March 17, 1970
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TRANSISTOR R-C FILTERS

Original Filed Dec. 2, 1964

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Fig. 9

Fig. 10

Fig. 11

Fig. 12

Fig. 13

Fig. 13a
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, 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:

\[ D(s) = d_0 + d_1 s + d_2 s^2 \]  

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

\[ d = \frac{d_1}{\sqrt{d_0 d_2}} \text{ and } w_c = \sqrt{\frac{d_2}{d_0}} \]

\[ D(s) = d_0 \left[ 1 + d_1 \frac{s}{w_c} + \left( \frac{s}{w_c} \right)^2 \right] \]

where: \( d = 1/Q \), and \( w_c \) = 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

\[ D_0(s) = 1 + d_1 s \frac{s}{w_c} \]

It can be shown that

\[ d = \frac{K_1 R_1 C_1 + H R_2 C_1 + B R_1 C_2 + K_2 R_2 C_2}{\sqrt{R_1 R_2 C_1 C_2}} \]

where:

\[ R_1 = \text{resistance of first resistive branch,} \]

\[ R_2 = \text{resistance of second resistive branch,} \]

\[ C_1 = \text{capacity of first capacitive branch,} \]

\[ C_2 = \text{capacity of second capacitive branch,} \]

\[ K_1, K_2, H \text{ and } B \text{ are all constants dependent upon the transistor and circuit configuration.} \]

Equation 4 can be rewritten as:

\[ d = K_1 \sigma + H \frac{R}{\rho} + B \frac{\sigma}{\rho} + K_2 \frac{R}{\rho} \]

where:

\[ \rho = \frac{R_1}{R_2} \]

\[ \sigma = \frac{C_1}{C_2} \]

\[ C = \sqrt{C_1 C_2} \]

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

\[ d_{\text{min}} = 2(\sqrt{K_1 K_2} + \sqrt{B H}) \]

if

\[ \rho^* = \frac{K_1 B}{K_2 H} \text{ and } \sigma^* = \frac{K_2 H}{K_1 B} \]

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 \( (h_{ie}) \) greater than 15. In many configurations, only a single common collector transistor having a positive feedback network need be employed. Reduction of \( d_{\text{min}} \) 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. 

$d_{\text{max}}$ 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 $d_{\text{max}}$ in this embodiment is also attained by employing the modified 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 for 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 provide 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 capacitive 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 components 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 1a, 2a and 3a 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 13a is a circuit diagram of a simple high pass filter that is optimally 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, band pass filters according to the invention.

Many components in the several figures are provided with reference numeral and subscript notations. The same subscript, e.g. $R_{e}$, 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 $I_{e}$ and $V_{e}$ while the output current and voltage are $I_{o}$ and $V_{o}$. In FIGURES 1 and 2, the emitter of transistors 11 and 12 are connected through negative feedback resistances, $R_{ef}$ 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 $R_{eb}$ 16.

The transistor collector is powered from a negative D.C. supply through load resistor $R_{c}$. 21

Conventional hybrid equivalent circuits, FIGURES 2a and 3a, 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 2a, comprises an admittance, $B_{e}G_{m}$ 18, connected between input terminal 19 and ground and a current generator $h_{fe}R_{e} 22$ shunted by load resistor 21 across output terminal 23 and ground. In the common base equivalent circuit of FIGURE 3a, 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, $h_{fe}K_{b}R_{e}$, shunted by load resistor 17.

The common collector equivalent circuit, FIGURE 1a, is a modified hybrid configuration in which input terminal 28 is connected to ground by the parallel combination of admittance 29, $B_{c}G_{m}$, and current generator 31, $I_{c}$, 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, $K_{v}V_{e}$, 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 1a, 2a and 3a, 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:

$$r_{e} = \frac{30 + 3L_{s}}{L_{e}} = \frac{22R_{e}}{V_{e}} \text{ or } r_{e} = \frac{32}{L_{e}}$$

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,
$h_{\text{be}}$ = collector-emitter current gain in the common base circuit,

$$B_1 = \frac{1}{1 + h_{\text{be}}}$$  \hspace{1cm} (8)

$$G_{\text{e}} = \frac{1}{r_e + R_e}$$  \hspace{1cm} (9)

$$h_{\text{be}} = \frac{h_{\text{be}}}{1 + h_{\text{be}}} = 1 - B_1$$  \hspace{1cm} (10)

$$K_e = K_{\text{be}} = \frac{1 - r_e}{R_e}$$  \hspace{1cm} (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 $r_e$, $K_{\text{be}}$, etc., such as $r_{c2}$, $r_{e2}$, $K_{\text{be}2}$, $K_{\text{ce}2}$. 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 $n^{\text{th}}$ 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, $R_o$, is provided with capacitor 41, $C_o$, connected between the transistor 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:

$$G_{\text{e}}(s) = \frac{h}{1 + ds + s^2}$$  \hspace{1cm} (12)

where $d$ is defined by Equation 4 and $h=1$. The values of the $K_{\text{be}}$, $K_{\text{ce}}$, $B$ and $H$ constants in Equation 4 for the FIGURE 4 configuration are:

