ACTIVE FILTERS WITH VOLTAGE-VARIABLE PASSBAND FOR APPLICATION TO RANGE-GATED MTI SYSTEMS

by

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UNITED STATES
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THESIS

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ABSTRACT

Range-gated moving target indicator (MTI) systems have some intrinsic advantages over delay-line canceler MTI systems. A further improvement in performance would result if the doppler cut-off frequency was adjustable by the radar operator. Because of the large number of filters involved in one system this would require a voltage-variable filter element.

A theory of variable filter response developed by Thiele was used as the basis for design using integrated-circuit operational amplifiers as the active elements. Three different types of active filter circuits are discussed. Implementation of voltage-variable filters and experimental frequency-response curves using field-effect transistors, Raysistors, and discrete switched resistors are described and illustrated.

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I. RANGE-GATED MTI SYSTEM

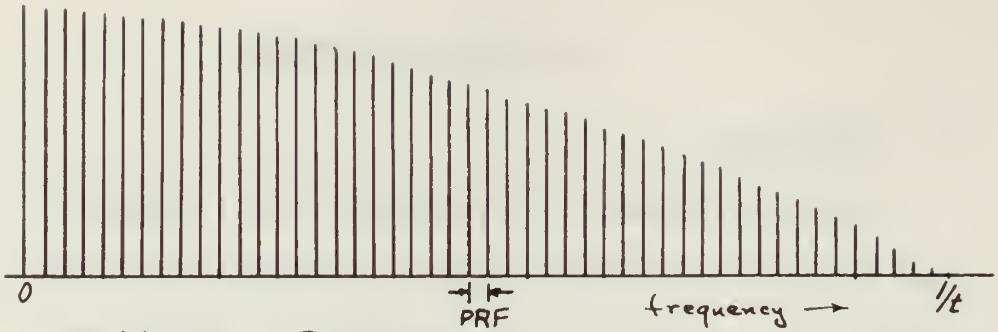
A. INTRODUCTION

Recent developments in integrated-circuit technique have brought about renewed interest in range-gated moving target (MTI) systems for pulse radars.[1,2,3] Unlike the delay-line canceler MTI which performs its filtering in the time domain, the range-gated MTI system performs its filtering in the frequency domain. This frequency-domain filtering would cause the loss of range information except that by using a timed gating system and a number of identical processing channels this range information is retained.

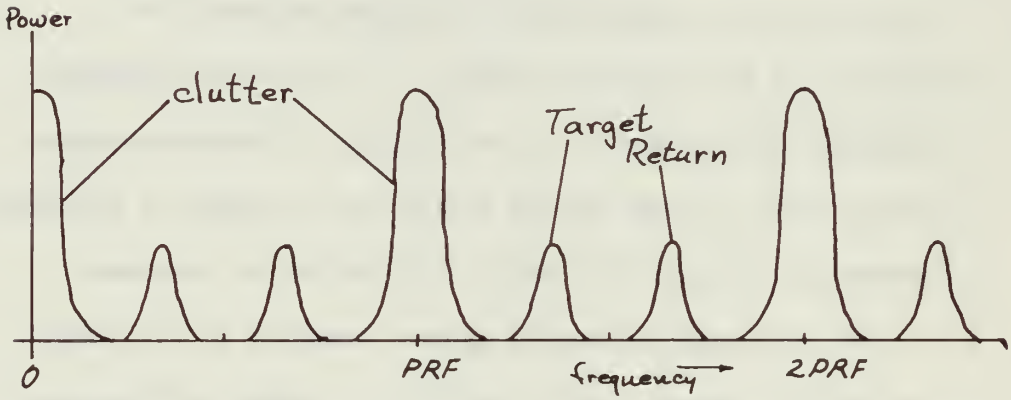
The advantage of such a system compared to the delay-line canceler stems from the ability to shape the frequency response of the filter to a more optimum one than can be obtained using a reasonably small number of delay lines. In at least one prototype range gated system an increase of fifteen decibels in clutter attenuation was obtained.[2]

If a video bandwidth greater than or equal to the reciprocal of the radar pulse width is assumed, the detected video signal of a pulse radar can be represented by the spectrum shown in Figure 1 (a). With no moving target return present the spectrum consists of clutter return centered about zero frequency (D.C.) and about multiples of the pulse repetition frequency.

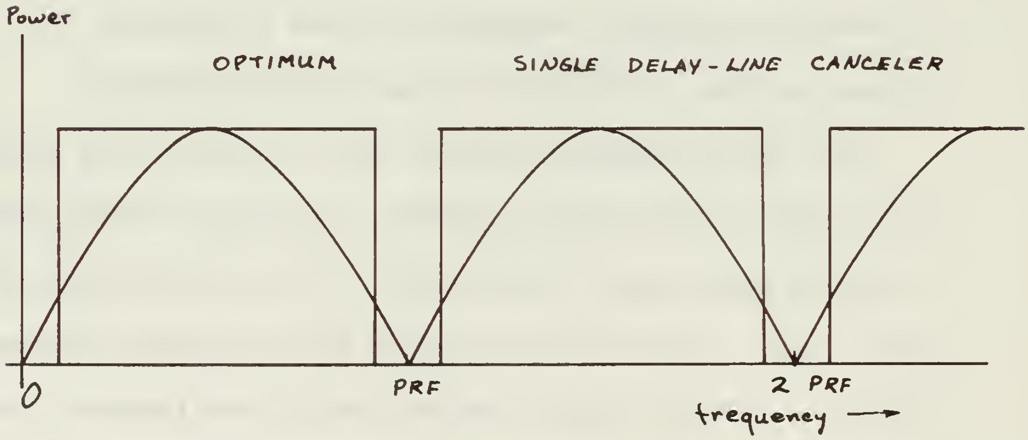
The power of this spectrum generally follows a $\frac{\sin^2 \omega t/2}{(\omega t/2)^2}$ envelope which is maximum at zero frequency and zero at the



(a) Video Spectrum



(b) Enlarged Clutter and Target Spectrum



(c) OPTIMUM AND SINGLE DELAY-LINE CANCELER RESPONSE

Figure 1. Some Frequency Spectra

reciprocal of the pulse width, t. The spacing of the diagram is somewhat exaggerated since for typical pulse radars the pulse repetition frequency is of the order of one one-thousandth of the reciprocal of the pulse width. This means that there would be about one thousand spectral lines shown rather than the few actually depicted.

Figure 1(b) represents a portion of this spectrum enlarged to show frequencies up to twice the pulse repetition frequency. The spectral lines of Figure 1 (a) are shown as bands since the return echos are modulated by antenna scanning and target motion, including incidental clutter motion. Unlike Figure 1 (a) it also includes representation of a moving target return whose radial velocity with respect to the radar is v. The doppler frequency of this target is given by $f_d = 2v/\lambda$ where λ is the wavelength of the radar radiation.

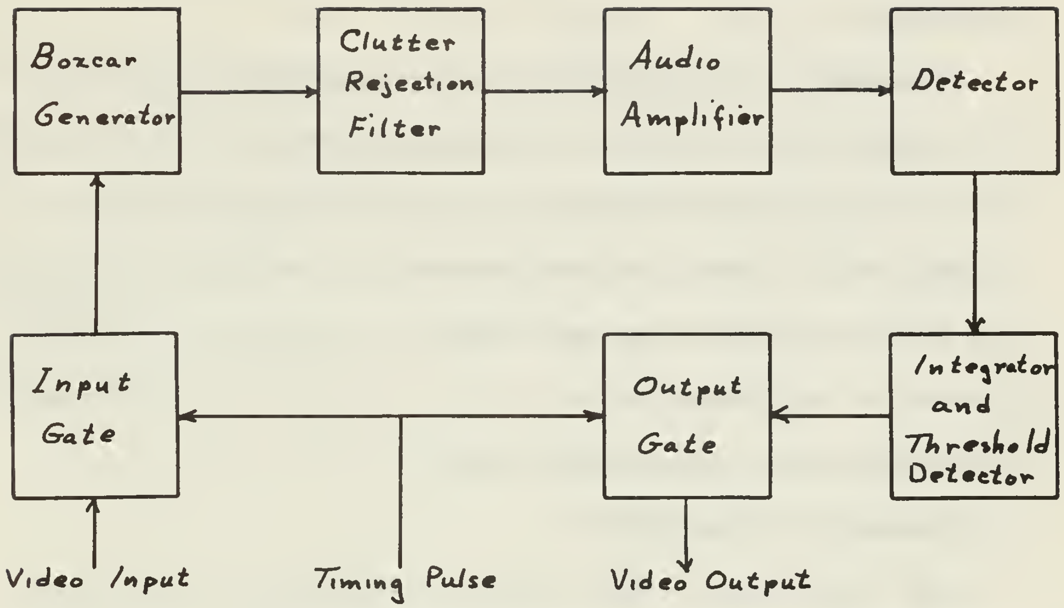
The ideal MTI filter response which would attenuate the radar clutter but allow target signals to be detected is shown superimposed on this spectrum in Figure 1 (c).

Most current MTI systems use a delay-line canceler to accomplish this filtering. This method consists of comparing an incoming video signal with the previous video signal which has been delayed one pulse repetition interval. The difference between the amplitude of successive pulses is allowed to continue as target video. Successive returns from stationary targets should show very nearly the same form and thus ideally they are eliminated. On the other hand, successive returns from moving targets are at slightly different ranges and subtraction does not eliminate the signal.

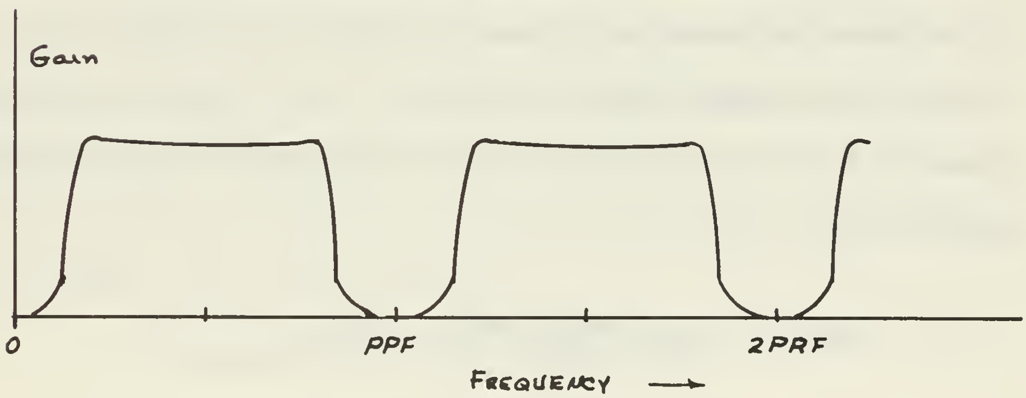
The heart of such a system is the delay line. Delay lines are costly and difficult to construct. Circuitry associated with the delay line requires extremely close tolerances on amplitude and phase control in order to get effective clutter cancellation. Also the pulse repetition frequency must be the reciprocal of the delay time to a high degree of exactness. Moreover, the effective frequency response of the single delay-line canceler as shown in Figure 1 (c) does not come very close to the optimum response. Some improvement can be obtained with a double delay-line canceler and higher-order cancelers are conceptually possible. The difficulties encountered in actual construction generally prohibit employment of these higher-order cancelers.

A typical range gated MTI system is represented in the block diagram of Figure 2 (a). The boxcar generator (pulse stretcher) acts as a low-pass filter. Its action is to hold the gated video signal over the pulse repetition period. This attenuates the frequencies above the pulse repetition frequency while accentuating the frequencies below this frequency. The clutter-rejection filter, in conjunction with the response of the boxcar generator, then determines the shape of the response. An equivalent frequency response can be considered to be that shown in Figure 2 (b). Proper filter design can make this response closely approach the optimum previously described.

The threshold detector provides discrimination against noise and may incorporate generation of standard video pulses when signal



(a) BLOCK DIAGRAM OF RANGE GATED MTI



(b) EFFECTIVE FREQUENCY RESPONSE

FIGURE 2 RANGE GATED MTI

is present, thus presenting very clean, uniform signals for further processing and display, in spite of signal strength variation.

The heart of the system is the bandpass filter which allows the doppler frequencies from moving-target returns to be detected but which filters out unwanted D.C. and low frequencies from clutter return. Since the width of the band of clutter frequencies is a function of the radar environment at the time of operation[4] it would be very desirable to be able to vary the rejection band of the filter over some limited range.

B. PROPERTIES OF THE FILTER

The filter used in each channel should have the following properties:

1. Size. The size of the filter must be small because in a medium-range search radar as many as one thousand identical channels might be required.

One design considers the number of channels equal to the ratio of the effective range to the range resolution (one pulse width).[3] Thus for a particular radar (the AN/UPS-1 search radar) the number of channels that would be required for eighty mile range operation is given by

$$\text{Number} = \frac{\text{Range (nm)}}{\text{Pulse width } (\mu\text{sec}) \times 0.081 \text{ (nm/}\mu\text{sec)}}$$

$$= \frac{80}{1.4 \times 0.081} = 715$$

2. Frequency response. The lower cutoff frequency should be just above the highest clutter frequency. Since clutter is not constant this portion of the response should be variable. The upper cutoff frequency should be at one half the pulse repetition frequency (PRF) of the radar because the spectrum of the received signal is folded about each multiple of the PRF so that it presents a mirror image about each multiple of one half the PRF.

Thus the ideal clutter-rejection filter for a range gated MTI system would be a physically very small constant transmission bandpass filter which passes frequencies between the upper limits of the clutter (which may be different from time to time) and one half the pulse repetition frequency. It might sometimes be desirable to provide for varying the pulse repetition frequency also. This is difficult with the delay-line canceler but is quite feasible with range gated systems.

Since the filter passband is in the very low audio range, inductors cannot be used because of size and weight limitations. The remainder of this paper develops a design procedure for filters which meet the above requirements and includes performance data for some typical filter configurations.

II. VARIABLE FILTERS

A. GENERAL

The break (or corner) frequency of a filter is ordinarily determined by two circuit elements for each pole. To change this break frequency it is necessary to change at least one of these circuit elements per pole. Thus a two-pole filter requires adjustment of at least two circuit elements; similarly, a three-pole filter requires adjustment of at least three components. A continuously variable frequency response imposes severe component tracking requirements. Discrete frequency-response changes are not as difficult to achieve, but the size and complexity of the circuit is then determined by the number of frequency-band changes required.

The variable filters described in this paper are based on a technique developed by Thiele.[5] Use of this method permits moving the break frequency of a three-pole filter by adjusting a single element.

