March 17, 1970 D. M. UETRECHT 3,501,709

TRANSISTOR R-C FILTERS

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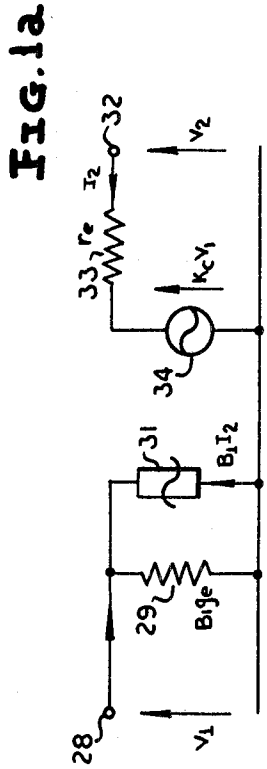


FIG. 1a

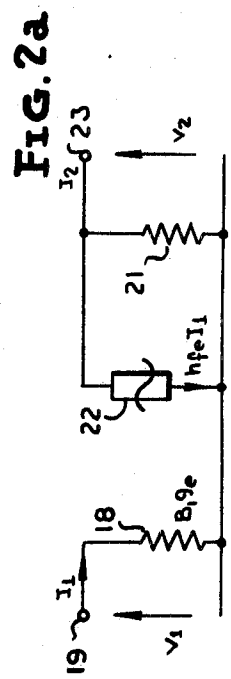


FIG. 2a

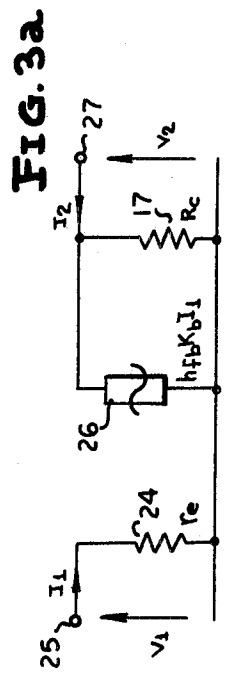


FIG. 3a

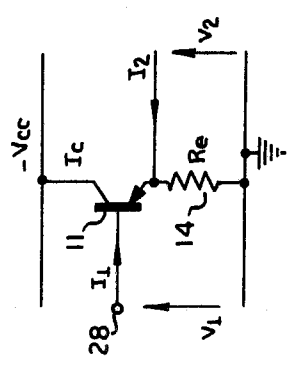


FIG. 1

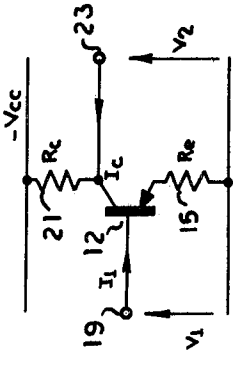


FIG. 2

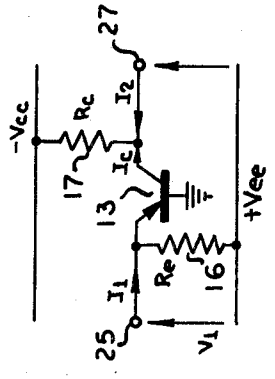


FIG. 3

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FIG. 4

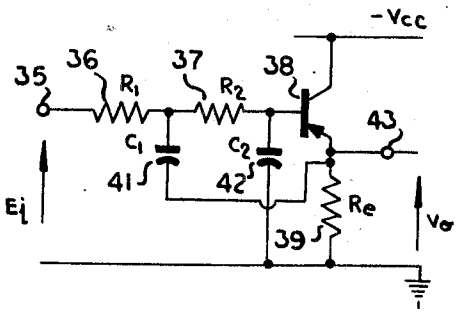


FIG. 5

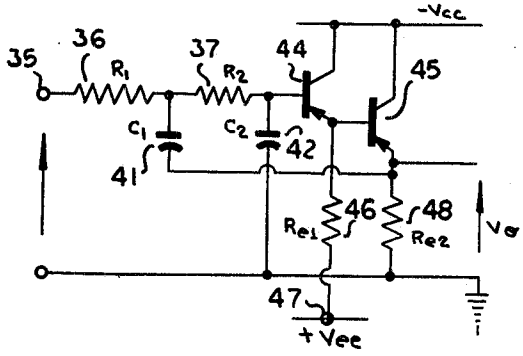


FIG. 6

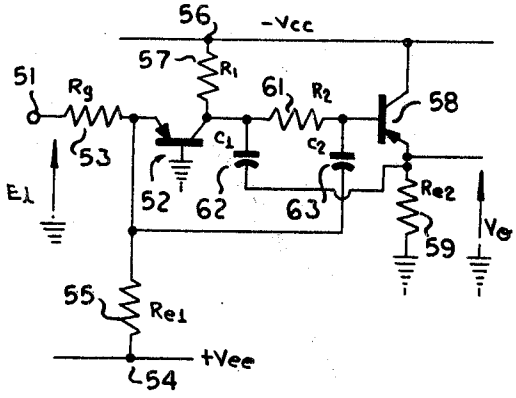
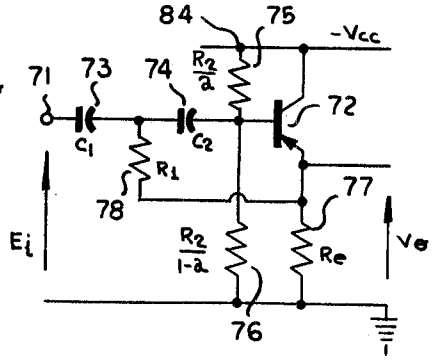


