 2 + (V_{L0}/V_T)^2 + (1/12)(-V_{L0}/V_T)^4 + \text{.....}} \\ &= \alpha \cdot I_s \cdot \exp(V_{IN}/V_T) \\ &\quad \times (\ln K) \cdot \{[1 - (\ln K)^2/12] - (V_{L0})^2/4V_T\} \text{.....} \\ &\quad \text{.....(47)} \end{aligned}$$

As a result, the current difference $\Delta I_{p1,q1}$ can be approximated as follows;

$$\begin{aligned} \Delta I_{p1,q1} &\approx \alpha \cdot I_s \cdot \exp(V_{IN}/V_T) \\ &\quad \times (\ln K) \cdot \{[1 - (\ln K)^2/12] - (V_{L0})^2/4V_T\} \\ &\quad \text{.....(48)} \end{aligned}$$

As seen from Eq. (48), the current difference $\Delta I_{p1,q1}$ can be approximated by an equation having the term of the square of V_{L0} as the input signal, so that almost of the



frequency components contained into the current difference $I_{pl,q1}$, as compared with the input signal frequency of f_{L0} , have a frequency of $2 f_{L0}$ when the DC component is removed therefrom. As a result, in this embodiment, it can be found that such a frequency multiplying operation that is effected in the third embodiment is carried out.

Next, if the input signal V_{IN} is set as below according to Eq. (26) similarly to the third embodiment;

$$V_{IN} = V_F + V_{RF},$$

Eq. (48) becomes below.

$$\begin{aligned} \Delta I_{pl,q1} &\approx \alpha \cdot I_0 \cdot \exp(V_{RF}/V_T) \\ &\times (\ln K) \cdot \{ [1 - (\ln K)^2/12] - (V_{L0})^2/4V_T \} \\ &\dots\dots (49) \end{aligned}$$

where I_0 is equal to that shown in Eq. (27).

And approximating Eq. (49) by expanding the logarithmic function of Eq. (49) in accordance with Eq. (8) with an assumption of $V_{RF} \ll V_T$, the following can be obtained as;

$$\begin{aligned} \Delta I_{pl,q1} &\approx \alpha \cdot I_0 \cdot (\ln K) \cdot \{ [1 - (\ln K)^2/12] - (V_{L0})^2/4V_T \} \\ &\times [1 + (V_{RF}/V_T) + (1/2)(V_{RF}/V_T)^2 + \dots\dots] \\ &= \alpha \cdot I_0 \cdot (\ln K) \\ &\times \{ [1 - (\ln K)^2/12] - (V_{L0})^2/4V_T \\ &\quad + [1 - (\ln K)^2/12] \cdot (V_{RF}/V_T) \} \end{aligned}$$



$$- [1/4(VT)^3] \cdot (VLO \cdot VRF) + \dots] \dots (50)$$

According to Eq. (50), the current difference $\Delta I_{p1,q1}$ contains the product of $(VLO)^2$ and VRF , or $(VLO)^2 \cdot VRF$, so that in the fourth embodiment shown in Fig. 8, a frequency doubling and mixing circuit can be provided that a double frequency signal of the input signal VLO and the input signal VRF are mixed to be outputted.

[Fifth Embodiment]

As shown in Fig. 9, an emitter resistor with a value of RE/K is inserted into an emitter of each of transistors $Q27$ and $Q30$ large in emitter size, and an emitter resistor with a value of RE is inserted into an emitter of each of transistors $Q28$ and $Q29$ small in emitter size, so that the amplitude level of an input signal VLO can be enhanced. As a result, by appropriately selecting the resistor value of an emitter resistor, the input signal VLO can be applied at any amplitude level.

As explained above, in the third to fifth embodiments of this invention, the circuit has two sets of differential transistor-pairs having emitter resistors in one



transistor of each pair or different emitter sizes, and two transistors respectively providing constant current sources to these two sets of differential transistor-pairs, so that the frequency doubling and mixing operations can be simultaneously carried out, which means that such a circuit can be practically realized with a small number of elements to be used, thus resulting in reductions of the circuit scale and power consumption.