$$K_1 = \frac{r_e}{R_e}$$

$$K_2 = 1$$

$$B = \frac{B_1}{1 + R_1}$$

$$H = 1$$

The values of $r_e$ and $B_1$ for transistor 38 are identical with those for the common collector equivalent circuit shown in FIGURE 1a. Of course, $R_1$ and $R_2$ are the values of resisters 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

$$W_c = \frac{1}{\sqrt{R_e C_o}}$$

To provide maximum circuit Q, the value of $d_{\text{min}}$ is selected in accordance with Equation 6. As a practical matter, for the circuits disclosed by FIGURES 5, 6, 8, 9, 10, 11 and 13, where more than one transistor stage is employed, it has been found that $d_{\text{min}}$ 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 $h_{\text{be}}$ values of at least 15 are utilized. However, for the

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 $r_{c1}/R_{c1}$ so the required Q values are attained, resistor $R_{c2}$, in the emitter circuit of transistor 44 is connected to positive D.C. source $V_{cc}$ at terminal 47. It is seen that the value of $r_{c1}/R_{c1}$ is minimized with this configuration when it is considered from Equation 7 that $r_{c1}$ is inversely related to $V_{cc}$. 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 $K_{\text{be}}$, and $B$ which have magnitudes indicated by:

$$K_1 = \frac{r_{c1}}{R_{e1}}$$

and

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

where: $r_{c1}$ and $r_{c2}$ are respectively the values of $r_c$ 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, $R_p$, 53. Bias potential for the emitter of transistor 52 is supplied by the positive D.C. voltage at terminal 54 through resistor 55, $R_6$. 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, $R_6$, 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:

$$K_1 = \frac{r_{c1} + r_{c2}}{R_{c1} + R_{c2}}$$

$$K_2 = 1$$

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

$$H = B_1 + \frac{r_{c1}}{R_{e1}}$$

$$K_3 = 1$$

$$W_c = \frac{1}{\sqrt{R_1 C_1 C_2}}$$

Again, suitable Q values are attained if transistors 52 and 58 have values of $h_{\text{be}}$ 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, $R_{eb}$, is returned to the positive voltage at bus 54 to minimize $r_e/R_e$. Feedback voltage is derived across resistor 67, $R_{eb}$, 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;

$$K_1 = \frac{r_e}{R_{ea}} + \frac{r_e}{R_{eb}} + \frac{r_e}{R_{eb}}$$
$$B = B_2B_3\left(\frac{1 + \frac{R_1}{R_{eb}}} + \frac{B_2R_1}{R_{eb}}\right)$$

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 it is illustrated a high pass filter having an output-input transfer function given by:

$$G_o(s) = \frac{h_s}{1 + ds + s^2}$$  \hspace{1cm} (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,

$$\frac{R_2}{a} \text{ and } \frac{R_3}{1-a}$$

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 $R_b$, a quantity appearing in basic design Equations 4–6. Positive feedback less than one for the emitter voltage developed across resistor 77, $R_{eb}$, 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:

$$K_1 = 1$$
$$K_2 = \frac{r_e}{R_e} + \frac{r_e}{R_2}$$
$$B = B_1$$
$$H = 1 + \frac{B_1 R_2}{R_2}$$

The constant term in the numerator of Equation 9, $h$, equals 1 and cut off frequency, $W_m$, 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, $R_{eb}$, 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, $R_{eb}$. In FIGURE 9, all parameter values are identical with those in FIGURE 8 except $K_2$, $B$ and $H$ which are:

$$K_2 = \frac{r_e}{R_{ea}} + \frac{r_e}{R_{eb}} + \frac{r_e}{R_{eb}}$$
$$B = B_1B_2 + \frac{r_e}{R_2}$$
$$H = 1 + \frac{B_1 R_2}{R_{ea}}$$

and

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

$$G_o(s) = \frac{1 + ds + s^2}{1 + ds + s^2}$$  \hspace{1cm} (14)

where: $d$ is given in Equation 5,

$$b = K_{th}\phi + H\phi + B_2\phi + \frac{K_{th}}{\phi}$$  \hspace{1cm} (15)

and $\phi$ and $\rho$ 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, $R_b$, in parallel with series connected capacitors 88 and 89, $C_1$ and $C_2$. A voltage divider including resistances 91 and 92, $R_1$ and $R_3$, connects the emitter transistor 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 $W_m = 1/RC$ while the slope of the filter skirts is inversely related to $b$. The center frequency attenuation, $G$ $(W_m)$, 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:

$$K_1 = 1$$
$$K_{th} = 1$$
$$B = B_1$$
$$H = 1$$

As in the other embodiments, $d_{max}$ 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:

$$K_1 = 1 + \frac{B_1 R_3}{R_{ea}}$$
$$K_{th} = 1 - a + \frac{B_1 R_3}{R_{ea}}$$
$$B = a B_1 B_3$$
$$H = 1 + \frac{B_1 R_3}{R_{ea}}$$

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

$$G_o(s) = \frac{h_s + f}{1 + ds + s^2}$$  \hspace{1cm} (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, $R_p$. 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 collector circuit via resistor 103, $R$. Output and feedback voltage at the emitter of transistor 97 is derived across resistor 104, $R_e$. 

$$b = K_{th}\phi + H\phi + B_2\phi + \frac{K_{th}}{\phi}$$  \hspace{1cm} (15)
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, \( R_{ep} \) 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,

\[
h = \frac{1 + \rho \cdot \sigma}{\epsilon}
\]

and the cut-off frequency of the passband for these circuits is given by \( w_0 = 1/RC \).