B. BUTTERWORTH AND CHEBYCHEV FILTERS

Two widely used filters are the Butterworth (maximally flat) response and the Chebychev (equal ripple) response.

1. Butterworth. In conventional filter theory the Butterworth low-pass filter response is given by

$$|G(j\omega)|^2 = \frac{1}{1 + \frac{\omega^{2n}}{\omega_o^{2n}}}$$

where n is the number of poles and ω_o is the angular break frequency. It derives the name maximally flat because as ω approaches zero so do all the derivatives.

The pole configuration in the s-plane consists of poles equally spaced on a semi-circle in the left-hand plane. The radius of the semi-circle is ω_o and the center is at the origin.

If conjugate poles are considered in pairs the response for the i'th pair is given by

$$G_i(j\omega) = \frac{1}{(\sigma_i'^2 + \omega_i'^2) + 2\,\sigma_i'(j\omega) + (j\omega)^2}$$

This transfer function can be scaled in frequency by multiplying through by ω_o^2.

$$G_i(j\omega) = \frac{1}{\left(\dfrac{\sigma_i'^2}{\omega_o^2} + \dfrac{\omega_i'^2}{\omega_o^2}\right) + 2\,\dfrac{\sigma_i'}{\omega_o}\,\dfrac{(j\omega)}{\omega_o} + \left(\dfrac{j\omega}{\omega_o}\right)^2}$$

where $\frac{\sigma_i'}{\omega_o}$ and $\frac{\omega_i'}{\omega_o}$ are the coordinates of the poles on the semi-circle scaled to unity radius. These normalized values are denoted by $\sigma_i = \frac{\sigma_i'}{\omega_o}$ and $\omega_i = \frac{\omega_i'}{\omega_o}$. These values are tabulated in standard circuit theory texts.[6]

Thus $$G_i(j\ ) = \frac{1}{\left(\sigma_i^2 + \omega_i^2\right) + 2\,\sigma_i\left(j\dfrac{\omega}{\omega_o}\right) + \left(\dfrac{j\omega}{\omega_o}\right)^2}$$

When there is an even number of poles the Butterworth response consists of n/2 such factors. When n is odd the response consists of the product of (n-1)/2 such factors and a single factor of the following form:

$$G_i(j\omega) = \frac{1}{\sigma_i' + j\omega} = \frac{1}{\frac{\sigma_i'}{\omega_o} + j\,\frac{\omega}{\omega_o}} = \frac{1}{1 + j\,\frac{\omega}{\omega_o}} \quad \text{(scaled)}$$

since $\sigma_1' = \omega_o$.

2. Chebychev. If the real part of each pole is multiplied by a factor $e < 1$ the poles will be shifted in the s-plane and they will lie on a semi-ellipse. This semi-ellipse is the pole configuration of a Chebychev filter response.

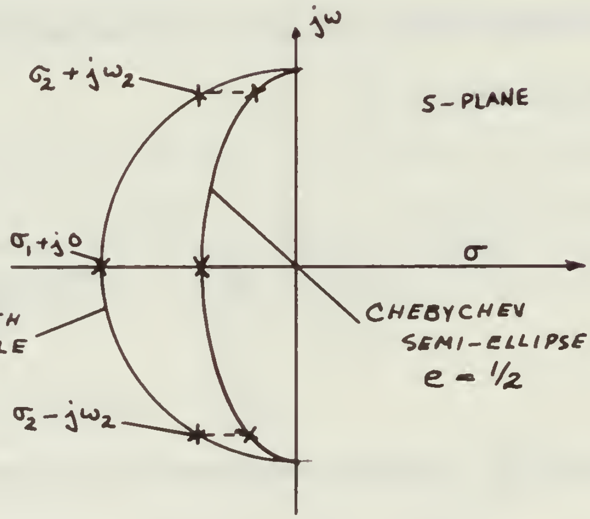


Figure 3. Three pole Butterworth filter transformed to a Chebychev three pole filter.

Within the passband where the Butterworth filter is maximally flat, the Chebychev filter has an equal ripple response. It approximates the ideal filter response by alternating between two values in the passband. The total number of maxima and minima is given by $|\Delta G(j\omega)| = \frac{1}{\sqrt{1 + e^2}}$.

C. A VARIABLE FILTER

With an alternate transformation of Butterworth to Chebychev due to Thiele[5], the break frequency can be made a function of a single circuit element. A semi-elliptical pole configuration in the s-plane can be obtained by multiplying the imaginary portion of the roots of the Butterworth filter by a factor $k > 1$. From the pole diagram alone it can be seen that this method has two significant features that differ from the previous method:

1. Varying k now affects the frequency response a great deal more.

2. The root on the real axis is unaffected by k.

Since the phasor $G_{i+}(j\omega)$ is given by

$$G_{i+}(j\omega) = \sigma_i + j\left(k\omega_i - \frac{\omega}{\omega_o}\right)$$

the gain from the conjugate roots becomes

$$G_i(j\omega) = \frac{1}{\left(\sigma_i^2 + k^2\omega_i^2\right) + 2\sigma_i j\frac{\omega}{\omega_o} + \left(\frac{j\omega}{\omega_o}\right)^2}$$

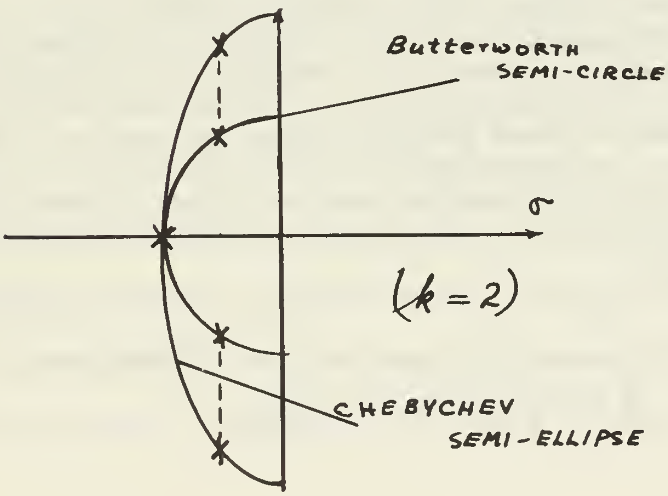


Figure 4. Chebychev poles derived from Butterworth poles by Thiele method.

The difference between this and the Butterworth transfer function is contained entirely within the zero order term. Thus if this term alone could be varied by changing a single parameter the response would shift from that of a Butterworth (k = 1) to that of a Chebychev response.

For convenience a new parameter K is defined:

$$\sigma_i = \cos\theta_i \qquad \text{and} \qquad \omega_i = \sin\theta$$

and $\sigma_i^2 + \omega_i^2 = 1$

Thus $\sigma_i^2 + k^2\omega_i^2 = \sigma_i^2 + \omega_i^2 + (k^2 - 1)\,\omega_i^2 = 1 + K\omega_i^2$

where $K = k^2 - 1$.

When K = 0 and k = 1 and the response is that of a Butterworth filter.

$$\text{Thus} \quad G_i(j\omega) = \frac{1}{(1 + K\omega_i^2) + 2\sigma_i \frac{j\omega}{\omega_o} + \left(j\frac{\omega}{\omega_o}\right)^2}$$

where σ_i and ω_i are coordinates of the relevant normalized Butterworth poles and K is a bandwidth widening factor.

The ratio of the three-decibel frequency, f_c, of the resulting Chebychev response to that of the original Butterworth, f_o, is given by $f_c/f_o = \sqrt{K + 1}$ when f_c is defined as the frequency at which the response is three decibels below the mean of the passband.

For convenience the response can be written in the following operational form:

$$G_i(s) = \frac{1}{\left(1 + K\omega_i^2\right) + 2\sigma_i \frac{s}{\omega_o} + \frac{s^2}{\omega_o^2}} \qquad \text{(Low Pass)}$$

Thus far the analysis has been for a low-pass filter. The variable response desired in the doppler filter is that of a high-pass filter. The function above can be transformed to a high-pass response by replacing s/ω_o by ω_o/s throughout.

$$G_i(s) = \frac{s^2}{(1 + K\omega_i^2)s^2 + 2\sigma_i s\omega_o + \omega_o^2} \quad \text{(High Pass)}$$

One factor which has been ignored up to this point is the insertion loss of the filters described. For each of the above gain factors an insertion loss is easily computed. For the low-pass function as s goes to zero $G_i(s)$ goes to $1/(1 + K\omega_i^2)$. Similarly for the high-pass function as s goes to infinity $G_i(s)$ goes to $1/(1 + K\omega_i^2)$. Thus to compensate for the dependence on K of the insertion loss the form of transfer function that is required is the following:

$$G_i(s) = \frac{(1 + K\omega_i^2)s^2}{(1 + K\omega_i^2)s^2 + 2\sigma_i\omega_o s + \omega_o^2}$$

The results of the above analysis indicate that if K can be varied the transfer function can be changed from a Butterworth response to varying forms of Chebychev response. If K is allowed to become slightly negative the response becomes what Thiele calls sub-Chebychev and over a limited range it remains useful in narrowing the bandpass. If K becomes too negative, however, the slope of the response approaches its asymptotic value too slowly to be useful.

D. APPLICATION

Most doppler filters used in recently developed range gated MTI systems have two-pole response for the high- and low-pass filters.[2,3] Since with the scheme described above the odd-numbered pole poses no particular added requirement for control, the three-pole filter response was chosen as the design goal. This would give an eighteen decibel per octave slope and the attendant increased clutter attenuation.

The pole coordinates for a third-order Butterworth filter are

$$\sigma_1 = -1$$

$$\sigma_2 \pm j\omega_2 = 1/2 \pm j\frac{\sqrt{3}}{2}$$

Thus the desired three-pole, high-pass transfer function is

$$G(s) = \frac{(1 + 3/4K)s^3}{(s + \omega_o)\,[(1 + 3/4K)s^2 + \omega_o s + \omega_o^2]}$$

Figure 5 shows computed response curves for this function.

In order to translate this function to one in terms of real components consider the following form of circuit transfer function:

$$G(s) = \frac{Axs^3}{(s + c)(xs^2 + as + b)}$$

where x is a single component (resistor or capacitor value) which appears only in the indicated positions. The other coefficients and terms (A,a,b, and c) are functions of other circuit elements only.

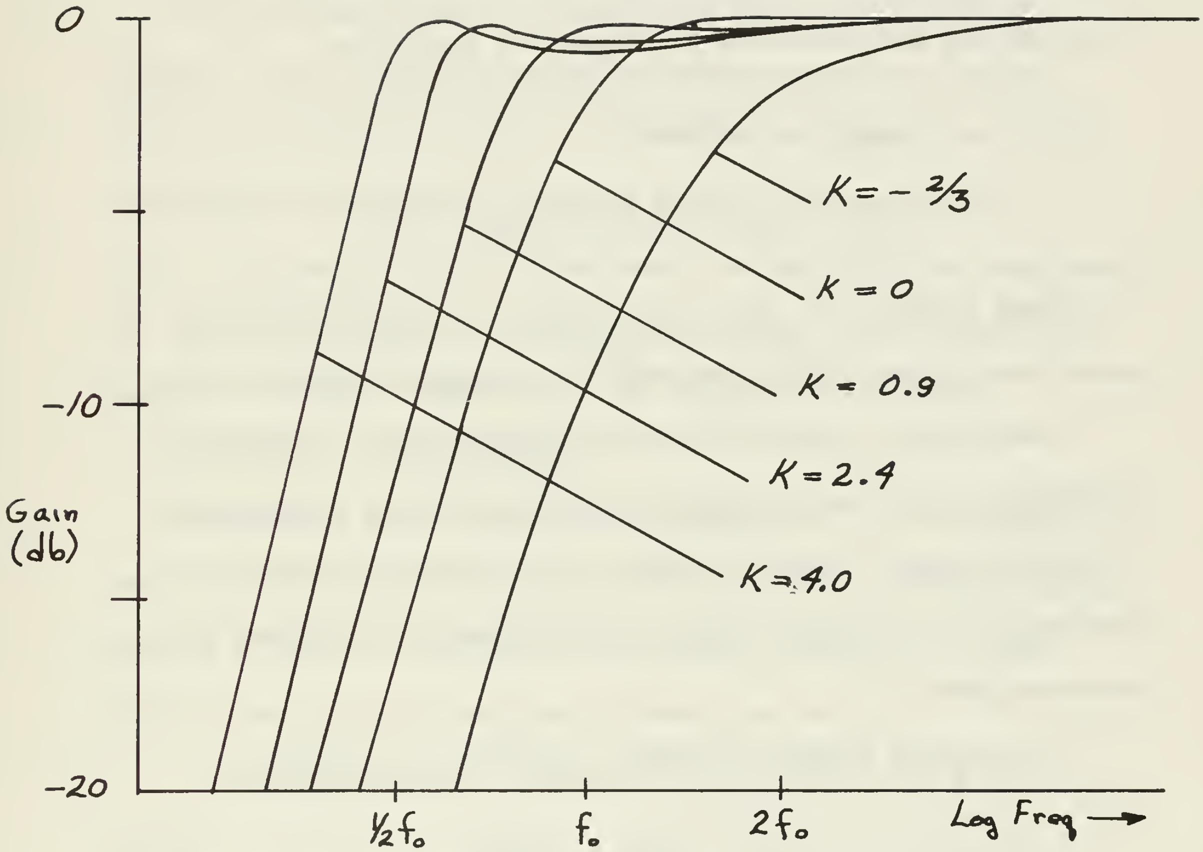


FIGURE 5 THEORETICAL RESPONSE OF THIRD ORDER THIELE FILTER

If one considers X_B as the value of X which corresponds to a Butterworth response, then

$$G(s) = \frac{A(X/X_B)s^3}{(s + c)\ [(X/X_B)s^2 + \frac{a}{X_B}s + \frac{b}{X_B}]}$$

Hence the component values must be chosen such that X is variable and the following relations are satisfied:

$$X/X_B = 1 + 3/4K$$

$$a/X_B = \omega_o,\ b/X_B = \omega_o^2,\ \text{and } c = \omega_o.$$

From Figure 5 it can be seen that a useful variation of about two octaves can be obtained by varying K from -2/3 to 4.

The bandpass doppler filter was designed as a cascade of double-pole high-pass, single-pole high-pass, and double-pole low-pass stages. Thus the design task consists primarily of obtaining second-order responses of the desired high-pass and low-pass form.

The overall bandpass transfer function is given by

$$G(s) = \frac{A(1 + 3/4K)s^3}{[(1 + 3/4K)s^2 + \omega_o s + \omega_o^2](s + \omega_o)(s^2/\omega_o^2 + \frac{s}{\omega_o} + 1)}$$

This provides a Butterworth second-order, high-frequency falloff.