[Sixth Embodiment]

Fig. 10 shows a sixth embodiment of this invention, in which 31 and 32 constitute a first input terminal pair to be applied with a local signal (voltage V_{L0}) to be doubled, 33 and 34 constitute a second input terminal pair to be applied with a mixing signal (voltage V_{IN}), and Q41 and Q42, Q43 and Q44 are differential transistor-pairs whose emitters are respectively voltage-connected with each other. The transistors Q42 and Q43 only of respective differential transistor pairs are provided with emitter resistors R_E . These two sets of differential transistor-pairs constitute a first differential amplifier. In these two sets, the collector of the transistors Q42 and Q43 each having an



emitter resistor and that of the transistors Q41 and Q44 each not having an emitter resistor are respectively connected in common, the base of the transistor Q41 not having an emitter resistor and that of the transistor Q43 having an emitter resistor are connected to one input terminal 31 of the first input terminal-pair, and yet the base of the transistor Q42 having an emitter resistor and that of transistor Q44 not having an emitter resistor are connected in common to the other input terminal 32 of the first input terminal-pair. The emitter of each of these two sets of differential transistor-pairs is connected to constant current source I0.

Then, transistors Q51 and Q52 constitute a second differential amplifier. The base of the transistor Q51 is connected to the input terminal 33 of the second input terminal-pair, the base of the transistor Q52 is connected to the input terminal 34 of the second input terminal-pair. The transistors Q45 and Q46, Q47 and Q48, and Q49 and Q50 respectively constitute a first current mirror circuit and a second current mirror circuit and a third current mirror circuit. Said first and second differential amplifiers are directly connected to each other by the first and second



current mirror circuits. V_{CC} is a power source voltage, R_L is a load resistance, and 35 is an output terminal from which the mixing output (voltage V_0) is taken out.

Next, the operational principle of the circuit structured as above will be explained below. In the first differential amplifier, between the base voltages V_{B41} and V_{B42} of respective differential transistor-pair Q_{41} and Q_{42} , the collector current I_{c42} of the transistor Q_{42} and the current amplification factor α , the following is established as:

$$V_{L0} + V_{B41} + (I_{c42}/\alpha) R_E - V_{B42} = 0 \quad \dots\dots(51)$$

Also, $V_{B41} \gg V_T$ and $V_{B42} \gg V_T$, so that if the collector current of the transistor Q_{41} is expressed as I_{c41} and the saturation currents of the differential transistor-pair Q_{41} and Q_{42} are respectively expressed as I_{s41} and I_{s42} , there exists the following relation between V_{B41} , V_{B42} and V_T as:

$$V_{B41} = V_T \cdot \ln(I_{c41}/I_{s41}) \quad \dots\dots(52)$$

$$V_{B42} = V_T \cdot \ln(I_{c42}/I_{s42}) \quad \dots\dots(53)$$

Here, supposing as $I_{s41} = I_{s42}$, Eq. (51) can be shown as follows:

$$V_{L0} + V_T \cdot \ln(I_{c41}/I_{c42}) + (I_{c42}/\alpha) R_E = 0 \quad \dots\dots(54)$$

Also, there exists the following relation between the



constant current source I_0 and the collector currents I_{c41} and I_{c42} as:

$$\alpha \cdot I_0 = I_{c41} + I_{c42} \quad \dots\dots\dots(55)$$

Hence, Eq. (54) can be made as follows:

$$V_{L0} + V_T \cdot \ln[(\alpha \cdot I_0 / I_{c42}) + 1] + (I_{c42} / \alpha) R_E = 0 \quad \dots\dots\dots(56)$$

Here, in order to obtain the gradient of the collector current I_{c42} with respect to the local signal (voltage V_{L0}) as a signal to be doubled, I_{c42} is differentiated with respect to V_{L0} as follows:

$$\frac{d I_{c42}}{d V_{L0}} = \frac{1}{V_T \frac{I_{c42} (\alpha \cdot I_0 - I_{c42})}{\alpha \cdot I_0} + \frac{R_E}{\alpha}} \quad \dots\dots\dots(57)$$

The point where the differentiating value (absolute value) is maximized is obtained when the following is satisfied as:

$$I_{c42} = I_{c41} = (\alpha \cdot I_0 / 2) \quad \dots\dots\dots(58)$$

Thus, the value of such point can be shown by the following equation:

$$\left| \frac{d I_{c42}}{d V_{L0}} \right|_{I_{c42} = (\alpha \cdot I_0 / 2)} = \frac{\alpha \cdot I_0}{4 V_T + R_E \cdot I_0} \quad \dots\dots\dots(59)$$



The input voltage V_{L0} in this case can be obtained as;

$$V_{L0} = -(1/2) R_E \cdot I_{c45} \quad \dots\dots\dots(60)$$

In the differential transistor-pair consisting of the transistors Q_{43} and Q_{44} , the determinations can be made in the same manner as shown above. The derivation can be achieved by inverting the polarity of the input voltage V_{L0} , thus being omitted here. Between the collector currents I_{c41} , I_{c42} , I_{c43} and I_{c44} of respective transistors determined as above and the input voltage V_{L0} , there exist the relations as shown in Fig. 11. Fig. 11 shows the case of being $R_E \cdot I_0 = 16 V_T$.