FIG. 8



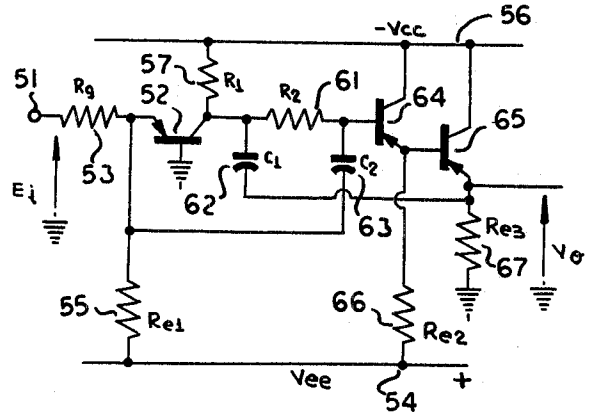


FIG. 7

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FIG. 9

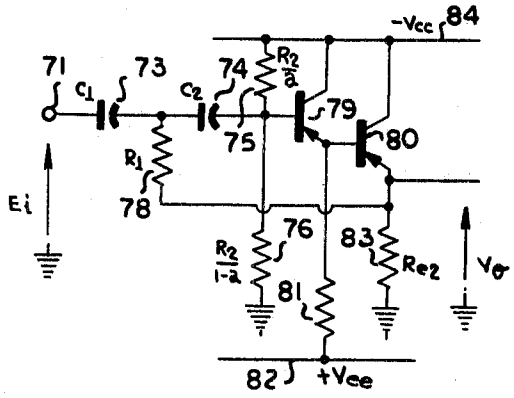


FIG. 10

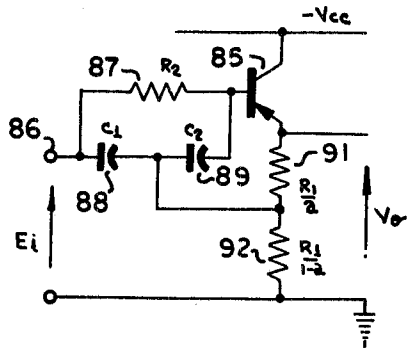


FIG. 11

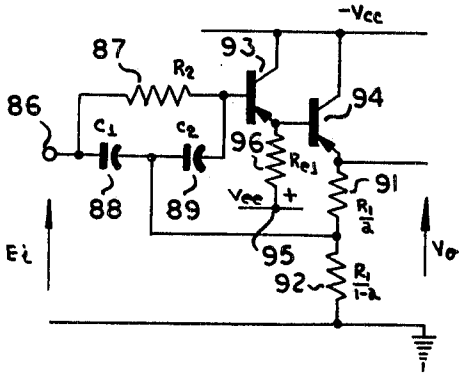


FIG. 12

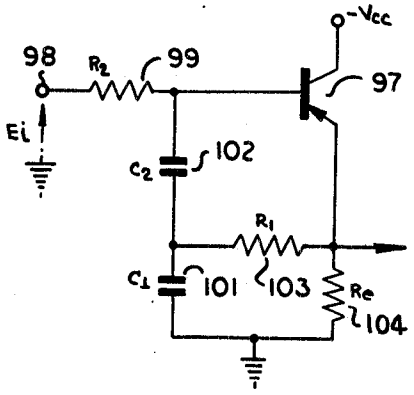


FIG. 13

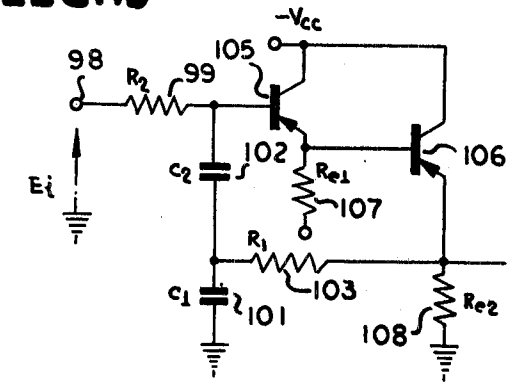
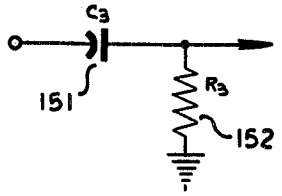


FIG. 13a



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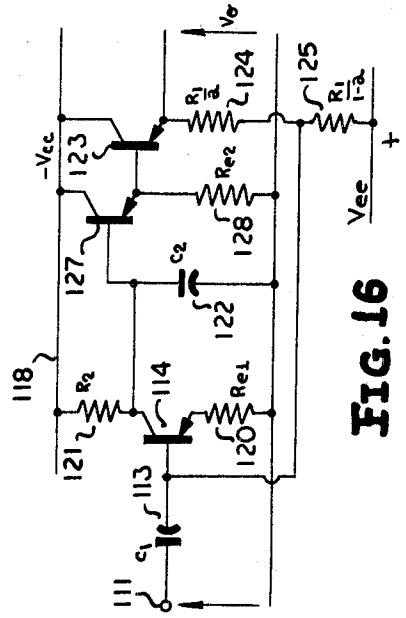