III. ACTIVE FILTERS

A. INTRODUCTION

Based on the type of active element employed there are currently three different approaches to active filter design: (1) Negative immittance converter technique, (2) Gyrator technique, and (3) Operational amplifier technique. Since integrated-circuit operational amplifiers are readily available this latter technique was selected for this investigation. Within this general method three basic variations were considered: (1) Integrator-summer building-block method, (2) Operational amplifier feedback method, and (3) Sallen-Key circuits.

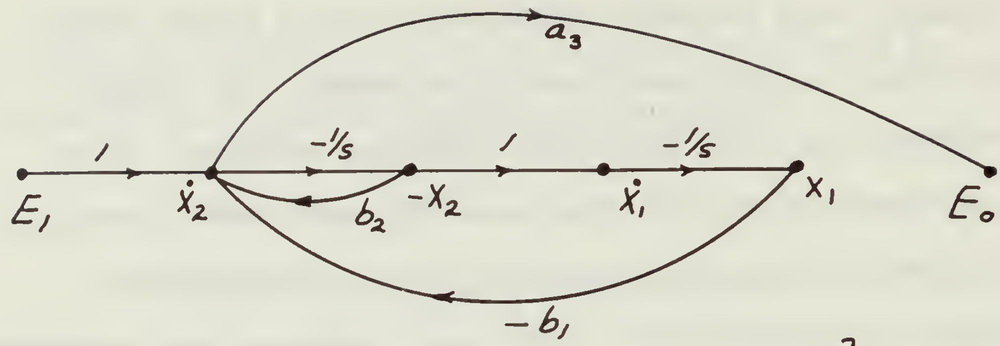
B. INTEGRATOR-SUMMER TECHNIQUE

Integrators and summers can be considered as the building blocks for realizing any transfer function. The discussion given below is a specific application of results derived in a more complete and general treatment by Kerwin et. al.[7]

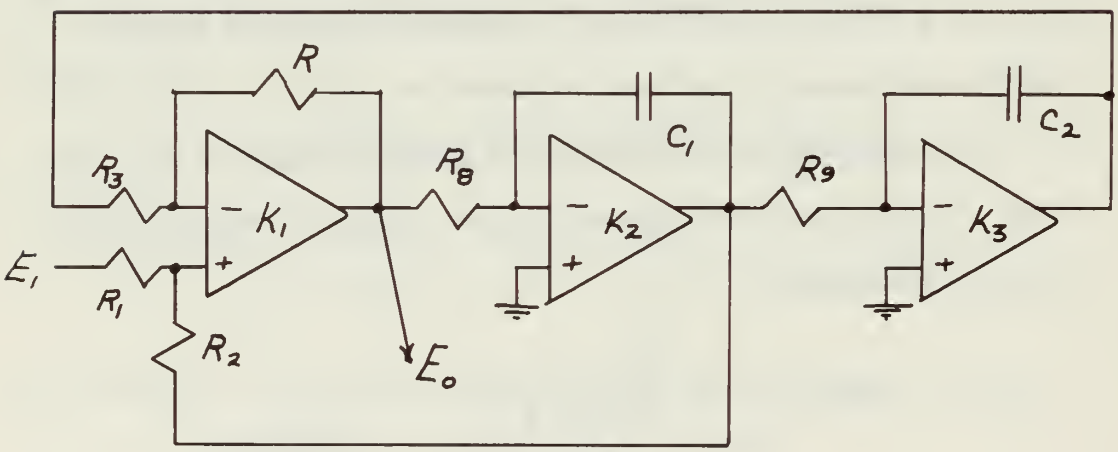
The flowgraph of the function is given by Figure 6 (a). In terms of components this results in the circuit of Figure 6 (b). In this realization

$$G(s) = \frac{C_1 s^2}{C_1 s^2 + \dfrac{R_1(R + R_3)}{R_3 R_8 (R_1 + R_2)} s + \dfrac{R}{R_3 R_8 R_9 C_2}}$$

which is nearly in the desired form for the variable response desired.



(a) FLOWGRAPH OF $\frac{E_0}{E_1} = \frac{a_3 s^2}{s^2 + b_2 s + b_1}$



(b) A CIRCUIT CONFIGURATION OF ABOVE FLOWGRAPH

FIGURE 6 INTEGRATOR - SUMMER CIRCUIT

Let C_B be the value of C_1 required for the Butterworth response of the system. Then

$$G(s) = \frac{(C_1/C_B)s^2}{(C_1/C_B)s^2 + \frac{R_1(R_1 + R_3)}{R_3R_8(R_1 + R_2)C_B} s + \frac{R}{R_3R_8R_9C_2C_B}}$$

Thus to obtain the proper response

$$1 + 3/4\ K = C_1/C_B \tag{1}$$

$$\omega_o = \frac{R_1(R + R_3)}{R_3R_8(R_1 + R_2)C_B} \tag{2}$$

and

$$\omega_o^2 = \frac{R}{R_3R_8R_9C_2C_B} \tag{3}$$

Kerwin et. al.[7] shows that a positive-gain integrator can be constructed. Shown in Figure 7, this integrator is considerably more complex than the standard negative-gain configuration, but it has the advantage that when used to obtain the above transfer function the variable element C_1 has one grounded terminal.

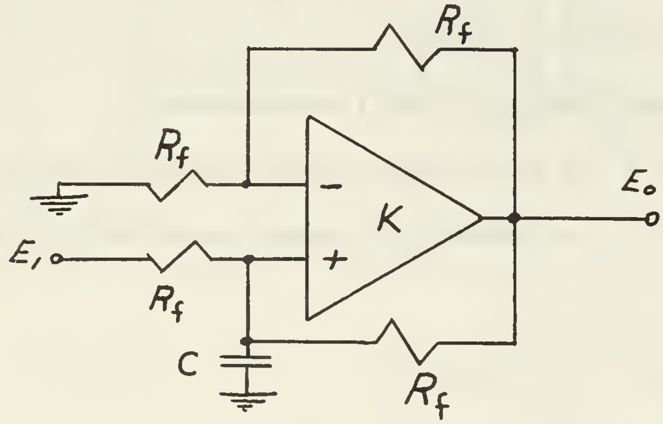


Figure 7. Positive-Gain Integrator.

Using this technique (either positive- or negative- gain integrators) requires a minimum of six operational amplifiers, twelve resistors (twenty four for positive gain integrators) and five capacitors for the complete bandpass filter. Even though the stability of this system is high it is not considered desirable for the doppler filter due to the large number of components required.

A two-pole high-pass circuit using this technique was designed, however. In order to eliminate the large number of variables the following arbitrary decisions were made concerning component values.

1. $R = R_1$
2. $R_2 = R_3 = 10R$
3. $f_o = 50$ hertz ($\omega_o = 2\pi f_o = 314$ radians/second)
4. $1 + 3/4\ K = C_1/C_B$ to range from 1/2 to 4

These assumptions substituted in equation (2) above give

$$\omega_o = 1/10R_8C_B = 314 \qquad (2')$$

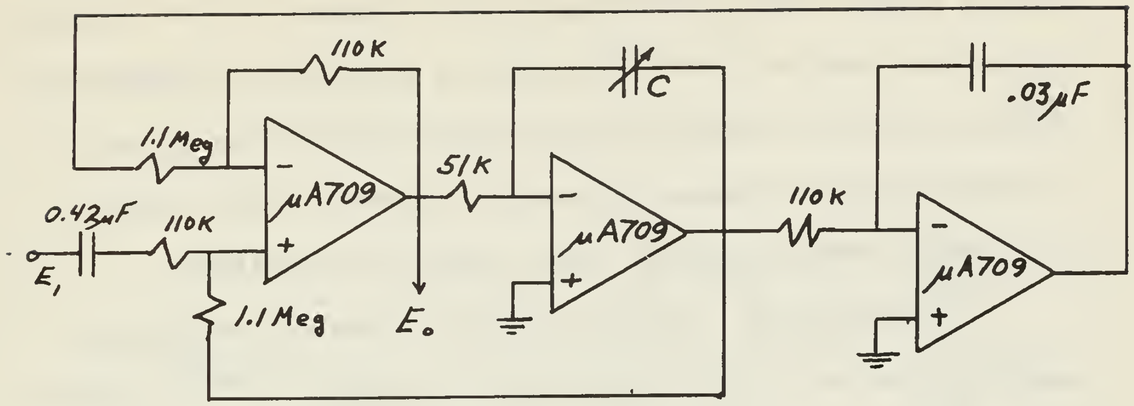
Thus if $R_8 = 50$ kilohms and $C_B = 0.00635$ microfarads. From equation (3) and the assumptions

$$\omega_o^2 = 1/10R_8R_9C_2C_B \qquad (3')$$

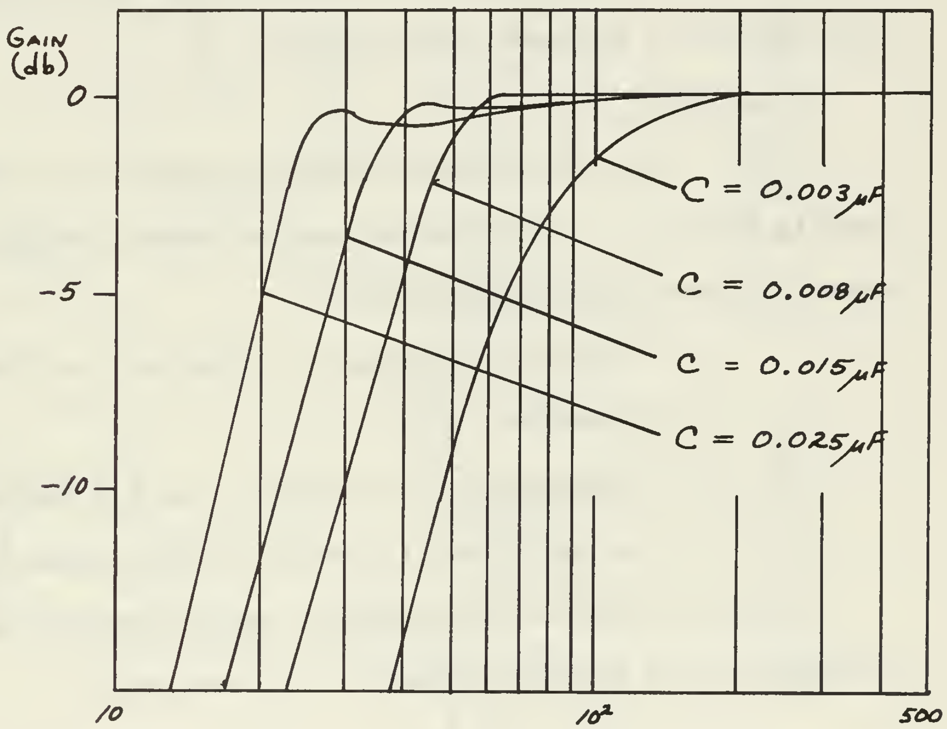
Dividing (3') by (2') $\omega_o = 1/R_9C_2$.

If $R_9 = 100$ kilohms then $C_2 = 0.0317$ microfarads.

Requiring $(1 + 3/4\ K)$ to range from a low value of 1/2 to a high of 4 results in requiring C_1 to range from $0.00317\mu F$ to $0.0254\mu F$.



(a) CIRCUIT CONFIGURATION (LESS FREQUENCY COMPENSATION COMPONENTS)



(b) FREQUENCY RESPONSE

FIGURE 8 INTEGRATOR - SUMMER CIRCUIT

The experimental results obtained using this configuration and the actual circuit values are shown in Figure 8. The resulting curves show that the corner frequency is a function of the control element (a decade capacitor in this case) and that the slope of the rejection-band response is nearly constant for all curves.

C. GENERAL FEEDBACK CIRCUIT USING OPERATIONAL AMPLIFIERS

1. Introduction. Various "general" forms of operational amplifier feedback circuits have been presented.[8,9,10,11] Most of these in fact assume a priori some limited and fairly specific configuration. Specifically one limitation sometimes imposed is that requiring a grounded input amplifier.[9]

2. Analysis

a. A General Feedback Circuit. Consider the circuit given in Figure 9. The following standard assumptions for ideal operational amplifiers will be made:

1. The input impedance is infinite; thus the amplifier will draw no current.

2. The gain, K, is infinite; thus E_A, the voltage input to the amplifier, is zero for finite output voltage, E_2.

From the operational amplifier assumptions and the network configuration the following relations are obtained:

$$E_2 = E_2^d = E_2^c$$

$$E_A = E_2^c = E_1^d = 0$$

$$E_1 = E_1^a$$

$$E_1^b = E_2^a = E_1^c$$

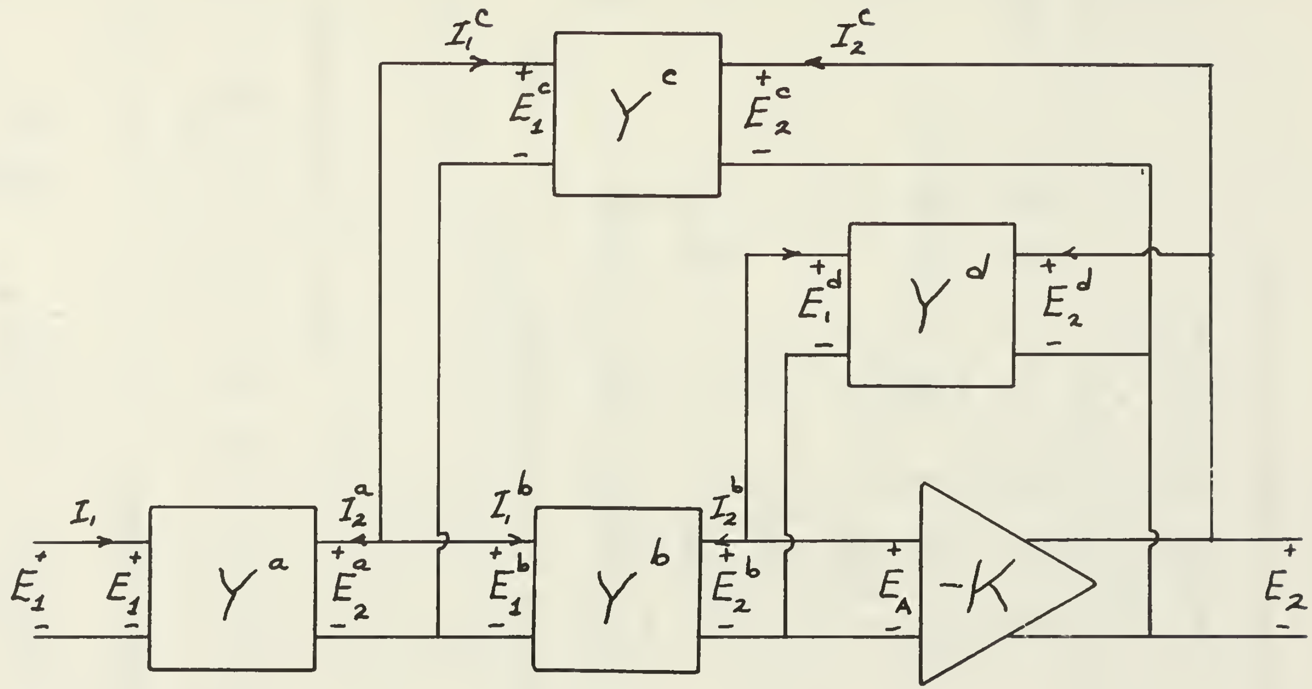


FIGURE 9 GENERAL FEEDBACK CIRCUIT

and

$$I_1 = I_1^a$$

$$I_2^a = -(I_1^c + I_1^b)$$

$$I_2 = I_2^c + I_2^d$$

$$I_1^d = -I_2^c$$

The equations of two-port network theory are

$$I_1^a = y_{11}^a E_1 + y_{12}^a E_2^a \quad (1)$$

$$I_2^a = y_{21}^a E_1 + y_{22}^a E_2^a \quad (2)$$

$$I_1^b = y_{11}^b E_1^b + y_{12}^b E_A = y_{11}^b E_1^b \quad (3)$$

$$I_2^b = y_{21}^b E_1^b + y_{22}^b E_A = y_{21}^b E_1^b \quad (4)$$

$$I_1^c = y_{11}^c E_1^b + y_{12}^c E_2 \quad (5)$$

$$I_2^c = y_{21}^c E_1^b + y_{22}^c E_2 \quad (6)$$

$$I_1^d = y_{11}^d E_A + y_{12}^d E_2 = y_{12}^d E_2 \quad (7)$$

$$I_2^d = y_{21}^d E_A + y_{22}^d E_2 = y_{22}^d E_2 \quad (8)$$

Since $I_1^d = -I_2^b$ equation (7) and (4) yield

$$E_1^b = -\frac{y_{12}^d}{y_{21}^b} E_2 \quad (9)$$

$I_2^a = -(I_1^c + I_1^b)$ results in the following

$$-y_{11}^c E_1^b - y_{12}^c E_2 - y_{11}^b E_1^b = y_{21}^a E_1 + y_{22}^a E_1^b \quad (10)$$

Using equation (9) this becomes

$$y_{21}^a E_1 = \frac{y_{11}^c y_{12}^d}{y_{21}^b} E_2 + \frac{y_{11}^b y_{12}^d}{y_{21}^b} E_2 + \frac{y_{22}^a y_{12}^d}{y_{21}^b} E_2 - y_{12}^c E_2$$

or

$$\frac{E_2}{E_1} = \frac{y_{21}^{a}}{\left(y_{11}^{c} + y_{11}^{b} + y_{22}^{a}\right)\frac{y_{12}^{d}}{y_{21}^{b}} - y_{12}^{c}} \tag{11}$$

b. Some Restrictions. In order to simplify the function derived above each two-port circuit was assumed to be a grounded, reciprocal network. Subject to these restrictions each network can be represented without any further loss of generality by the PI network in Figure 10.

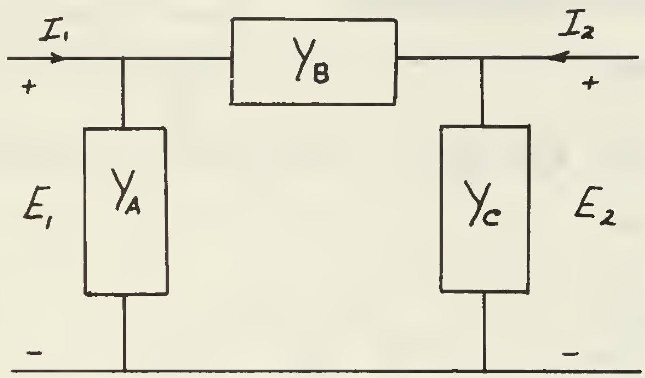


Figure 10. PI equivalent of reciprocal grounded two-port circuit.

For this two-port network

$$y_{11} = Y_A + Y_B$$

$$y_{12} = y_{21} = -\,Y_B$$

$$y_{22} = Y_B + Y_C$$

The transfer of the general network then becomes

$$\frac{E_2}{E_1} = \frac{-\,Y_B^a}{(Y_A^c + Y_B^c + Y_A^b + Y_B^b + Y_B^a + Y_C^a)\ \dfrac{Y_B^d}{Y_B^b} + Y_B^c}$$

It is evident from this function that not all elements of the PI network contribute to the transfer function. Figure 11 (a) shows the general network redrawn using the PI configuration for each two port. The hatched elements are those which do not contribute to the transfer function. Figure 11 (b) combines these elements into a simpler equivalent circuit where

$$Y_1 = Y_B^a$$

$$Y_1 = Y_C^a + Y_A^b + Y_A^c$$

$$Y_3 = Y_B^b$$

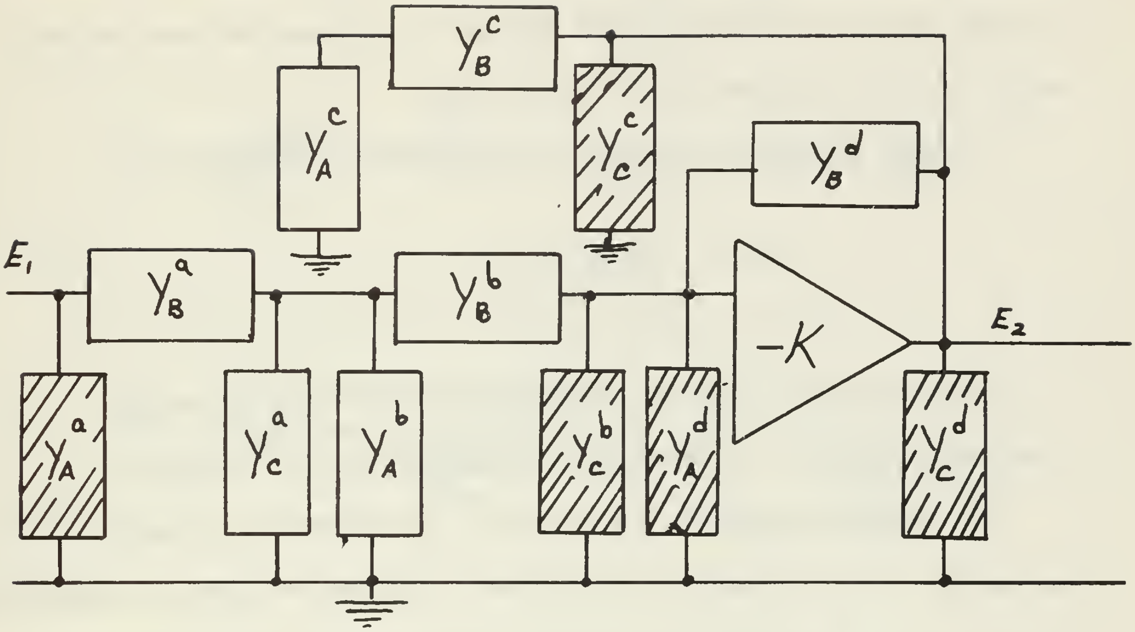
$$Y_4 = Y_B^c$$

$$Y_5 = Y_B^d$$

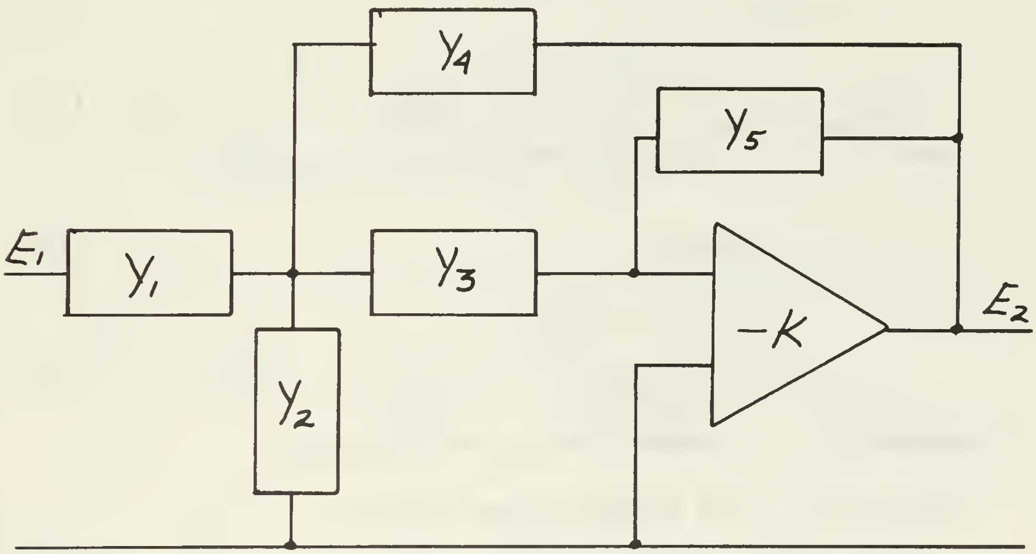
Thus the resultant transfer function becomes

$$\frac{E_2}{E_1} = \frac{-\,Y_1\,Y_3}{Y_5(Y_1 + Y_2 + Y_3 + Y_4) + Y_3\,Y_4}$$

3. <u>Synthesis</u>. Since Y_2 in the simpler circuit appears only in the factor of sums it contributes nothing to the transfer function which cannot be derived from the other elements. If the blocks indicated are assumed to be simple components (either resistors or capacitors) two useful configurations can be obtained. Figure 12



(a) GENERAL OPERATIONAL AMPLIFIER CIRCUIT



(b) SIMPLIFIED EQUIVALENT

FIGURE 11

shows these circuits and their transfer functions. These two are similar tuned-circuit type functions whose major practical difference is that in one case the control element is a capacitor and in the other the control element is a resistor.

In both cases the transfer function is of the form

$$G(s) = \frac{Axs}{XS^2 + Bs + C}$$

Thus two such circuits in cascade with proper choice of components results in the desired bandpass circuit response.

4. Design Example. Two stages of the circuit of Figure 12 (a) in cascade with a single-pole, high-pass filter provide one realizable configuration. From the transfer functions given the following requirements result:

a. High-Pass Stage (two pole).

$$1 + 3/4\ K = C_2/C_B \qquad (1)$$

$$\omega_o = (R_2 + R_1)/(R_1 R_2 C_B) = 1/RC_B \qquad (2)$$

where R is the parallel combination of R_1 and R_2

$$\omega_o^2 = 1/(R_1 R_2 C_1 C_B) \qquad (3)$$

Dividing (3) by (2)

$$\omega_o = 1/C_1(R_1 + R_2) \qquad (4)$$

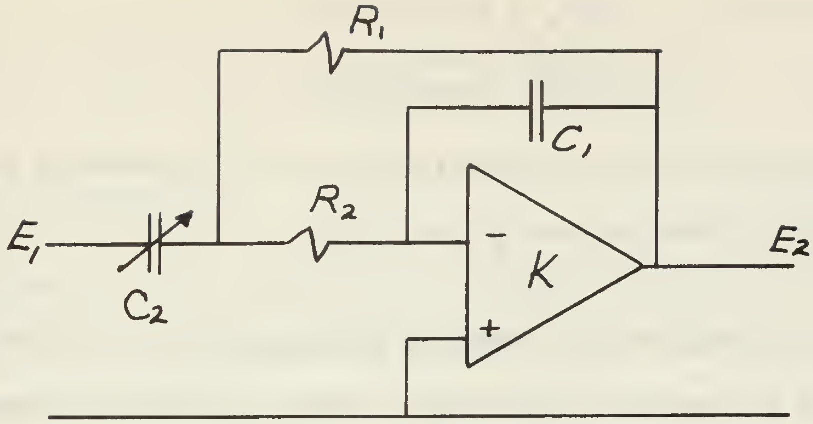
Assume f_o is 50 hertz, then ω_o = 314 rad/sec.

Let $R_1 = R_2$ = 110 kilohms, then from (4)

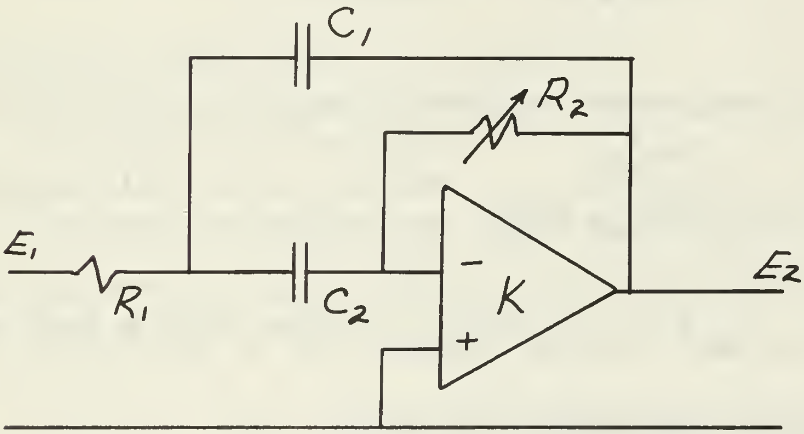
$$C_1 = 1/\omega_o(R_1 + R_2) = 1/314 \times 220 \times 10^3$$

or

$$C_1 = 0.0145 \text{ microfarads}$$



(a) $$\frac{E_2}{E_1} = \frac{-s\left(\frac{C_2}{C_B}\right)\left(\frac{1}{R_2 C_1}\right)}{\left(\frac{C_2}{C_B}\right)s^2 + \frac{(R_1+R_2)}{R_1 R_2 C_B}s + \frac{1}{R_1 R_2 C_1 C_B}}$$



(b) $$\frac{E_2}{E_1} = \frac{-s\left(\frac{R_2}{R_B}\right)\left(\frac{1}{C_1 R_1}\right)}{\left(\frac{R_2}{R_B}\right)s^2 + \frac{(C_1+C_2)}{C_1 C_2 R_B}s + \frac{1}{R_1 R_B C_1 C_2}}$$

FIGURE 12. TWO BANDPASS CIRCUITS

From (2) $C_B = 1/R\omega_o = 1/314 \times 55 \times 10^3$

or $C_B = 0.0578$ microfarads

For (1 + 3/4 K) to range from 1/2 to 4, C_2 should be controlled from $C_B/2 = 0.0289$ μF to $4C_B = 0.231$ μF.

b. Low-Pass Stage. The same technique used for the high-pass stage can be used here except that, since no variable frequency response is desired, it becomes somewhat simplified.