As seen from Fig. 11, the tangent at the point where the gradient is maximized has a point of being $V_{L0} = 4 V_T$ as the starting point and passes through points of being $V_{L0} = -(1/2) R_E \cdot I_0$ and $I_{c42} = -(1/2) \alpha \cdot I_0$. In this case, the starting point of $4 V_T$ is made constant independently of the value of emitter resistor R_E and that of constant current source I_0 .

Here, I_5 and I_6 are defined as follows;

$$I_5 = I_{c41} + I_{c44} \quad \dots\dots\dots(61)$$

$$I_6 = I_{c42} + I_{c43} \quad \dots\dots\dots(62)$$

Then, it is found that each of the currents I_5 and I_6



is a differential current and has a both wave rectification property for the input voltage V_{L0} . As a result, by optimizing the value of the emitter resistance and that of the constant current source I_0 , each of the currents I_5 and I_6 can be considerably approximated in characteristic to the square characteristic of the input voltage V_{L0} . That is, this first differential amplifier constitutes a frequency doubling circuit, and each of the currents I_5 and I_6 is an even function with respect to the input voltage V_{L0} , so that even if approximation is made up to the fourth term of the input voltage V_{L0} as the primary approximation of each of the currents I_5 and I_6 , it can be expected that an error to be generated is small. Therefore, the current difference $I_{5,6}$ between the currents I_5 and I_6 can be approximated with a , b and c each as a constant as follows:

$$\begin{aligned}\Delta I_{5,6} &= I_5 - I_6 \\ &\approx \alpha \cdot I_0 \cdot (a - b \cdot V_{L0}^2 - c \cdot V_{L0}^4) \quad \dots\dots\dots(63)\end{aligned}$$

This current difference $I_{5,6}$ is generated by the first and second current mirror circuits respectively consisting of the transistors Q_{45} and Q_{46} and the transistors Q_{47} and Q_{48} . And, an electric current substantially equal to this current difference $I_{5,6}$ is flowed to the collector of the



transistor Q48, and this electric current becomes a control current of the third current mirror circuit consisting of the transistors Q49 and Q50. If the collector currents of the transistors Q51 and Q52 are expressed as I_{c51} and I_{c52} , respectively, the difference therebetween, or the output current I_{OUT} of the second differential amplifier can be shown as follows;

$$\begin{aligned} I_{OUT} &= I_{c51} - I_{c52} \\ &= \alpha \cdot I_{5.6} \cdot \tanh(V_{IN}/2V_T) \end{aligned} \quad \dots\dots\dots(64)$$

Here, with $V_{IN} \ll 2V_T$, $\tanh(V_{IN}/2V_T)$ is expanded in series according to Eq. (13), and further Eq. (63) is substituted for $I_{5.6}$. Thus, the output current I_{OUT} can be approximated as follows;

$$\begin{aligned} I_{OUT} &\approx \alpha^2 \cdot I_0 (a - b \cdot V_{LO}^2 - c \cdot V_{LO}^4) \\ &\quad \times \{ (V_{IN}/2V_T) - (1/3)(V_{IN}/2V_T)^3 \} \end{aligned} \quad \dots\dots\dots(65)$$

Eq. (65) can be further expanded as follows;

$$\begin{aligned} I_{OUT} &= [(\alpha^2 \cdot I_0 \cdot a)/(2V_T)] V_{IN} \\ &\quad - [(\alpha^2 \cdot I_0 \cdot a)/(24V_T^3)] V_{IN}^3 \\ &\quad - [(\alpha^2 \cdot I_0 \cdot b)/(2V_T)] V_{IN} \cdot V_{LO}^2 \\ &\quad + [(\alpha^2 \cdot I_0 \cdot b)/(24V_T^3)] V_{LO}^2 \cdot V_{IN}^3 \\ &\quad - [(\alpha^2 \cdot I_0 \cdot c)/(2V_T)] V_{IN} \cdot V_{LO}^4 \\ &\quad + [(\alpha^2 \cdot I_0 \cdot c)/(24V_T^3)] V_{IN}^3 \cdot V_{LO}^4 \end{aligned}$$



.....(66)

Eq. (66) contains the product $(VLO)^2 \cdot VIN$ of $(VLO)^2$ and VIN , so that it is found that the output current $IOUT$ contains frequency components of $(2fLO + fIN)$ and $(2fLO - fIN)$ or $(2fIN - 2fLO)$. In fact, if $a \ll b$, and $c \ll b$, these frequency components become predominant. As a result, from the output terminal 35 is outputted by mixing the double frequency of the input signal VLO with the input signal VIN , which means that this second differential amplifier becomes a mixing circuit.