FIG. 16

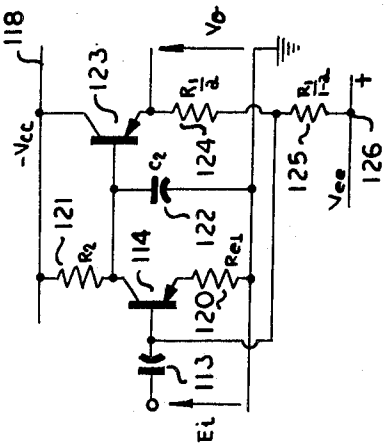


FIG. 15

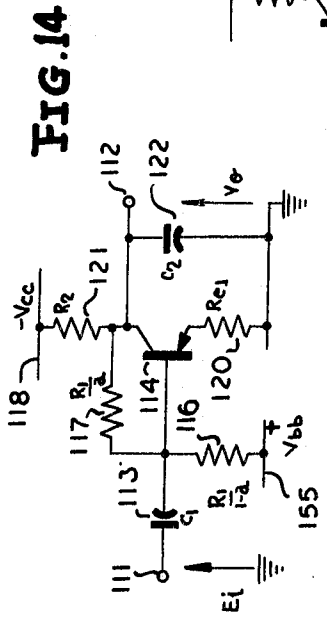


FIG. 14

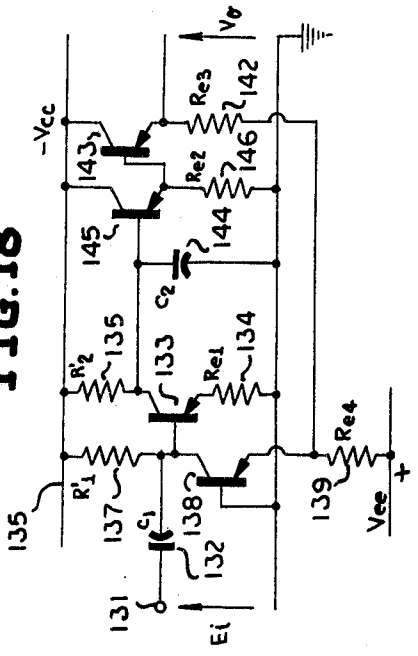


FIG. 18

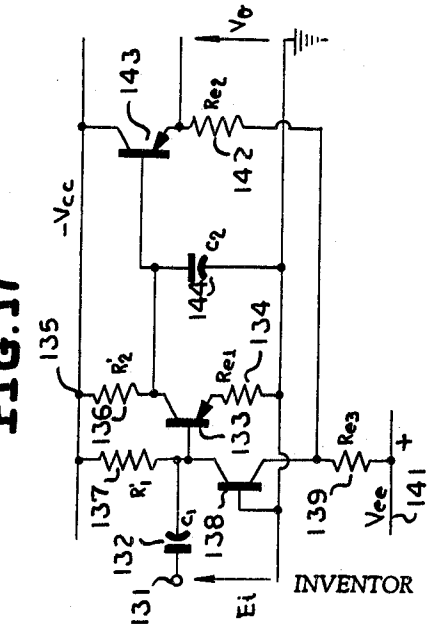


FIG. 17

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United States Patent Office

3,501,709
Patented Mar. 17, 1970

3,501,709

TRANSISTOR R-C FILTERS

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Continuation of application Ser. No. 415,236, Dec. 2, 1964. This application Aug. 14, 1968, Ser. No. 754,107
Int. Cl. H03f 3/04
U.S. Cl. 330—21 1 Claim

ABSTRACT OF THE DISCLOSURE

The present system relates to the transistorization of so-called Sallen and Key RC active filters. Sallen and Key have taught how to design the class of filters with which they were concerned, assuming vacuum tubes as the active elements. When transistors are substituted for tubes it is found that the design criteria provided by Sallen and Key are inadequate, and modified design criteria must be provided.

This application is a continuation of application Ser. No. 415,236, filed Dec. 2, 1964, and now abandoned.

The present invention relates generally to active filter networks and more particularly to transistor filter networks having only capacitive reactances in which the component values are selected to provide steep roll off just beyond the filter band pass.

Filters having steep roll off (at least 6 db per octave) in the vicinity of their cut off frequencies have generally, in the past, required an inductance and capacitance. At audio frequencies, however, the use of inductances is generally to be avoided, for economic and space reasons.

The prior art indicates that filter characteristics substantially the same as those attained with inductance, capacitance networks can be obtained with resistance capacitance networks connected with active elements, such as tubes or transistors. One of the most significant articles on the subject, written by Sallen and Key, appeared on pages 74–85 in the March 1955 issue of the "IRE Transactions on Circuit Theory." This article indicates that attainment of filters with high Q values, (5) a necessity in achieving sharp cut off at frequencies just beyond the filter band pass, requires highly stabilized, complicated active elements, and carefully adjusted passive elements having close tolerances. The same article indicates that if these precautions are not taken Q's of only 2 are realized. Also, the idealized responses calculated by Sallen and Key ignore the effects of finite input and output impedance. In mass production circuits adapted for commercial use, it is not practical to employ complicated negative feedback stabilizing circuits or expensive components having extreme tolerances. For economic reasons, transistors, elements having input impedances that cannot be ignored, are preferably employed.

According to the present invention, high Q (at least 5) transistorized filters having no inductances are provided in which the circuit values are selected on a predetermined generalized, mathematical basis. Each filter circuit includes two capacity branches and two resistive

branches so that the transfer function denominators can be represented by the form:

$$D(s)=d_o+d_1s+d_2s^2 \quad (1)$$

where: s is the Laplace operator, and d_0, d_1 and d_2 are constants. By letting

$$d=\frac{d_1}{\sqrt{d_od_2}} \text{ and } w_o=\sqrt{\frac{d_o}{d_2}}$$

$$D(s)=d_o\left[1+d\frac{s}{w_o}+\left(\frac{s}{w_o}\right)^2\right] \quad (2)$$

where:

$d=1/Q$, and

w_o=cut off angular frequency of a low or high pass filter and the center frequency of a bandpass or elimination filter.

Normalizing Equation 2, D(s) becomes

$$D_n(s)=1+ds+s^2 \quad (3)$$

It can be shown that

$$d=\frac{K_1R_1C_1+HR_1C_2+BR_2C_1+K_2R_2C_2}{\sqrt{R_1R_2C_1C_2}} \quad (4)$$

where:

R_1=resistance of first resistive branch,

R_2=resistance of second resistive branch,

C_1=capacity of first capacitive branch,

C_2=capacity of second capacitive branch,

K_1, K_2, H and B are all constants dependent upon the transistor and circuit configuration.