The transfer function is the following

$$\frac{E_2}{E_1} = \frac{-\,s\,\dfrac{1}{R_2C_1}}{s^2 + \dfrac{R_1 + R_2}{R_1R_2C_2}\,s + \dfrac{1}{R_1R_2C_1C_2}}$$

The second-order Butterworth low-pass filter function is given by $G(s) = \dfrac{A\,\omega_{o2}^2}{s^2 + \omega_{o2}s + \omega_{o2}^2}$ where ω_{o2} is the radian corner frequency.

$$\omega_{o2} = (R_1 + R_2)/(R_1R_2C_2) = 1/RC_2 \text{ where R again is} \qquad (1)$$

the parallel combination of R_1 and R_2.

$$\omega_{o2}^2 = 1/(R_1R_2C_1C_2) \qquad (2)$$

Dividing (2) by (1)

$$\omega_{o2} = 1/C_1(R_1 + R_2)$$

If f_{o2} is 400 hertz then $\omega_{o2} = 800\pi$ rad/sec and if $R_1 = R_2 = 100$ kilohms then

$$C_1 = 1/(R_1 + R_2)\ \omega_{o2} = 1/(200\text{x}10^3)(800\pi)$$

or

$$C_1 = 0.00198\ \mu F$$

From equation (1) $\quad C_2 = 1/R\ \omega_{o2} = 1/(50\text{x}10^3)(800\pi)$

or $\quad C_2 = 4C_1 = 0.00798\ \mu F$

For the overall response of these two stages in cascade the numerators of the transfer functions are related to the standard form in the following manner:

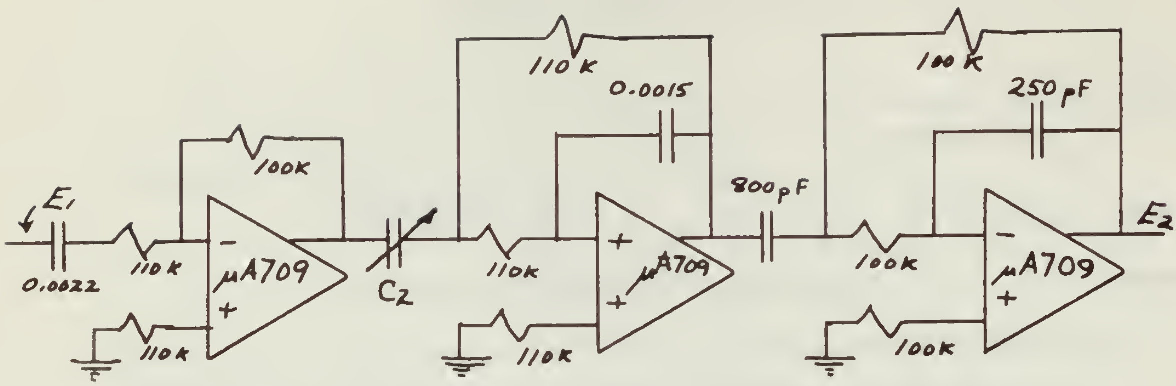
$$A\omega_{o2}^2 = \frac{1}{R_2C_1}\ \text{High Pass} \approx \frac{1}{R_2C_1}\ \text{Low Pass}$$

Solving for the insertion loss, A, results in

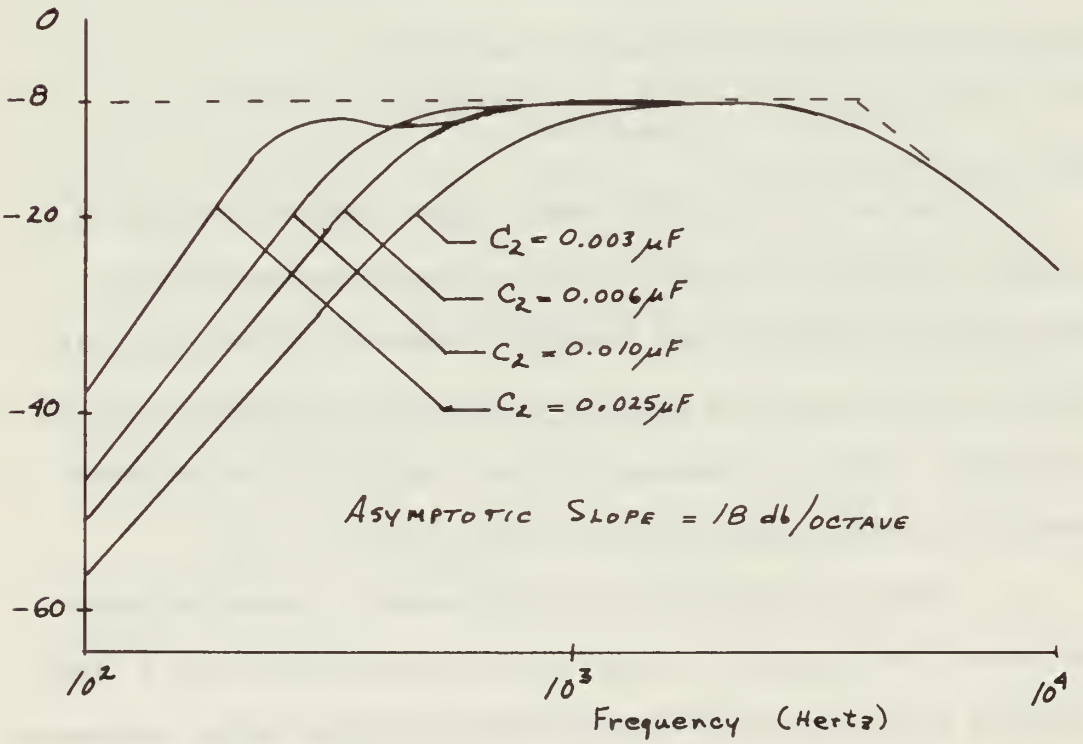
$$A = 0.503 = -\ 5.8 \text{ decibels}$$

c. Single-Pole High-Pass Stage. This stage was made up of a simple high-pass RC circuit using the input impedance of the operational amplifier as the resistance element. This method was chosen so that this stage could be considered an integral block in the cascade. Thus, if necessary, overall gain could be obtained merely by providing gain in this stage.

d. Circuit Configuration and Performance. Because of signal-generator limitations, frequency values were scaled up by a factor of ten by using capacitors of one-tenth the design values previously computed. Figure 13 shows the actual implementation of this design and the resulting experimental frequency response. In this case C_2 was a decade capacitor. The experimental insertion loss of eight



(a) Circuit Configuration (Less Compensating Circuits)



(b) Experimental Frequency Response

FIGURE 13 A VARIABLE FILTER

decibels is not far from the theoretical value of 6.8 db resulting from the loss of the single-pole stage and the 5.8 db derived above. All of the above theoretical results assume ideal operational amplifiers. In practice the feedback resistor values used (100 kilohms) are large enough to affect the ideal operational amplifier assumption and introduce some error.

D. SALLEN-KEY CIRCUITS

Another method of transfer-function synthesis uses active circuits having the following properties:

1. Finite gain, K. Often K equals unity.

2. Very high input impedance--assumed infinite.

3. Output impedance assumed to be zero.

Circuits using this type of active element are often called Sallen-Key circuits because an extensive catalog of these was made by Sallen and Key in 1955.[12]

1. Analysis of General Form. Consider the general feedback network shown in Figure 14.

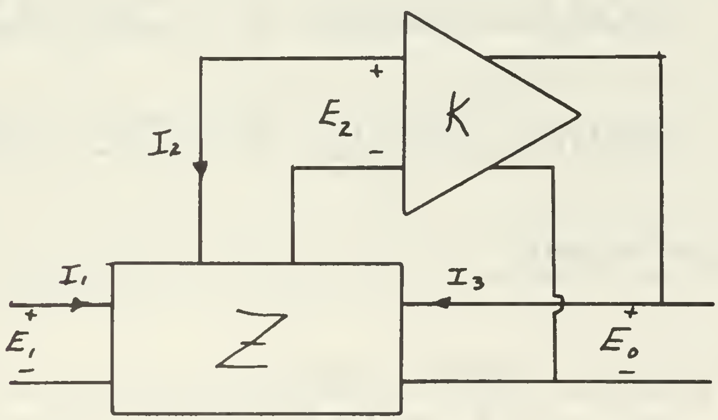


Figure 14. Generalized Sallen-Key Circuit.

The assumptions made above on the character of the active circuit place the following restrictions on the quantities given.

$$I_2 = 0$$

$$E_o = KE_2$$

If Z is considered to be a three-port network the standard circuit equations for the system are

$$E_1 = z_{11} I_1 + z_{13} I_3 \qquad (1)$$

$$E_2 = z_{21} I_1 + z_{23} I_3 \qquad (2)$$

$$E_3 = z_{31} I_1 + z_{33} I_3 \qquad (3)$$

Solving equations (1) and (2):

$$I_1 = \frac{E_1 z_{23} - E_2 z_{13}}{\Delta z}$$

and $I_3 = \dfrac{E_2 z_{11} - E_1 z_{21}}{\Delta z}$ where $\Delta z = z_{11} z_{23} - z_{13} z_{21}$

Substituting these values into equation (3):

$$E_o = \frac{E_1(z_{23} z_{31} - z_{33} z_{21}) + E_2 (z_{11} z_{33} - z_{13} z_{31})}{\Delta z}$$

$$= \frac{E_1(z_{23} z_{31} - z_{33} z_{21}) + \frac{E_o}{k} (z_{11} z_{33} - z_{13} z_{31})}{\Delta z}$$

Solving for the gain

$$\frac{E_o}{E_1} = \frac{z_{23} z_{31} - z_{33} z_{21}}{(z_{11} z_{23} - z_{13} z_{21}) + \frac{1}{k} (z_{13} z_{31} - z_{11} z_{23})}$$

2. <u>Specific Useful Sallen-Key Circuits</u>. The analysis above applied to the circuits of Figure 15 yields the given responses which can be used in realizing variable-frequency filters on the type desired.

The high-pass transfer function is equivalent to

$$G(s) = \frac{(R_2/R_B)s^2}{(R_2/R_B)s^2 + \frac{C_1 + C_2}{C_1C_2R_B} s + \frac{1}{C_1C_2R_1R_B}}$$

where R_B is the value of R_2 for a Butterworth response. As before the following equations must hold:

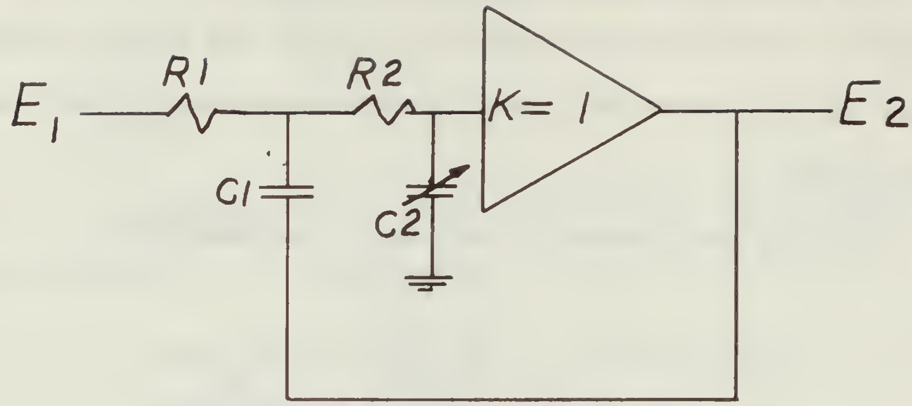
$$1 + 3/4\ K = R_2/R_B$$

$$\omega_o = (C_1 + C_2)/C_1C_2R_B$$

$$\omega_o^2 = 1/C_1C_2R_1R_B$$

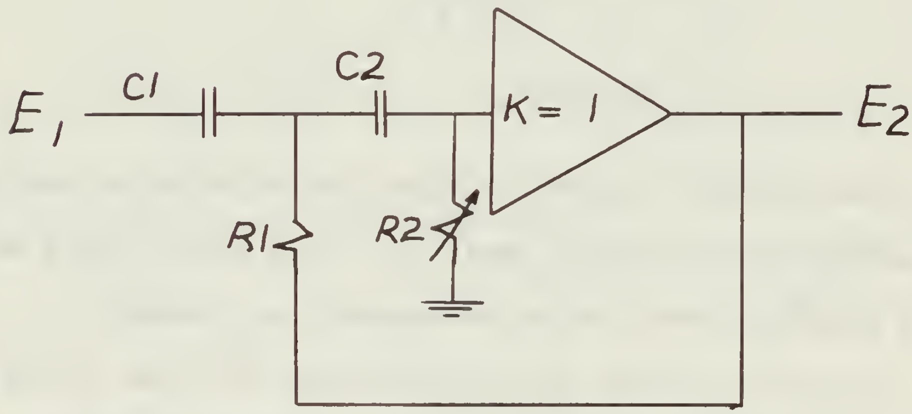
The low-pass transfer functions can similarly be used to make a variable break frequency where C_2 of this circuit would be used as a variable element in the same manner as R_2 above.

These two circuits have one advantage over most of the previously derived circuits because the control element is tied to ground alleviating some problems caused by lack of isolation of a voltage-variable element between the controlled parameter and the controlling voltage.



$$G(S) = \frac{1/C_2}{1/C_2 + S\ (R_1+R_2) + S^2 C_1 R_1 R_2}$$

(a) LOW PASS



$$G(s) = \frac{R_2 s^2}{R_2 s^2 + s\,\frac{C_1+C_2}{C_1 C_2} + \frac{1}{C_1 C_2 R_1}}$$

(b) HIGH PASS

FIGURE 15 SALLEN – KEY CIRCUITS

IV. VOLTAGE-VARIABLE ELEMENTS

A. INTRODUCTION

It is apparent from the above analysis that the break frequency of a three-pole filter can be adjusted over a sufficiently wide range by changing the value of a single component while still preserving the general overall shape of the response. The remaining problem is one of obtaining a voltage-variable element which will be useable in one of the circuits previously described.

In order to determine what kind of element might be successfully employed some requirements should be considered.

1. It may be either a resistor or a capacitor. However, most circuits which resulted in a variable capacitor as the control element for the high-pass configuration required isolation of the capacitor from ground. Of those circuits previously derived only one employed a grounded capacitor: the integrator-summer type using positive-gain integrators.