As explained above, the circuit shown in Fig. 10 makes it possible to effect the frequency doubling operation of the input signal VLO and the mixing operation of the signal thus frequency-multiplied with the input signal VIN by one unit of circuit. In addition, the frequency doubling circuit and the mixing circuit are connected directly with each other through a current mirror circuit, and an output current of the frequency doubling circuit is used as the driving current of the mixing circuit, so that a bias circuit of the mixing circuit can be eliminated, being adapted to be formed into semiconductor integrated circuit. It is needless to say that a filter to be equipped outside



does not need, making possible to widen the frequency band to be applied.

[Seventh Embodiment]

Fig. 12 is for explaining a seventh embodiment of this invention. In this embodiment, a constant current source I00 is inserted in parallel in the transistor Q49 of the third current mirror circuit as the driving source of the second differential amplifier shown in the sixth embodiment thereby to enhance the conversion gain of the mixing circuit.

In the third current mirror circuit consisting of the transistors Q49 and Q50, a current obtained by subtracting a constant DC value I00 of the constant current source (I00) from the output of the current mirror circuit of the preceding stage ($\Delta I_{5,6}$ shown by Eq.(63)) is made as the driving current of the second differential amplifier (mixing circuit), so that if it is expressed as I' , Eq.(63) already shown above can be given as follows in this embodiment:

$$\begin{aligned}\Delta I' &= I_1 - I_2 - I_{00} \\ &\approx \alpha \cdot I_0 \cdot (a' - b \cdot V_{L0}^2 - c \cdot V_{L0}^4) \dots\dots\dots(67)\end{aligned}$$

Provided, in Eq.(67), between the constant current sources



I00 and I0, there exists the following relation as;

$$I00 = (a - a') \alpha \cdot I0 \quad \dots\dots\dots(68)$$

As a result, by substituting a of Eqs.(65) and (66) with a', the output current I00' of the mixing circuit can be given as follows;

$$\begin{aligned} I' \text{ OUT} &= Ic51 - Ic52 \\ &\approx \alpha^2 \cdot I0 \cdot (a' - b \cdot VLO^2 - c \cdot VLO^4) \\ &\quad \times \{ (VIN/2VT) - (1/3) (VIN/2VT)^3 \} \\ &\quad \dots\dots\dots(69) \end{aligned}$$

$$\begin{aligned} I' \text{ OUT} &= [(\alpha^2 \cdot I0 \cdot a') / (2VT)] VIN \\ &\quad - [(\alpha^2 \cdot I0 \cdot a') / (24VT^3)] VIN^3 \\ &\quad - [(\alpha^2 \cdot I0 \cdot b) / (2VT)] VIN \cdot VLO^2 \\ &\quad + [(\alpha^2 \cdot I0 \cdot b) / (24VT^3)] VLO^2 \cdot VIN^3 \\ &\quad - [(\alpha^2 \cdot I0 \cdot c) / (2VT)] VIN \cdot VLO^4 \\ &\quad + [(\alpha^2 \cdot I0 \cdot c) / (24VT^3)] VIN^3 \cdot VLO^4 \\ &\quad \dots\dots\dots(70) \end{aligned}$$

Here, $a > a'$, so that the proportion of the product $(VLO)^2 \cdot VIN$ contained into the output current I00' becomes larger by an amount that the proportions of VIN and $(VLO)^2$ are reduced than the product $(VLO)^2 \cdot VIN$ contained into the output current I00. This means that the conversion gain of the mixing circuit can be enhanced. In addition, it is needless



to say that the circuit subtracting the constant DC value from the current difference $\Delta I_{5,6}$ is not limited to the circuit shown in Fig. 12.