Equation 4 can be rewritten as:

$$d=K_1\rho\sigma+H\frac{\rho}{\sigma}+B\frac{\sigma}{\rho}+\frac{K_2}{\rho\sigma} \quad (5)$$

where:

$$\rho=\frac{R_1}{R}=\frac{R}{R_2},$$

$$\sigma=\frac{C_1}{C}=\frac{C}{C_2},$$

$$R=\sqrt{R_1R_2}, \text{ and}$$

$$C=\sqrt{C_1C_2}.$$

Thus, R and C are resistance and capacity values necessary to satisfy the relationships for σ and ρ. It can be shown that the minimum value of d is given by:

$$d_{min}=2(\sqrt{K_1K_2}+\sqrt{BH}) \quad (6)$$

if

$$\rho^4=\frac{K_2B}{K_1H} \text{ and } \sigma^4=\frac{K_2H}{K_1B}$$

I have found that these mathematical concepts enable transistorized low pass, high pass and band pass active filters having minimum d values of 0.2 to be attained with resistors and capacitors having 10% tolerances as well as transistors having common emitter current gains (h_{fe}) greater than 15. In many configurations, only a single common collector transistor having a positive feedback network need be employed. Reduction of d_{min} is attained by modifying the single transistor circuit so the emitter follower output is replaced with a modified Dar-

lington circuit in which positive feedback is attained from a second emitter follower stage cascaded with the first emitter follower.

d_{min} can also be reduced, according to the present invention, by connecting a common base transistor in the base circuit of a positive voltage feedback circuit. The base of the emitter follower is connected in a positive feedback network to the emitter of the common base transistor. When this configuration is employed as a low pass filter, each of the feedback impedances is a capacitor. Further reduction of d_{min} in this embodiment is also attained by employing the modied Darlington network.

In the positive feedback embodiments, no stability problems arise because the feedback magnitude cannot exceed unity. Unity gain cannot be exceeded since the output voltage of an emitter follower can never be greater than its base input voltage and the output current of a common base transistor cannot exceed its emitter input current.

The generalized teachings of the present invention can also be extended to transistorized, active bandpass filters employing negative, D.C. stabilizing feedback networks. The negative feedback networks, however, generally require transistors with appreciably larger values of h_{fe} than is required with the positive feedback arrangements.

It is, accordingly, an object of the present invention to provide new and improved active filter networks employing only capacitive reactances.

It is another object of the invention to provide new and improved, transistorized active filters, requiring no inductances and having resistances and capacitances selected in a predetermined manner to provide maximum Q.

An additional object of the invention is to provide new and improved high Q transistorized active filters requiring no inductances and utilizing components having standard tolerances.

A further object of the invention is to provided transistorized active filters requiring no inductances or complicated stabilizing networks and having Q's on the order of at least 5.

It is another object of the invention to provide transistorized active filters employing only capactive reactances in which maximum possible values of Q are attained by utilizing predetermined component values.

A further object of the invention is to provide a transistorized active filter requiring no inductances, which filter employs a pair of stabilized positive feedback networks.

Still another object of the invention is to provide an active filter that is inexpensive since it uses a minimum number of conponents and transistors having values of h_{fe} in excess of 15.

The above and still further objects, features and advantages of the present invention will become apparent upon consideration of the following detailed description of one specific embodiment thereof, especially when taken in conjunction with the accompanying drawings, wherein:

FIGURES 1, 2 and 3 are circuit diagrams for the common collector, common emitter and common base circuits, respectively;

FIGURES 1*a*, 2*a* and 3*a* are the equivalent circuits of FIGURES 1, 2 and 3, respectively;

FIGURES 4–7 are circuit diagrams illustrating different embodiments of low pass filters according to the present invention;

FIGURES 8 and 9 are circuit diagrams for different embodiments of high pass filters according to the present invention;

FIGURES 10 and 11 are circuit diagrams of band elimination networks of the present invention;

FIGURES 12 and 13 are circuit diagrams of positive feedback, low pass filters according to the invention;

FIGURE 13*a* is a circuit diagram of a simple high pass filter that is optimumly connected with the circuits of

FIGURES 12 and 13 to convert them into band pass filters; and

FIGURES 14–18 are circuit diagrams of negative feedback, bandpass filters according to the invention.

Many components in the several figures are provided with reference numeral and subscript notations. The same subscript, e.g. R_1, is applied to apparently different components in the different figures. However, the common subscripts are utilized to designate the same elements in Equations 4–6.

In FIGURES 1–3, there are illustrated circuit diagrams for PNP transistors **11–13**, respectively connected in the conventional common collector, common emitter and common base configurations. In each figure, the input current and voltage are designated I_1 and V_1 while the output current and voltage are I_2 and V_2. In FIGURES 1 and 2, the emitter of transistors **11** and **12** are connected through negative feedback resistances, R_e, **14** and **15** to ground. In FIGURE 2, the transistor collector is connected to a negative biasing potential via load resistor, R_c, **21** while the collector of transistor **11**, FIGURE 1, is directly connected to the bias source. In both FIGURES 1 and 2, an audio signal source, such as derived from an electronic organ, is applied between the base of each transistor and ground.

In the common base configuration, FIGURE 3, the audio signal is applied to the emitter base junction of transistor **13** that is shunted by biasing resistor R_e, **16**. The transistor collector is powered from a negative D.C. supply through load resistor R_c, **17**.

Conventional hybrid equivalent circuits, FIGURES 2*a* and 3*a*, are employed for the common emitter and common base configurations. In each circuit, a number of approximations is made because audio frequency signals are employed as sources and relatively small output impedances are fed by the transistors. Thus, effects of reverse voltage ratio can be ignored, and it can be assumed that the transistor output impedance introduces zero phase shift.

The common emitter equivalent circuit, FIGURE 2*a*, comprises an admittance, B_1G_e, **18** connected between input terminal **19** and ground and a current generator $h_{fe}I_1$, **22** shunted by load resistor **21** across output terminal **23** and ground. In the common base equivalent circuit of FIGURE 3*a*, resistance **24**, r_e, is connected between input terminal **25** and ground while the output circuit between terminal **27** and ground comprises current generator **26**, $h_{fb}K_bI_1$, shunted by load resistor **17**.