2. For adequate range of the break frequency the element requires an eight-to-one ratio of variability.

3. The element needs to be sufficiently small so that the overall size restriction of the filter is not exceeded.

4. Its parameter variation as a function of the control voltage needs to be sufficiently predictable so that reasonable engineering design can be made with confidence that all circuits will behave as predicted, with only minor final adjustments required.

5. Within the range from a few millivolts to a few volts the maximum signal voltage can be arbitrarily placed at any value by proper choice of amplifiers. Thus the element must have the capability of operating somewhere in this range. It must not require extremely large signal levels and, conversely, it must accept sufficiently large signals so that normal processing can be accomplished without the noise level becoming unduly large with respect to the signal.

Two variable capacitor elements which, at first glance, might seem to be useable are the voltage-variable capacitance diode and reactance circuits. Reactance circuits use feedback technique to multiply the capacitance in proportion to the gain of the active circuit employed. If this gain can be varied then the effective capacitance of the circuit will also vary.

The voltage-variable capacitance diodes, in general, do not have the required tuning ratio and they are in the picofarad range. At doppler frequencies the capacitance required is of the order of a microfarad for reasonable circuit impedance. For higher-frequency applications and for smaller frequency variations these diodes might be suitable. The biasing problem for non-grounded elements remains, however.

Reactance circuits and capacitance multiplier circuits generally require isolation of the signal circuit from the power supply since otherwise the Q of the circuit is ruined by the shunting effect of the power supply. This is generally done with chokes which are not practicable at the low audio frequencies.[13]

Because of the apparent difficulty in obtaining variable capacitance elements and because of the limitation listed in the first requirement concerning capacitors, the goal of finding an adequate variable element was primarily directed towards resistance elements.

B. FIELD-EFFECT TRANSISTOR

If the characteristics of field-effect transistors for small drain-to-source voltage levels are investigated it is found that in this region they exhibit reasonable linearity and the drain-to-source resistance, R_d, is controlled by the gate-to-source voltage, V_{GS}. Figure 16 shows characteristics measured for an RCA 40468 insulated-gate field-effect transistor chosen at random.

The resistances of three different types of field-effect transistors are shown in Figure 17. The variation of samples within a general type is apparent from these curves which are of samples chosen at random and then measured.

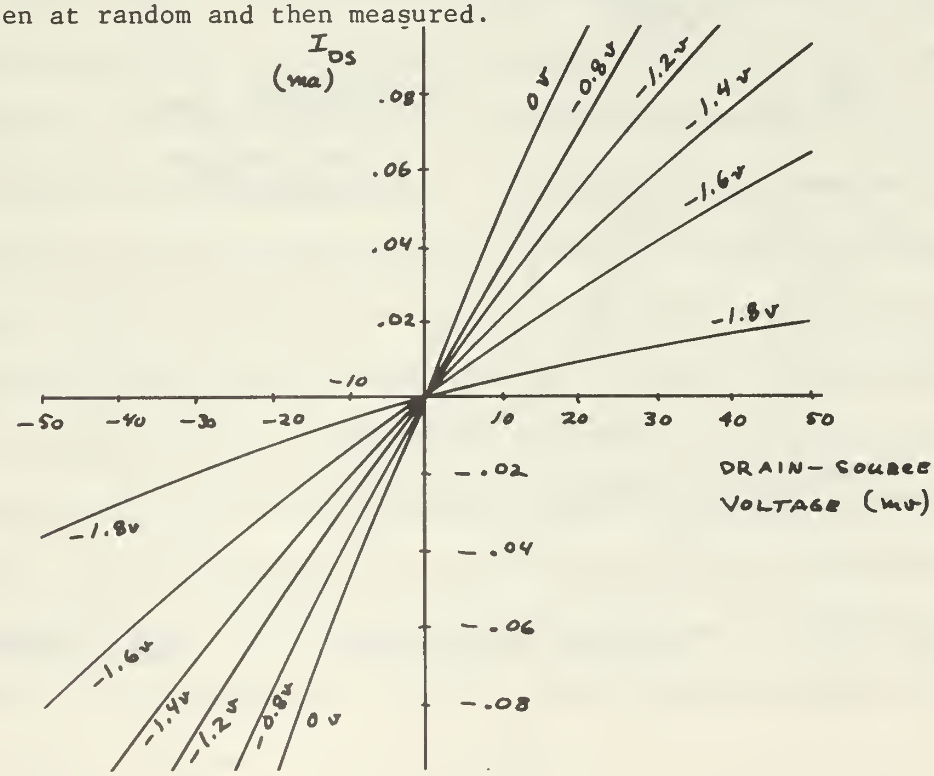


Figure 16. Small-signal characteristics of a particular RCA 40468 MOSFET.

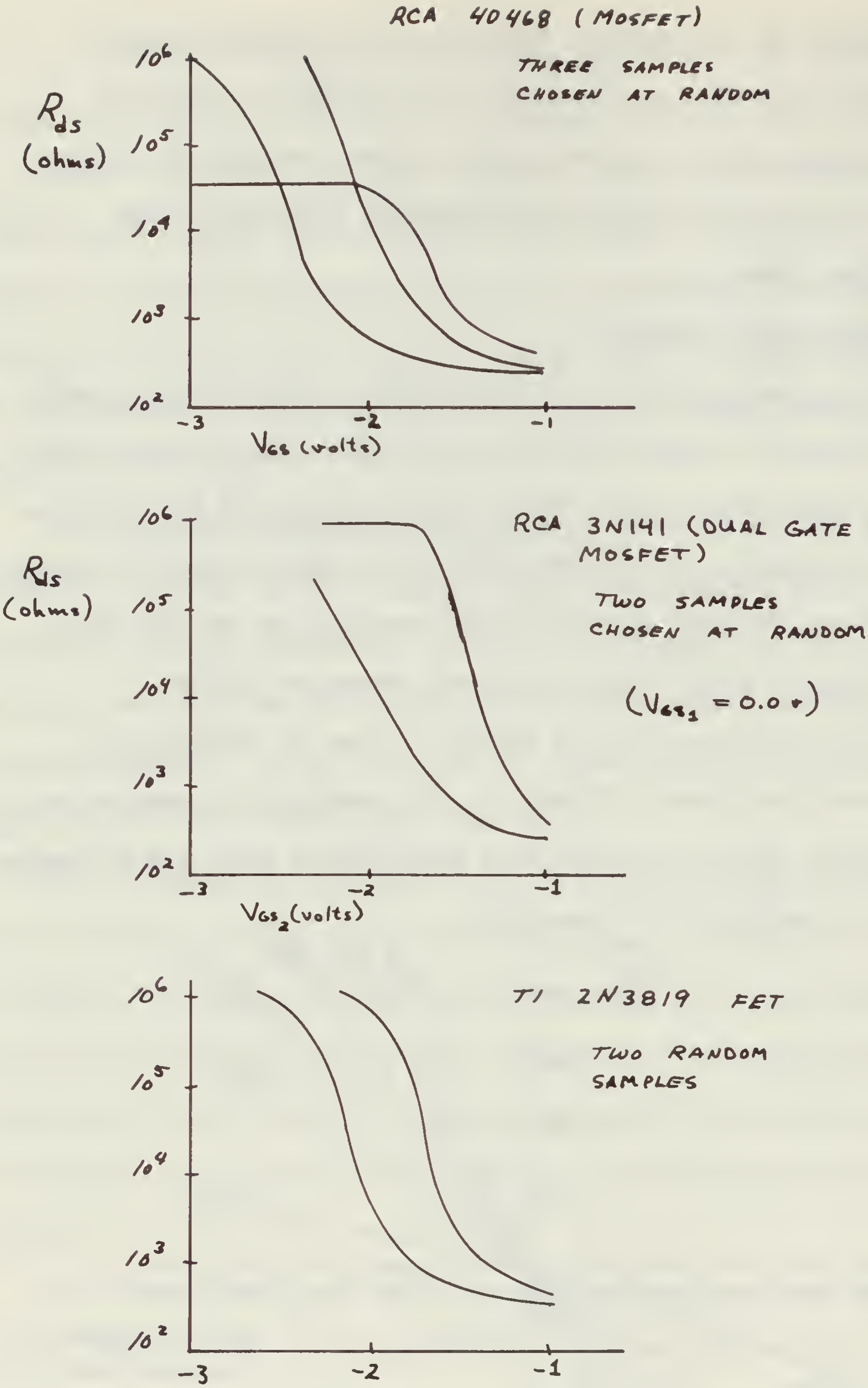
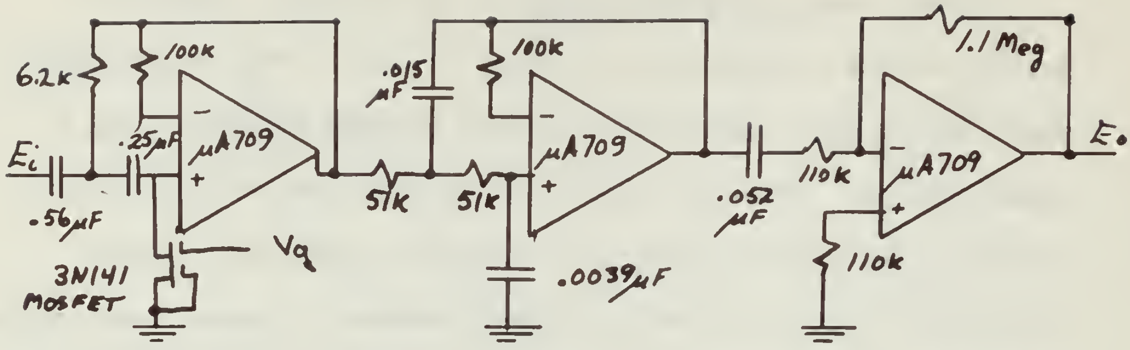


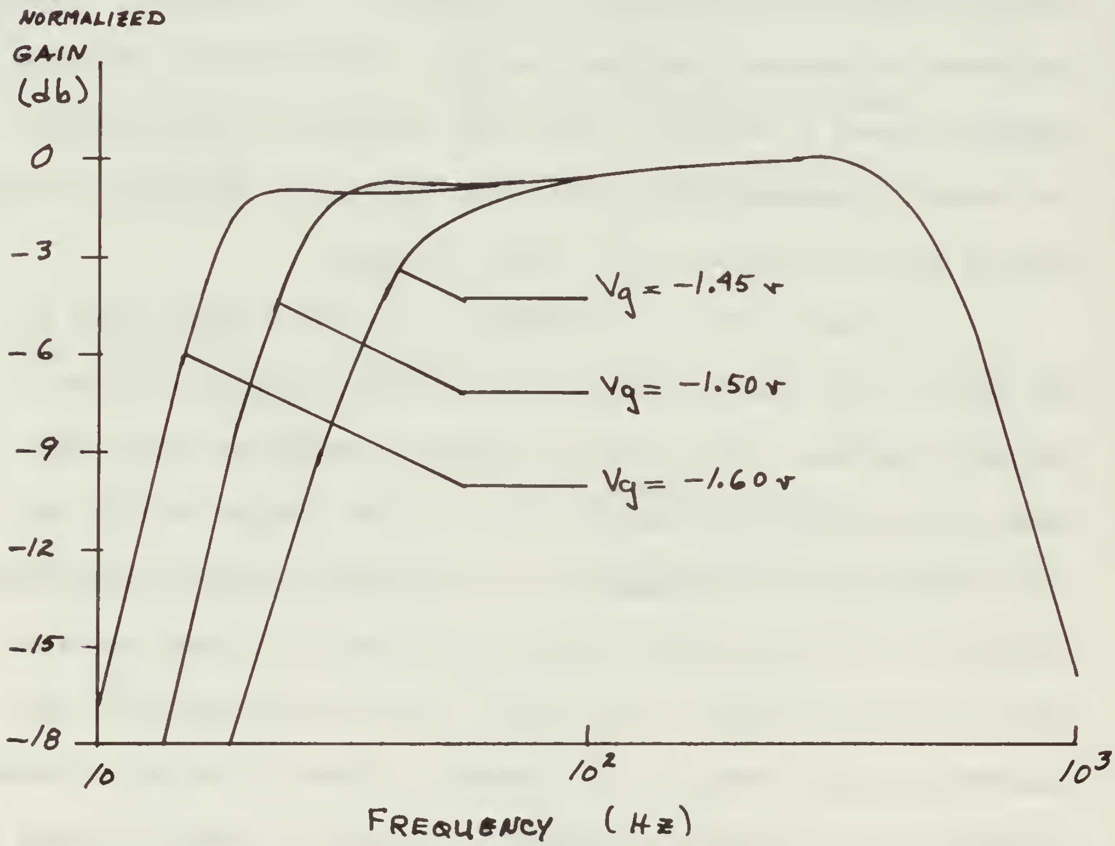
FIGURE 17 CHANNEL RESISTANCE OF SOME FET's

Some attempts were made to use a field-effect transistor in the circuit configuration of Figure 12 (b). With an isolated power supply there would be no serious problem in controlling a single filter. However, isolated power supplies are impractical. Moreover, for a bank of filters requiring control by a single voltage the voltage source must have a common or ground termination. In this circuit neither the source nor the drain of the field-effect transistor is at ground potential. Both have signal voltage. Thus when applying control voltage to the gate, D.C. voltage with respect to ground is added to the signal voltage. The resultant effective feedback causes a nonlinear action and distortion of the waveform. No biasing arrangement was found which would allow effective control without gross distortion of the signal voltage.

The Sallen-Key circuit of Figure 15 (b) was designed with an RCA 3N141, dual, insulated-gate, field-effect transistor as the variable resistor. The method of computing component values was identical to that illustrated in the previous chapter with R_B of the MOSFET chosen at 30 kilohms. The required unity-gain amplifier consists of the operational amplifier and the 100 kilohm resistor which clamps the input to the output. The input signal was a one-hundred-millivolt peak-to-peak sinusoid. Figure 18 (a) is a diagram of the circuit used and the frequency response is shown in Figure 18 (b). Note that the last stage is a single-pole high-pass stage with a gain of ten. This illustrates how the filter itself can incorporate a stage with gain. By proper choice of components the gain can be made considerably greater than ten. The frequency response curve has been normalized to the midband gain.



(a) CIRCUIT CONFIGURATION



(b) FREQUENCY RESPONSE

FIGURE 18 A VOLTAGE VARIABLE FILTER

Many problems exist in the employment of field-effect transistors as voltage-variable elements. Even within a given type the resistance-versus-control-voltage characteristic is subject to wide variation. One cause of this appears to be the lack of manufacturer concern for this particular parameter of field-effect transistors. Very few data sheets include this in the device description. One data sheet which does give curve values for this is that describing the RCA 3N138, but no samples were available for testing for consistency.