As explained above, in the sixth and seventh embodiments, the first differential amplifier constitutes a frequency doubling circuit, and the second differential amplifier constitutes a mixing circuit. The frequency doubling circuit and mixing circuit thus constituted are connected directly with each other through a current mirror circuit, so that the frequency doubling operation of a signal to be multiplied and the mixing operation of the signal thus frequency-doubled with a mixing signal can be achieved by one unit of circuit.

In addition, in the seventh embodiment, a constant direct current is subtracted from each of output currents of the first differential amplifier as the frequency doubling circuit thereby to drive the second differential amplifier as the mixing circuit, so that the conversion gain of the mixing circuit can be improved.



CLAIMS:



1. A frequency doubling and mixing circuit comprising:
a first set of differential transistor-pair which has emitters connected in common and an emitter size of K:1 (K is larger than 1);

a second set of differential transistor-pair which has emitters connected in common and an emitter size of K:1;

a first constant current source for supplying a constant current to said first set of differential transistor-pair;

a second constant current source from supplying a constant current to said second set of differential transistor-pair;

means for deriving a differential current between a common collector output of the transistors each having a larger emitter size of said first and second sets of differential transistor-pairs and a common collector output of the transistors each having a smaller emitter size of said first and second sets of differential transistor-pairs;
and

a differential amplifying circuit having said differential current as its driving current source.

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wherein a signal to be doubled is supplied across common base contacting points of the transistors which are different in emitter size from each other of said first and second differential transistor-pairs, a mixing signal is supplied from a differential input terminal of said differential amplifying circuit, and a circuit output is taken out from an output terminal of said differential amplifying circuit.

2. A frequency doubling and mixing circuit as claimed in claim 1, wherein a resistor is inserted in series into each emitter of said first and second sets of differential transistor-pairs.

3. A frequency doubling and mixing circuit comprising:
a first input terminal-pair for receiving a signal to be doubled;

a second input terminal for receiving a mixing signal;

two sets of differential transistor-pairs having an emitter resistor in one transistor only of each pair, wherein a collector of the transistors each having an emitter resistor and a collector of the transistors each not having an emitter resistor of these two sets of differential transistor-pairs are connected in common, one





of the bases of the transistors each having an emitter resistor and one of the bases of the transistors each not having an emitter resistor are connected in common to one input terminal of said first input terminal-pair, and the other bases thereof are connected in common to the other input terminal of said first input terminal-pair; and

two transistors whose collectors are connected to corresponding emitters of the transistors each not having an emitter resistor of said two sets of differential transistor-pairs, and the bases of which are connected in common to said second input terminal.

4. A frequency doubling and mixing circuit as claimed in claim 3, wherein two sets of differential transistors-pairs each having an emitter size of $K:1$ (K is larger than 1) are provided instead of said two sets of differential transistor-pairs in which one transistor only of each pair has an emitter resistor.

5. A frequency doubling and mixing circuit as claimed in claim 4, wherein an emitter resistor with a resistance value of RE/K is inserted into an emitter of each of said transistors each having a larger emitter size, and an emitter resistor with a resistance value of RE is inserted

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into an emitter of each of said transistors each having a smaller emitter size.

6. A frequency doubling and mixing circuit comprising:

a first input terminal-pair for receiving a signal to be doubled;

a second input terminal-pair for receiving a mixing signal;

a first differential amplifying circuit consisting of two sets of differential transistor-pairs having an emitter resistor inserted into one of the emitters to be connected in common, wherein a collector of the transistors each having an emitter resistor and a collector of the transistors each not having an emitter resistor of these two sets of differential transistor-pairs are connected in common, one of the bases of the transistors each having an emitter resistor and one of the bases of the transistors each not having an emitter resistor are connected in common to one input terminal of said first input terminal pair, the other bases thereof are connected in common to the other input terminal of said first input terminal pair, and the emitters of the transistors of each pair are connected to a constant current source;





a second differential amplifying circuit consisting of a differential transistor-pair wherein one base thereof is connected to one input terminal of said second input terminal pair and the other base thereof is connected to the other input terminal of said second input terminal pair; and

a current mirror circuit for generating a differential current from each output current of said first differential amplifying circuit and driving said second differential amplifying circuit according to the differential current thus obtained.

7. A frequency doubling and mixing circuit as claimed in claim 6, having a circuit for controlling said current mirror circuit so as to drive said second differential amplifying circuit with an electric current obtained by subtracting a constant direct current value from said differential current.

8. A frequency doubling and mixing circuit as claimed in claim 7, wherein said control circuit includes a constant current source inserted in parallel into one of the transistors forming said current mirror circuit.