The common collector equivalent circuit, FIGURE 1*a*, is a modified hybrid configuration in which input terminal **28** is connected to ground by the parallel combination of admittance **29**, B_1G_e, and current generator **31**, B_1I_2, that feeds current into node **28**. The output circuit between terminal **32** and ground comprises the series combination of resistor **33**, r_e, and voltage source **34**, K_cV_1, having a polarity such that the voltage at its ungrounded end corresponds with the polarity of the input voltage at terminal **28**.

In each of FIGURES 1*a*, 2*a* and 3*a*, as well as in the remainder of the specification, like reference letters denote similar circuit parameters. In the grounded base configuration, it has been found that the emitter base resistance, r_e, can be expressed almost independently of the transistor selected in accordance with:

$$r_e \doteq \frac{30+3I_e}{I_e} \doteq \frac{32R_e}{V_e} \text{ or } \frac{r_e}{R_e} \doteq \frac{32}{V_e} \qquad (7)$$

where: I_e and V_e are the emitter bias or average current and voltage, respectively, with I_e, in milliamperes, V_e in millivolts and R_e as well as r_e in ohms. In each configuration, the parameters are defined as:

h_{fe}=collector-base current gain in the common emitter circuit,

h_{fb}=collector-emitter current gain in the common base circuit,

$$B_1=\frac{1}{(1+h_{fe})} \quad (8)$$

$$G_e=\frac{1}{r_e+R_e} \quad (9)$$

$$h_{fb}=\frac{h_{fe}}{1+h_{fe}}=1-B_1 \quad (10)$$

$$K_c=K_b=1-\frac{r_e}{R_e} \quad (11)$$

In certain circuit configurations to be described the subscript 1 for B_1 is changed to a 2, i.e. B_2, and subscripts are added to r_e, K_c, etc., such as r_{e1}, r_{e2}, K_{c1}, K_{c2}. In these instances, more than one transistor is employed in the circuit under consideration, and the several transistors are cascaded. The subscript 1 always has reference to parameters of the transistor directly responsive to the input, the subscript 2 to parameters of the second cascaded transistor, i.e. the one directly responsive to the output of the first transistor. The nomenclature is extended so the parameters of the n^{th} cascaded transistor bear the subscript n.

Reference is now made to the low pass filter of FIGURE 4 wherein an audio source is connected between input terminal **35** and ground. Terminal **35** is connected via series connected resistors **36** and **37**, R_1 and R_2, to the base of emitter follower or common collector PNP transistor **38**. A positive feedback path for the voltage developed across emitter resistor **39**, R_e, is provided with capacitor **41**, C_1, connected between the transistor emitter and the junction between resistors **36** and **37**. Connected between the transistor base and ground is capacitor **42**, C_2. While the FIGURE 4 circuit contains a positive feedback loop, no possibility of instability arises since feedback cannot exceed +1. This is because the emitter feedback voltage can never exceed the base input voltage and it is true of each positive feedback, emitter follower configuration disclosed.

It can be shown that the transfer function between input and output terminals **35** and **43** of the circuit illustrated in FIGURE 4 is represented by:

$$G_n(s)=\frac{h}{1+ds+s^2} \quad (12)$$

where d is defined by Equation 4 and $h=1$. The values of the K_1, K_2, B and H constants in Equation 4 for the FIGURE 4 configuration are:

$$K_1=\frac{r_e}{R_e}+\frac{r_e}{R_1},$$

$$K_2=1$$

$$B=B_1\left(1+\frac{R_1}{R_e}\right), \text{ and}$$

$$H=1$$

The values of r_e and B_1 for transistor **38** are identical with those for the common collector equivalent circuit shown in FIGURE 1*a*. Of course, R_1 and R_2 are the values of resistors **36** and **37**, respectively, while C_1 and C_2 are the values of capacitors **41** and **42**, respectively. The cut off frequency of the filter is given by

$$W_o=\frac{1}{RC}=\frac{1}{\sqrt{R_1R_2C_1C_2}}$$

To provide maximum circuit Q, the value of d_{min} is selected in accordance with Equation 6. As a practical matter, for the circuits disclosed by FIGURES 5, 6, 8, 9, 11 and 13, where more than one transistor stage is employed, it has been found that d_{min} must not always be utilized to attain acceptable Q's, i.e. Q's at least equal to 5. Q's on the order of 5 are derived if transistors having h_{fe} values of at least 15 are utilized. However, for the single transistor embodiments illustrated in FIGURES 4, 7, 10 and 12, strict adherence to Equation 6 is necessary.

The circuit of FIGURE 4 can be modified, as shown in FIGURE 5, to provide lower values of d by replacing the single emitter follower stage **38** with a cascaded pair of emitter follower transistors **44** and **45**. To minimize the value of r_{e1}/R_{e1} so the required Q values are attained, resistor **46**, R_{e1}, in the emitter circuit of transistor **44** is connected to positive D.C. source V_{ee} at terminal **47**. It is seen that the value of r_{e1}/R_{e1} is minimized with this configuration when it is considered from Equation 7 that r_e is inversely related to V_{ee}. Positive feedback, essentially as shown in FIGURE 4, is derived across load resistor **48**, R_{e2}, that is connected between the emitter of transistor **45** and ground.

The circuit of FIGURE 5 has virtually the same transfer function as the FIGURE 4 configuration. The only distinctions are in the values of K_1, and B which have magnitudes indicated by:

$$K_1=\frac{r_{e1}}{R_{e1}}+\frac{r_{e2}}{R_{e2}}+\frac{r_{e2}}{R_1}, \text{ and}$$

$$B=B_1\frac{R_1}{R_{e1}}+B_1B_2\left(1+\frac{R_1}{R_{e2}}\right)$$

where: r_{e1} and r_{e2} are respectively the values of r_e for transistors **44** and **45** in accordance with Equation 7, and

B_1 and B_2 are respectively values of the transistor parameters for stages **44** and **45**, as indicated by B_1 in Equation 8.