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 \( r_e/Rh \) 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 \( C_1, C_2 \), of the simple RC filter network shown in FIGURE 13a to the emitter of transistor 97 or 106. Shunt resistor \( R_0 \), of the simple filter is connected from capacitor \( C_1 \) or \( C_2 \) to ground. By selecting \( C_2 = (C_1 + C_0) \) and \( R_0 = R_1 \), the constant \( f \) term in the numerator of Equation 16 is eliminated so the band pass transfer function of

\[
G_a(s) = \frac{h_o}{1 + ds + s^2}
\]

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:

\[
G_a(s) = \frac{h_o}{1 + ds + s^2}
\]

Such as disclosed in FIGURES 4–17, 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 \( C_0 \) to the base of common emitter PNP transistor 114. Base bias is established with the positive D.C. voltage at terminal 115 through resistors 116 and 117,

\[
\frac{R_1}{1 - \alpha} \quad \text{and} \quad \frac{R_1}{\alpha}
\]

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, \( R_{eb} \), and load resistor 121, \( R_b \), 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:

\[
K_1 = \frac{1}{\rho} \frac{1}{K}; \quad K_2 = \frac{1 + B_s a (1 - a)}{K}; \quad \frac{B}{\rho} = \frac{B_s}{K}; \quad H = 0; \quad h = \frac{-a (1 - a) \sigma}{\mu K}; \quad w_0 = \frac{K}{RC}
\]

where, by definition:

\[
K^2 = [\rho^2 + a(1 - a)] (1 + \frac{B_1 R_1}{R_{es}} + \frac{a R_1}{R_{es}})
\]

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,

\[
\frac{R_1}{a} \quad \text{and} \quad \frac{R_1}{1 - \alpha}
\]

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:

\[
K_1 = \frac{1 + B_s a (1 - a)}{\rho^2} \quad ; \quad K_2 = \frac{1 + B_s a (1 - a)}{K}; \quad \frac{B}{\rho} = \frac{B_s}{K}; \quad H = 0; \quad h = \frac{-a (1 - a) \sigma}{\mu K}; \quad w_0 = \frac{K}{RC}
\]

where:

\[
K^2 = [\rho^2 + B_s a(1 - a)] (1 + \frac{B_1 R_1}{R_{es}} + \frac{a R_1}{R_{es}})
\]

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 \( B_s a \) 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 \( d_{mho} \), prevents substantially loading of transistor 114.
The parameters of FIGURE 16 are given by:

\[
\begin{align*}
K_1 &= \frac{1 + B_1 R_2}{K}, \\
K_2 &= \frac{1 + B_1 R_2 (1 - a)}{K}, \\
R_{eq} &= \frac{B}{R} = \frac{B_2 R_2}{K}, \\
H &= 0, \\
h &= -\sigma (1 - a) \mu K, \\
w_e &= \frac{K}{RC}.
\end{align*}
\]

where:

\[
K' = \left[ a^2 + B_2 R_2 d(1 - a) \right] \left( 1 + \frac{B_1 R_1}{R_{eq}} \right) + a R_1 + \frac{a R_1}{R_{eq}}
\]

Another manner for attaining band-pass characteristics 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_{eq}, to ground. Collector bias for transistor 133 is supplied from the negative, D.C. voltage on bus 135 through load resistor 136, R'_e. Base bias is provided with the voltage divider comprising resistors 137, R'_e, 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_{eq}, 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:

\[
\begin{align*}
K_1 &= K_2 = \frac{1}{K}, \\
B &= H = 0, \\
h &= R_1 \mu K; \text{ and} \\
w_e &= \frac{K}{RC},
\end{align*}
\]

where:

\[
K = 1 + \frac{B_1 R_2}{R_{eq}} R_{eq} \quad R_1 = \frac{R_1}{1 - \frac{B_1 R_2}{R_{eq}}}
\]

\[
R_2 = \frac{B_2}{1 - \frac{B_2 R_2}{R_{eq}}}
\]

The circuit of FIGURE 17 is modified according to FIGURE 18 by connecting emitter follower 145, having resistor 146, R_{eq}, 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_{eq} in the former is replaced with R_e in the latter and

\[
R_e = \frac{R_2}{1 - \frac{B_2 R_2}{R_{eq}}}
\]

I claim:

1. An active filter for audio signal provided by an audio source, said filter having only capacitative 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 capacitative 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:

\[
d = \frac{K_1 R_2 C_1 + H R_3 C_2 + B R_3 C_1 + K_2 R_2 C_2}{\sqrt{R_1 R_2 C_1 C_2}}
\]

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 shorting 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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U.S. Cl. X.R
330--20, 25, 26, 28, 31; 333--70