Thus though field-effect devices can be used they fail to satisfy the fourth requirement for variable elements; the available types were not sufficiently predictable to provide design data. Other circuit parameters could be determined only after the individual device was measured. This is an impractical method of design especially where a large number of circuits are necessary.

Another property of the field-effect transistors which affects its employment as the variable resistance element is the extreme sensitivity of the device to the control voltage in its useful range of control. From the curves in Figure 17 (b) it can be seen that in the useful range the resistance doubles for each four-tenths volt of gate voltage applied.

The temperature dependence of the field-effect transistors was not investigated experimentally. It has been shown that the effect can be significant but that the magnitude and direction of the effect is a function of the manufacturing technique used.[14]

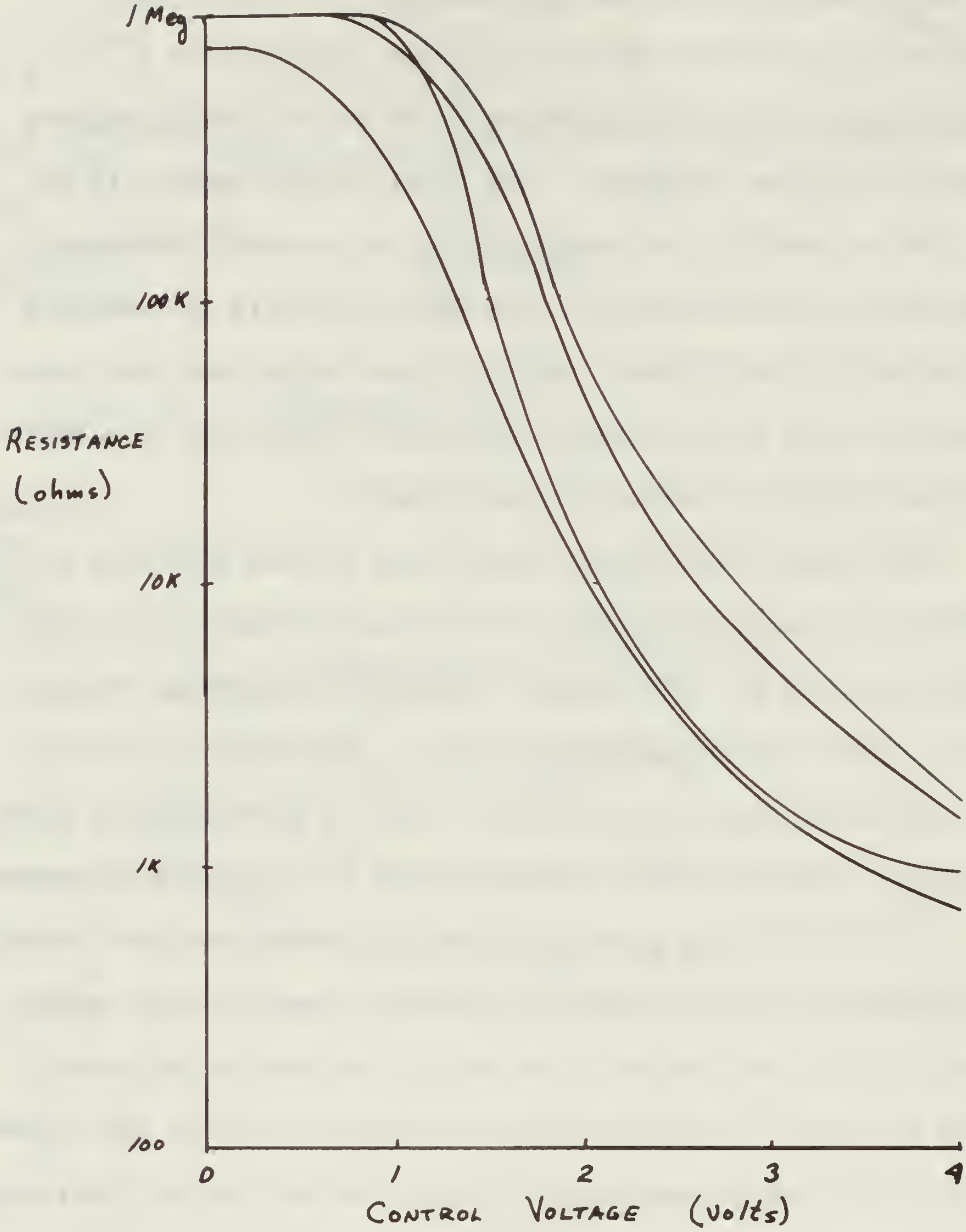


FIGURE 19 RAYSISTOR CHARACTERISTICS

C. RAYSISTORS

Raysistors are commercial devices made by Raytheon which consist of a light source and a photosensitive resistor enclosed within a light-tight case. Varying the voltage across the lamp causes variable light intensity which varies the resistance of the photo-resistor. At full lamp brightness the resistance is low; at low voltage or low brillance the resistance is high.

These devices are available in various packages including TO-5 cans. Four terminals are provided: two for the lamp current and two for the resistor terminals. Since the controlling medium is light there is complete electrical isolation maintained between the control source and the controlled device. Thus such an element can be used in any of the variable resistor configurations discussed above.

Four samples of Raysistor model CK 1122 were obtained. The catalog description indicated the packaging as TO-5; the devices actually were contained in crystal-type cans whose dimensions were 1.8 cm by 0.8 cm by 1.9 cm (excluding terminal lengths). This is still small enough to fulfill the requirements imposed on physical size.

The resistance characteristics of these samples as a function of the lamp control voltage are shown in Figure 19. Among samples the shape of the resistance curves appear somewhat more consistent than the behavior of the field-effect transistor samples within a given type. At a given control voltage there was considerable difference in resistance between devices. For example, at two volts of control voltage the lowest resistance recorded was twelve

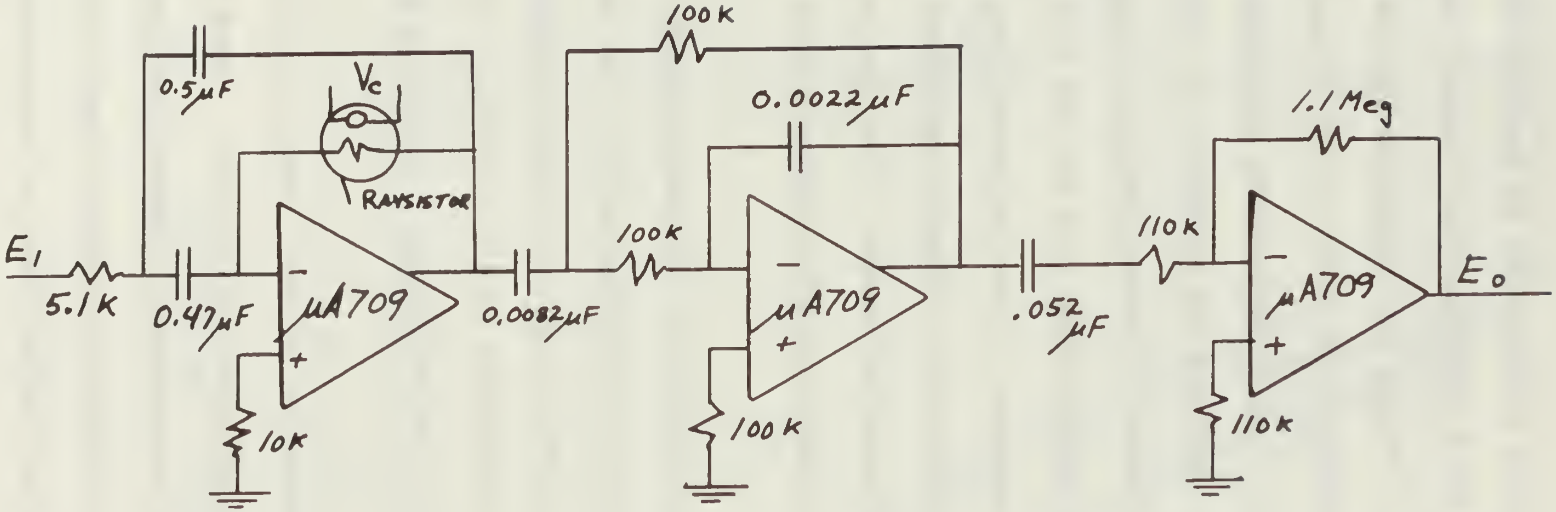


FIGURE 20 A VOLTAGE VARIABLE FILTER USING A RAYSISTOR AS THE CONTROL ELEMENT

and one half kilohms while the highest was ninety kilohms. This would require voltage offset between filters and careful calibration.

Even though they could be used in a Sallen-Key circuit similar to the one in which FETs were used, the configuration of Figure 20 was chosen to demonstrate the isolation of the device. The circuit was designed for a control range from ten kilohms to one hundred kilohms, providing a ten to one ratio. Thus R_B was twenty kilohms. The control voltages corresponding to these resistance values were found and each device was then controlled over its own range. In practice this would require separate voltage-divider circuits for each filter, to convert the overall control applied to a bank of such filters. Figure 21 illustrates the tracking achieved from all four circuits at high and low values of resistance levels.

Some of the same objections that were raised against the field-effect transistor are also present with the Raysistor; that is, the ability to predict the response is quite limited and the sensitivity to the control voltage is high. This sensitivity (the resistance doubles for each 0.35 volt change of control voltage in the center of the control range) is very similar to that of the field-effect transistor.

D. STEP TUNING RESPONSE

It is possible by using voltage switching devices such as transistors, field-effect transistors, and similar devices to cause the low break frequency to change in discrete steps.

One simple method of accomplishing this is shown in Figure 22, where four steps were chosen for convenience, but any reasonable number could be incorporated into the filter circuit. Further, in

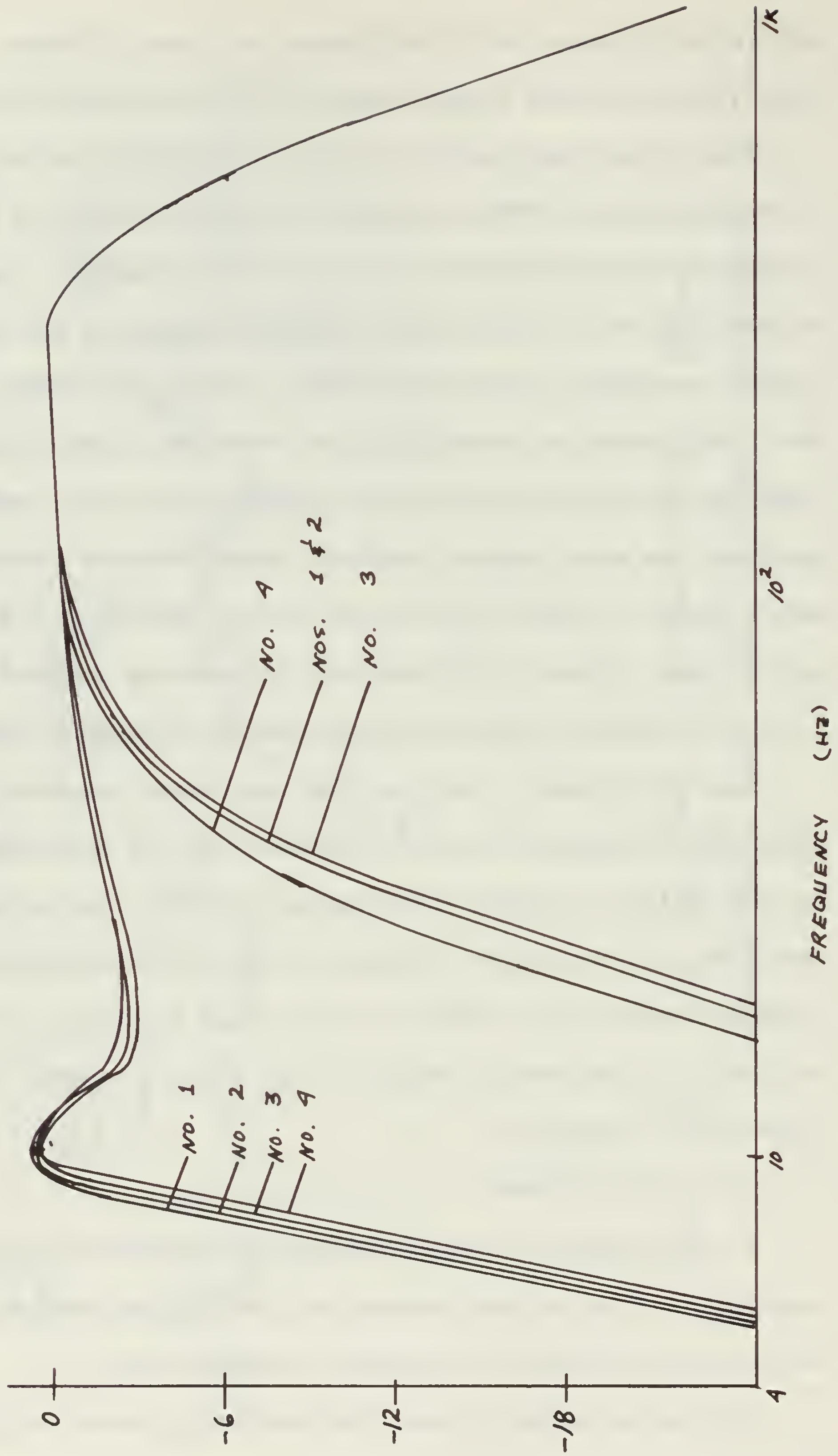

RAYSISTOR TRACKING
NO. 1
NO. 2
NO. 3
NO. 4
NO. 4
NOS. 1 & 2
NO. 3
0
-6
-12
-18
4
10
10^2
1K
FREQUENCY (Hz)

FIGURE 21

a bank of similar filters all the filters can be similarly controlled by connecting like gates to the same control voltage. The field-effect transistors with zero gate voltage then shunt the resistors which they are across while those with large negative voltage (in the case of an n-channel depletion-mode device) present about a megohm resistance to the signal,allowing the resistors to have control. Thus the unpredictable midrange of the device is not used.

One attribute of this particular version of switching circuit is that very little power is required to control the switches since isolation of the gate for an insulated-gate transistor is good. The Raysistor could also be used in this mode (without the low-power advantage cited) but the polarity for control would be reversed: zero control voltage causes an effective open circuit and eight volts causes the Raysistor to approach short circuit.