Reference is now made to the circuit configuration of FIGURE 6 wherein an audio source is applied from terminal **51** to the emitter of grounded base PNP transistor **52** via resistor, R_g, **53**. Bias potential for the emitter of transistor **52** is supplied by the positive D.C. voltage at terminal **54** through resistor **55**, R_{e1}. Collector bias is provided by the negative D.C. potential at terminal **56** via load resistor **57**, R_1. Resistor **57** is connected with the base emitter junction of emitter follower transistor **58** in virtually the same manner as resistor **36** is connected to transistor **38**, FIGURE 4. In particular, the emitter voltage of transistor **58**, developed across resistor **59**, R_{e2}, is coupled back to the base input resistor **61**, R_2, via capacitor **62**, C_1. Shunt capacitor **63**, C_2, is returned from the base of transistor **58** to the emitter of transistor **52** to provide a positive feedback path that must have a value less than one since the collector current of a grounded base stage is always less than the emitter current.

The low pass transfer function of FIGURE 6 is of exactly the same form as given in Equation 12. The values of the constants differ widely from the low pass circuits of FIGURE 4 and 5 since:

$$K_1=\frac{r_{e2}}{R_{e2}}+\frac{r_{e2}}{R_1}$$

$$K_2=1$$

$$B=B_2\left(1+\frac{R_1}{R_{e2}}\right)$$

$$H=B_1+\frac{r_{e1}}{R_{e1}}$$

$$h=\frac{R_1}{R_g}$$

$$W_o=\frac{1}{RC}=\frac{1}{\sqrt{R_1R_2C_1C_2}}$$

Again, suitable Q values are attained if transistors **52** and **58** have values of h_{fe} in excess of 15.

Reference is now made to FIGURE 7 wherein the circuit of FIGURE 6 is modified so that emitter follower **58** is replaced with cascaded emitter followers **64**, **65**. Emitter resistance **66**, R_{e2}, is returned to the positive voltage at bus **54** to minimize r_e/R_e. Feedback voltage is derived across resistor **67**, R_{e3}, between the emitter of transistor **65** and ground. The parameter values for FIGURES 6 and 7 are the same except those for K_1 and B, in which for FIGURE 7;

$$K_1=\frac{r_{e2}}{R_{e2}}+\frac{r_{e3}}{R_{e3}}+\frac{r_{e3}}{R_1}, \text{ and}$$

$$B=B_2B_3\left(1+\frac{R_1}{R_{e3}}\right)+B_2\frac{R_1}{R_{e2}}$$

The circuit of FIGURE 7 is generally preferred over the one of FIGURE 6 since the response is less sensitive to transistor variations and common base transistor **52**, in combination with common emitter transistor **65**, provides greater isolation between the summed feedback and input signals.

Reference is now made to FIGURE 8 of the drawings, where there is illustrated a high pass filter having an output-input transfer function given by:

$$G_n(s)=\frac{hs^2}{1+ds+s_2} \tag{13}$$

Audio signal at terminal **71** is fed to the base of common collector PNP transistor **72** via series connected capacitors **73** and **74**, C_1 and C_2. Base bias for transistor **72** is established by a voltage divider including resistors **75** and **76**,

$$\frac{R_2}{a} \text{ and } \frac{R_2}{1-a}$$

which divider is connected between the negative D.C. potential at bus **84** and ground. It is noted that the parallel equivalent resistance of resistances **75** and **76** equals R_2, a quantity appearing in basic design Equations 4–6. Positive feedback less than one for the emitter voltage developed across resistor **77**, R_e, is provided by resistance **78**, R_1, to the junction of capacitors **73** and **74**.

In FIGURE 8, the values of the circuit parameters to define K_1, K_2, B and H in Equation 5 are given by:

$$K_1=1$$

$$K_2=\frac{r_e}{R_e}+\frac{r_e}{R_2}$$

$$B=B_1$$

$$H=1+\frac{B_1R_2}{R_e}$$

The constant term in the numerator of Equation 9, h, equals 1 and cut off frequency, W_o, equals $1/RC$. Suitable Q values are easily obtained by correct selection of the circuit components in accordance with Equation 5 if transistor **72** has an h_{fe} of at least 15.

FIGURE 9 is a modification of FIGURE 8 so that the single emitter follower **72** is replaced with cascaded emitter followers **79** and **80**. Emitter resistance **81**, R_{e1}, for transistor **79** is returned to the D.C. positive bias potential at terminal **82** while feedback is derived from the emitter of transistor **80** across grounded resistor **83**, R_{e2}. In FIGURE 9, all parameter values are identical with those in FIGURE 8 except K_2, B and H which are:

$$K_2=\frac{r_{e1}}{R_{e1}}+\frac{r_{e2}}{R_{e2}}+\frac{r_{e2}}{R_2}$$

$$B=B_1B_2+\frac{r_{e2}}{R_2}$$

and

$$H=1+\frac{B_1R_2}{R_{e1}}$$

There are provided in FIGURES 10 and 11, band elimination filters having an output-input transfer function given by:

$$G_n(s)=\frac{1+bs+s^2}{1+ds+s^2} \tag{14}$$

where: d is given in Equation 5,

$$b=K_{1n}\rho\sigma+H_n\frac{\rho}{\sigma}+B_n\frac{\sigma}{\rho}+\frac{K_{2n}}{\rho\sigma} \tag{15}$$

and

σ and ρ are given supra.

The circuit of FIGURE 10 comprises grounded collector PNP transistor **85** having its base connected to an audio signal source at terminal **86** via the combination of resistor **87**, R_2, in parallel with series connected capacitors **88** and **89**, C_1 and C_2. A voltage divider including resistances **91** and **92**,

$$\frac{R_1}{a} \text{ and } \frac{R_1}{1-a}$$

connects the transistor emitter to ground. The tap between resistors **91** and **92** is connected to the junction of capacitors **88** and **89** to establish a positive feedback network.

The center frequency of the rejection band is given by $W_o=1/RC$ while the slope of the filter skirts is inversely related to b. The center frequency attenuation, G (W_o), is determined by the ratio b/d. Therefore, it is necessary that d be large enough to obtain the desired b/d.

The values of the constants in the FIGURE 10 circuit required to satisfy Equations 5 and 11 are given by:

$K_1=1$; $K_{1n}=1$;
$K_2=1-a$; $K_{2n}=0$;
$B=B_1a(2-a)$; $B_n=aB_1$; and
$H-1$; $H_n=1$.

As in the other embodiments, d_{min} in the circuit of FIGURE 10 can be decreased by substituting for the single emitter follower, a pair of cascaded emitter followers **93**, **94**, as shown in FIGURE 11. The emitter of transistor **93** is connected to the positive voltage at terminal **95** via resistance **96** while feedback is from the top of the voltage divider in the emitter circuit of transistor **94**. The values used for determining the circuit constants required in FIGURE 11 are:

$$K_1=1+\frac{B_1R_2}{R_{e1}};\ K_2=1-a+\frac{ar_{e1}}{R_{e1}}+\frac{a^2(1-a)r_{e2}}{R_1}$$

$$B=aB_1B_2;\ H=1+\frac{B_1R_2}{R_{e1}};\ K_{1n}=1$$

$$K_{2n}=0;\ B_n=aB_1B_2;\text{ and } H_n=1$$

Reference is now made to the low pass filters of FIGURES 12 and 13 having transfer functions in accordance with:

$$G_n(s)=\frac{hs+f}{1+ds+s^2} \tag{16}$$

where: f is a constant equal to one and each of the other quantities is defined supra. In FIGURE 12, audio input signal is applied to the base of PNP transistor **97** from terminal **98** via resistance **99**, R_2. The transistor base circuit is shunted by series capacitors **101** and **102**, C_1 and C_2, the tap between which is connected in a positive feedback loop to the transistor emitter via resistor **103**, R_1. Output and feedback voltage at the emitter of transistor **97** is derived across resistor **104**, R_e.

FIGURE 13 is the cascaded emitter follower modification of FIGURE 12 in which emitter biasing voltage for transistor **105** and base biasing voltage for transistor **106** is derived from the positive D.C. voltage at terminal **107** via resistor **107**, R_{e_1}, and feedback voltage is derived across grounded output resistor **108**.

The values of K_1, K_2, B and H for the circuits of FIGURES 12 and 13 are identical with those in the circuits of FIGURES 8 and 9, respectively. For the circuits of FIGURES 12 and 13,

$$h=\rho\left(\frac{1+\sigma^2}{\sigma}\right)$$

and the cut-off frequency of the passband for these circuits is given by $w_o=1/RC$.

In the circuits of FIGURES 4–13, desired Q values are attained with a minimum number of transistors because of the positive feedback networks employed. In general, it can be stated that r_e/R_e should be as small as possible and h_{fe} as large as possible to obtain suitable Q's. Stability is insured, despite positive feedback, because closed loop gain cannot exceed unity.

The high and low pass filters of FIGURES 4–9, 12 and 13 can be cascaded together to provide band pass filters or band pass filters can be formed by connecting simple, passive high or low pass RC networks in cascade with the outputs of these active filters. I have found, however, that with the embodiments of FIGURES 12 and 13, a band pass filter having the desired d values can be attained by connecting capacitor **151**, C_3, of the simple RC filter network shown in FIGURE 13*a* to the emitter of transistor **97** or **106**. Shunt resistor **152**, R_3, of the simple filter is connected from capacitor **151** to ground. By selecting $C_3=(C_1+C_2)$ and $R_3=R_1$, the constant f term in the numerator of Equation 16 is eliminated so the band pass transfer function of

$$G_n(s)=\frac{hs}{1+ds+s_2} \qquad (17)$$

is attained. In Equation 17 each of the parameters is defined to conform with those set forth for the embodiments of FIGURES 12 and 13.

In negative feedback bandpass circuits having response transfer functions given by:

$$G_n(s)=\frac{hs}{1+ds+s^2} \qquad (18)$$

Such as disclosed in FIGURES 14–17, generally more than one stage is required to attain suitable Q's. The circuit of FIGURE 14 is an exception if a very low impedance audio source is connected to terminal **111** and a high impedance is connected to terminal **112**. In FIGURE 14, the signal source at terminal **111** is connected through low frequency attenuating capacitor **113**, C_1, to the base of common emitter PNP transistor **114**. Base bias is established with the positive D.C. voltage at terminal **155** through resistors **116** and **117**,

$$\frac{R_1}{1-a} \text{ and } \frac{R_1}{a}$$

the latter also providing negative feedback stabilizing effects. The emitter and collector of transistor **114** are connected to ground and the negative D.C. bias voltage at terminal **118** by negative feedback resistor **120**, R_{e_1}, and load resistor **121**, R_2, respectively. High frequency attenuation is attained by shunting the path between the collector of transistor **114** and ground with capacitor **122**.