The response curves obtained experimentally with the circuit of Figure 22 are shown in Figure 23. This technique fulfills all of the requirements listed above. It allows only discrete changes of the response shape but for application to range-gated MTI this is not too serious a restriction. If the radar operator has the option of choosing four or five different settings of the filter he should be able to effect a satisfactory compromise between maximum target reception and minimum clutter.

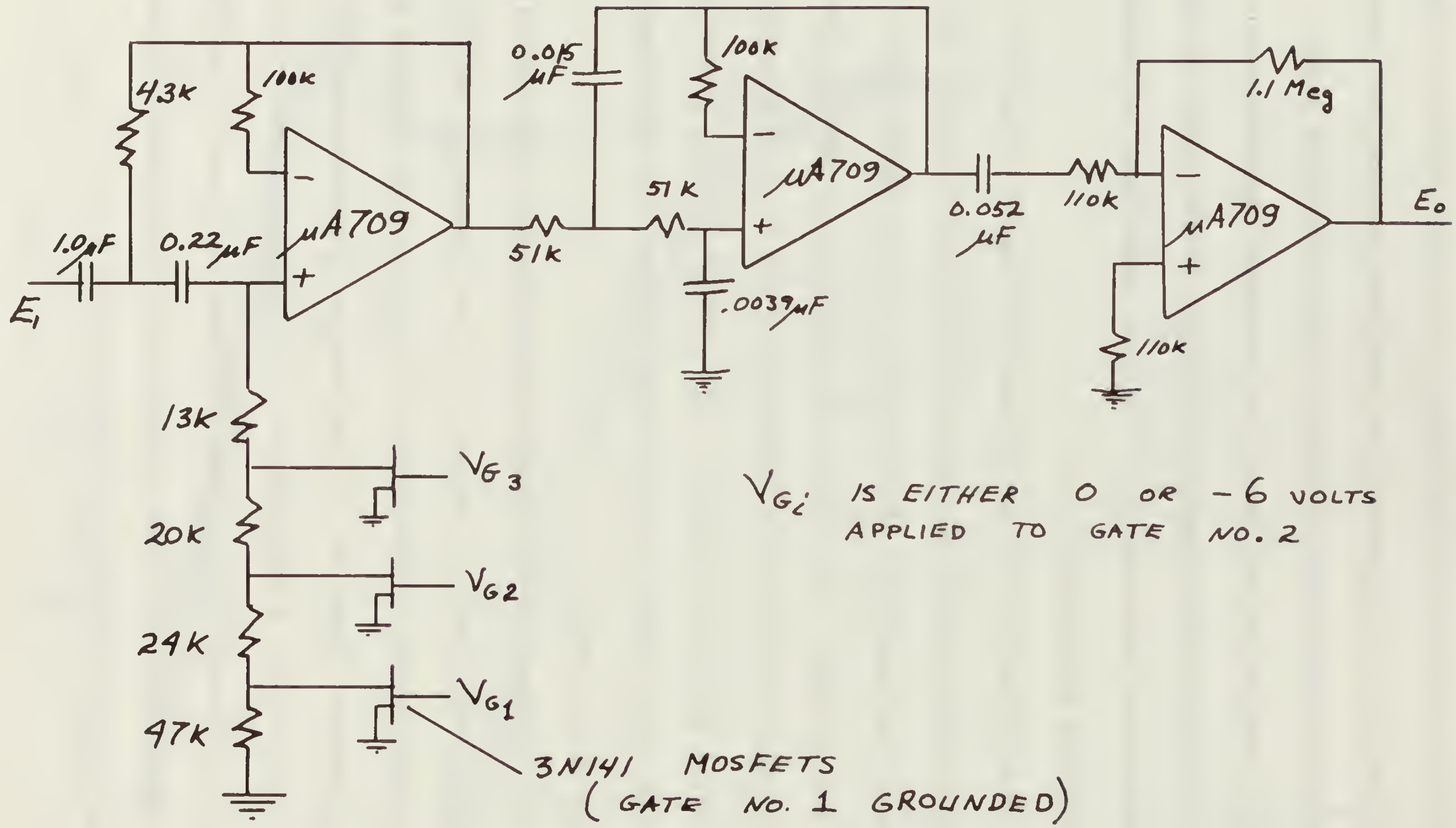


FIGURE 22 A STEP VOLTAGE VARIABLE FILTER

FREQUENCY RESPONSE OF STEP VOLTAGE VARIABLE FILTER

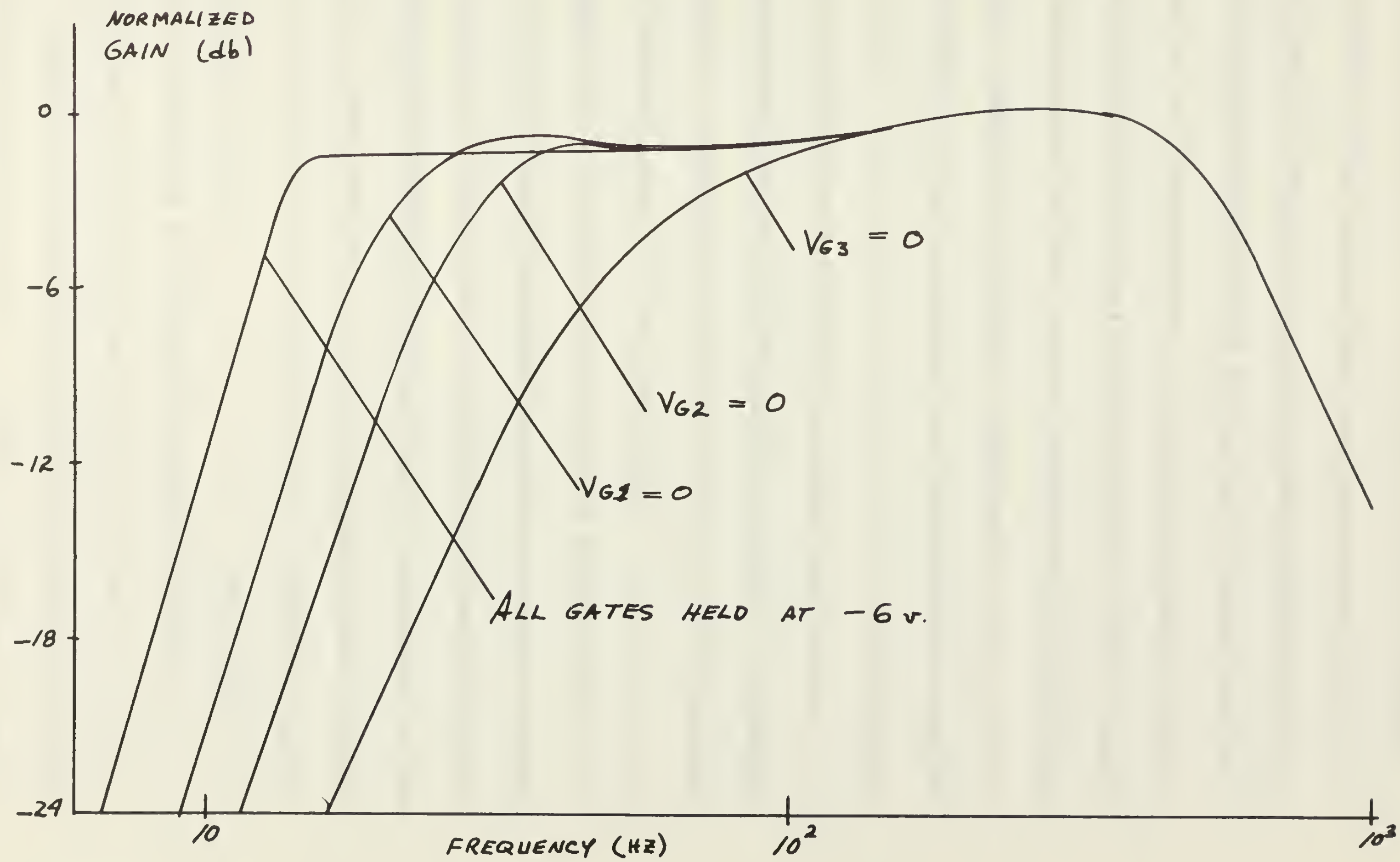


FIGURE 23

V. RESULTS AND CONCLUSIONS

A. SUMMARY

It was shown above that nearly three octaves of variation in break frequency can be obtained by use of the Thiele method of filter design. This method provides the capability of adjusting the frequency response of a three-pole filter by varying a single element. Though some configurations were obtained with capacitors as the variable element most of these required isolation from ground which is difficult to attain. Further, voltage-variable capacitors having the required values are not available.

Two variable-resistance elements were tested and used in appropriate circuit configurations. Both the field-effect transistor and the Raysistor lacked uniformity of response within a given type. In addition they both showed extreme sensitivity to control voltage in the range of control. Other types of these elements might exhibit more uniformity especially if the manufacturing process was established with variable-resistance control as the objective. Present goals for field-effect devices are pointed towards application in the nonlinear region of the device characteristics.

B. CONCLUSIONS

If discrete voltage variability is acceptable it is not difficult to devise circuits using the filters described above where the frequency response is changed by switching in or out discrete resistance elements. One circuit was developed which successfully provided such control.

Nearly all the elements used in the circuits above are capable of being fabricated within integrated circuitry. The Raysistor and the large capacitors are probable exceptions to this. The circuit using discrete tuning could probably be entirely integrated except for the two large capacitors in the high-pass circuit. Monolithic integrated circuitry utilizing compatable thin-film technique for the resistors appears feasible.[15] Thus the requirement for small physical size would be fulfilled to an extent not possible with discrete components.

Some extensions of the design method described above include making band-pass filters that are variable on both high and low-frequency rolloff. This could provide a selective IF filter either manually or automatically varied by signal strength. In communication circuits, audio filters could be constructed that are adaptable to noise environment.

One method of obtaining variable response would appear to be by use of a Sallen-Key circuit where the gain of the active device would be the variable parameter. No circuit was found that would give this response. However, no exhaustive study was made of this technique and such a circuit may exist. If one could be found, an amplifier with automatic-gain-control provision could provide the break-frequency variation.

BIBLIOGRAPHY

[1]C. D. Shupe, R. T. Paca and T. T. Monaghan, "Integrated Circuits for Combat Area Equipment -- Report No. 3", Motorola, Inc. Military Electronics Division, Scottsdale, Arizona, February 1966.

[2]W. L. Cappadona, "Microminiaturized MTI Group AN/GPA (XW-2)", Tech. Rpt. No. RADC-TR-65-442, Rome Air Development Center, Griffiss Air Force Base, New York, March 1966.

[3]R. J. Berg and P. N. Marshall, "Spotting Targets on the Wing", Electronics, vol. 40, pp. 58-65, 25 December 1967.

[4]M. I. Skolnik, Introduction to Radar Systems, New York: McGraw-Hill, 1962, p. 146.

[5]A. N. Thiele, "Filters with Variable Cut-off Frequencies", Proc. I.R.E.E. Australia, vol. 26, pp. 284-300, September 1965.

[6]M. E. Van Valkenburg, Introduction to Modern Network Synthesis, New York, John Wiley & Sons, Inc., 1962. p. 376.

[7]W. J. Kerwin, L. P. Huelsman, and R. W. Newcomb, "State-Variable Synthesis for Insensitive Integrated Circuit Transfer Functions", IEEE J. Solid State Circuits, vol. SC-2, pp. 87-92, September 1967.

[8]M. J. Somerville and G. H. Tomlinson, "Filter Synthesis Using Active RC Networks", J. Electronics and Control, vol. 12, pp. 401-420, May 1962.

[9]H. C. Pande and R. S. Shukla, "Synthesis of Transfer Functions Using an Operational Amplifier", Proc. I.E.E., vol. 112, pp. 2208-2212, December 1965.

[10]R. C. Foss and B. J. Green, "Q Factor, Q Stability and Gain in Active Filters", Electronic Letters, vol. 2, pp. 99-100, March 1966.

[11]A. Meltzer, "Activating the Passive RC Network", Electronics, vol. 41, pp. 114-118, May 27, 1968.

[12]R. P. Sallen and E. L. Key, "A Practical Method of Designing RC Active Filters", I.R.E. Trans. Circuit Theory, vol. CT-2, pp. 74-85, March 1955.

[13]J. D. Ryder, Electronic Fundamentals and Applications, Englewood Cliffs, N. J.: Prentice-Hall, Inc., 1964.

[14] J. T. Wallmark and H. Johnson, Field-Effect Transistors, Englewood Cliffs, N. J.: Prentice-Hall, Inc., 1966, pp. 286-289.

[15] R. M. Warner, Jr., Integrated Circuits, New York: McGraw Hill, 1965.

[16] Fairchild Semiconductor Data Sheet for μA709C, March 1967.

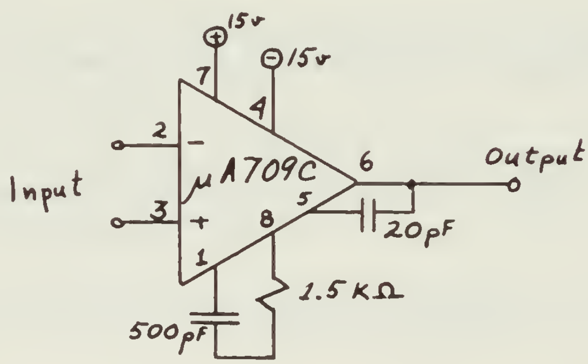
FAIRCHILD OPERATIONAL AMPLIFIER μA709

Electrical Characteristics[16]

	MIN	TYP	
Input Resistance	50	250	kilohms
Output Resistance		150	ohms
Large-Signal Voltage Gain	15,000	45,000	

Frequency Compensation Circuit

This circuit was used for frequency compensation in all the applications discussed within this paper but it was omitted from the circuit diagrams for clarity.



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13. ABSTRACT

Range-gated moving target indicator (MTI) systems have some intrinsic advantages over delay-line canceler MTI systems. A further improvement in performance would result if the doppler cut-off frequency was adjustable by the radar operator. Because of the large number of filters involved in one system this would require a voltage-variable filter element.

A theory of variable filter response developed by Thiele was used as the basis for design using integrated-circuit operational amplifiers as the active elements. Three different types of active filter circuits are discussed. Implementation of voltage-variable filters and experimental frequency-response curves using field-effect transistors, Raysistors, and discrete switched resistors are described and illustrated.

14 KEY WORDS	LINK A		LINK B		LINK C	
	ROLE	WT	ROLE	WT	ROLE	WT
Filters, Active						
Filters, Voltage-variable						
Moving target indicator system, range-gated						