To attain desired values of d in the circuit of FIGURE 14 in accordance with Equations 4–6, the K_1, K_2, B and H parameters are given by:

$$\frac{K_1}{\rho}=\frac{1}{K};$$

$$\frac{K_2}{\rho}=\frac{1+B_1a(1-a)}{K};$$

$$\frac{B}{\rho}=\frac{a}{K}$$

$$H=0;$$

$$h=\frac{-a(1-a)\sigma}{\rho K};$$

$$w_o=\frac{K}{RC};$$

where, by definition:

$$K^2=[\rho^2+a(1-a)]\left(1+\frac{B_1R_1}{R_{e1}}\right)+\frac{aR_1}{R_{e1}}$$

To enable a circuit similar to that illustrated in FIGURE 14 to operate satisfactorily with higher impedance sources and lower output impedances, the modification of FIGURE 15 is provided. In FIGURE 15, the negative feedback network comprising resistors **116** and **117**, FIGURE 14, is replaced with emitter follower stage **123** having resistors **124** and **125**,

$$\frac{R_1}{a} \text{ and } \frac{R_1}{1-a}$$

connecting its emitter to the positive biasing potential at terminal **126**. The tap between resistors **124** and **125** is connected to the base of transistor **114** to form negative feedback stabilizing network. In FIGURE 15, the circuit parameters are given by:

$$\frac{K_1}{\rho}=\frac{1+\frac{B_2a(1-a)}{\rho^2}}{K};$$

$$\frac{K_2}{\rho}=\frac{1+B_1a(1-a)}{K};$$

$$\frac{B}{\rho}=\frac{B_2a}{K};$$

$$H=0;$$

$$h=\frac{-a(1-a)\sigma}{\rho K};$$

$$w_o=\frac{K}{RC}$$

where:

$$K^2=[\rho^2+B_2a(1-a)]\left(1+\frac{B_1R_1}{R_{e1}}\right)+\frac{aR_1}{R_{e1}}$$

It is noted that the parameters for the circuits of FIGURES 14 and 15 are generally similar, the main difference being that a in the former is replaced with B_2a in the latter.

FIGURE 16 is a modification of FIGURE 15 in that emitter follower stage **127**, having emitter load resistance **128** connected to ground, is connected between the output of transistor **114** and the input of transistor **123**. The added stage, in addition to decreasing the value of d_{min}, prevents substantially loading of transistor **114**.

The parameters of FIGURE 16 are given by:

$$\frac{K_1}{\rho}=\frac{1+\frac{B_2R_2}{R_{e2}}}{K};$$

$$\frac{K_2}{\rho}=\frac{1+B_1a(1-a)}{K};$$

$$\frac{B}{\rho}=\frac{B_2B_3a}{K}$$

$$H=0;$$

$$h=\frac{-a(1-a)\sigma}{\rho K};$$

$$w_o=\frac{K}{RC};$$

where:

$$K^2=[\rho^2+B_2B_3a(1-a)]\left(1+\frac{B_1R_1}{R_{e1}}\right)+\frac{aR_1}{R_{e1}}$$

Another manner for attaining bandpass characteristics indicated by Equation 13 with negative feedback is shown in FIGURE 17. Audio signal at terminal **131** is fed through low frequency attenuating capacitor **132**, C_1, to the base of common emitter transistor **133**, the emitter of which is connected through resistor **134**, R_{e1}, to ground. Collector bias for transistor **133** is supplied from the negative, D.C. voltage on bus **135** through load resistor **136**, R_2'. Base bias is provided with the voltage divider comprising resistor **137**, R_1', the collector emitter path of grounded base transistor **138**, and resistor **139** that is connected between the negative and positive supplies at buses **135** and **141**. Transistor **138** serves as a signal responsive variable resistor for stabilizing the operating point of transistor **133** and is controlled in response to the negative feedback voltage applied through resistor **142**, R_{e2}, to its emitter by the voltage deriving from the emitter of transistor **143**. To provide high frequency attenuation, capacitor **144**, C_2, connects the collector of transistor **133** to ground.

The component parameters to give the required values of d in the FIGURE 17 circuit are:

$$K_1=K_2=\frac{1}{K};$$

$$B=H=0;$$

$$h=\frac{R_1\sigma}{R_{e1}\rho K};\text{ and}$$

$$w_o=\frac{K}{RC},$$

where:

$$K=1+\frac{R_1R_2}{R_{e1}R_{e2}}\quad R_{1'}=\frac{R_1}{1-\frac{B_1R_1}{R_{e1}}}\text{ and}$$

$$R_{2'}=\frac{R_2}{1-\frac{B_2H_2}{R_{e2}}}$$

The circuit of FIGURE 17 is modified according to FIGURE 18 by connecting emitter follower **145**, having resistor **146**, R_{e2}, between the output of transistor **133** and input of transistor **143**. The K_1, K_2, etc. parameters

of the FIGURE 18 circuit are identical with those of FIGURE 17 except that each occurrence of R_{e2} in the former is replaced with R_{e3} in the latter and

$$R_{2'}=\frac{R_2}{1-\frac{B_2R_2}{R_{e2}}-\frac{B_2B_3R_2}{R_{e3}}}$$

I claim:

1. An active filter for audio signal provided by an audio source, said filter having only capacitive reactances, comprising transistor means having an input circuit responsive to said signal, an output circuit, and a feedback circuit between said input and output circuits, said circuits including first and second resistive branches having values of R_1 and R_2, respectively, and first and second capacitive branches having values of C_1 and C_2, respectively, a source of electrode biasing potential, biasing resistance means connecting at least one electrode of said transistor with said biasing source, means interconnecting said branches with said transistor means such as to provide a network having a transfer function with a normalized denominator represented by:

$$1+ds+s^2$$

where:

$$d=\frac{1}{Q}=\frac{K_1R_1C_1+HR_1C_2+BR_2C_1+K_2R_2C_2}{\sqrt{R_1R_2C_1C_2}}$$

K_1, K_2, H and B are predetermined circuit parameter functions of transistor beta and emitter impedance, the biasing resistance means and internal transistor characteristics, the values of the impedances of said branches, the biasing resistance means and the internal characteristics of said transistor means being arranged such that d equals or is less than 0.2, wherein said transistor means comprises a common collector configuration, said first and second resistive branches being connected between said audio source and the base input of said configuration, said first capacitor branch being connected between the emitter output of said configuration and the junction between said resistive branches, said second capacitor branch shunting said base input, and wherein said transistor means further comprises a grounded base transistor stage connecting said audio source to said base input, the emitter of said grounded base stage being connected to be responsive to said audio source and with one end of the second capacitor branch to form a positive feedback loop having gain less than one.

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