APPLIED ELECTRONICS IN PERSPECTIVE

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FOREWORD

This textbook is addressed to exploration geophysicists whose responsibility is mainly planning or interpretation of field data, and is intended for participants having a background of elementary radio engineering.

In spite of tremendously widespread scope of the associated techniques the coverage of this brochure is, of necessity, quite restricted because of limited time we are given. In addition, somewhat peculiar point of view may be reflected in the selection of materials: a number of important topics are omitted and only cursory derivations of each subject are made here.

Emphasis has been placed on understanding the functions of miscellaneous electronic systems, from the standpoint of those participants who are not majored in the present speciality, rather than on detailed analysis of specific circuit, because it is the feeling of the instructor that our work is primarily concerned with the grasp of underlying principles of this highly developed technology.

It is therefore expected that geophysical instrumentation related to particular method will be reviewed in corresponding course in the curriculum: such as magnetometer in magnetics.

The instructor has consulted numerous references in the preparation of the text. What follows may particularly be suggestive for those who are interested in studying further:

Millman, J., and Taub, H.: Pulse and Digital Circuits

Nodelman, H. M., and Smith, F. W.: Mathematics for Electronics with Applications

Van Valkenburg, M. E.: Network Analysis

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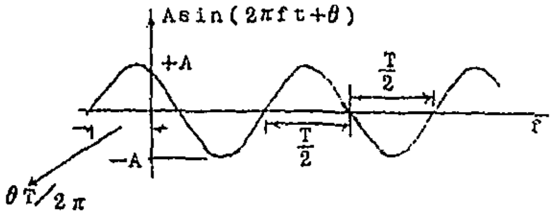
§1. Recalling Fundamental Concepts



amplitude A

Period T

Frequency $f = 1/T$

Phose angle θ

Fig. 1-1 A real sinusoidal time function.

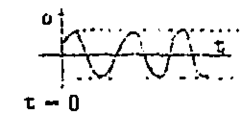
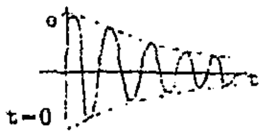
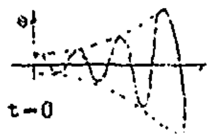
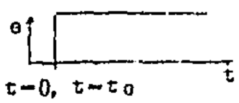
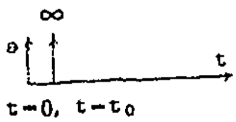
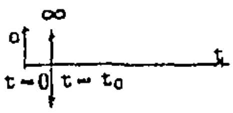
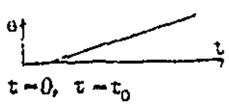
1		sinsoidal wave	$t<0,\ e=0$ $t>0,\ e=E\cos(\omega t+\theta)$
2		exponential wave	$t<0,\quad e=0$ $t>0,\quad \alpha<0$ $e=E\varepsilon^{\alpha t}\cos(\omega t+\theta)$
3		ditto	$t<0,\quad e=0$ $t>0,\quad \alpha>0$ $e=E\varepsilon^{\alpha t}\cos(\omega t+\theta)$
4		unit step function	$t<0,\quad e=0$ $t>0,\quad e=E$ $e=U_{-1}(t-t_0)$
5		unit impulse	$t<0,\quad e=0$ $t>0$ $e=U_0(t-t_0)$
6		unit doublet	$t<0,\quad e=0$ $t>0$ $e=U_1(t-t_0)$
7		linear ramp	$t<0,\quad e=0$ $t>0$ $e=U_{-2}(t-t_0)$

Fig. 1-2 Signal Waveforms

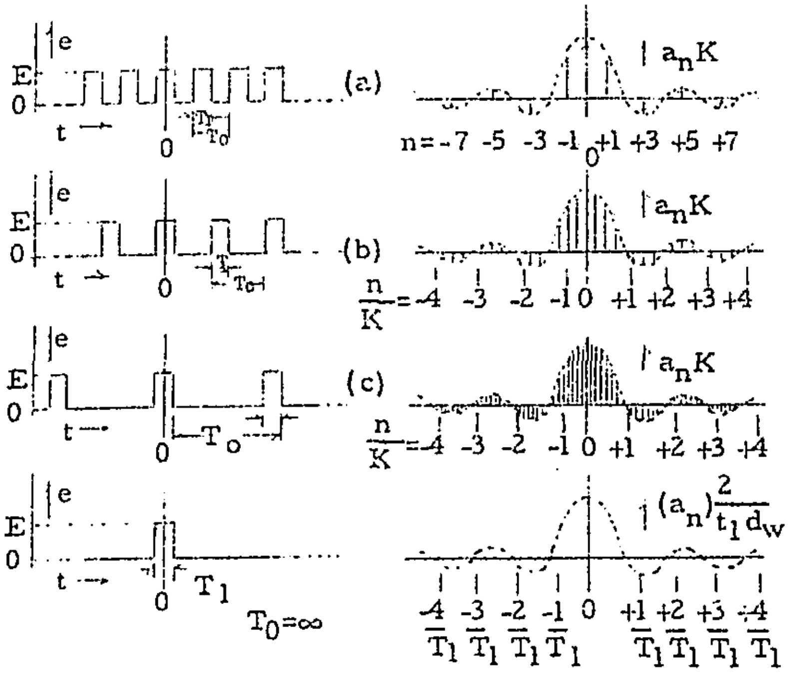


Fig. 1-3 Spectrum of Rectangular Pulse

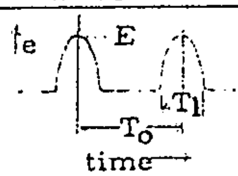
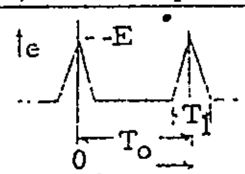
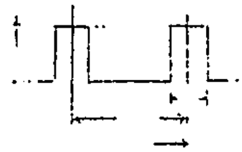
Waveform	Wave Equation	n^{th} Harmonic Amplitude
 (a)Half Cosine	$e = E\cos\frac{\pi}{T_1}t$ $\left(-\frac{T_1}{2}<t<+\frac{T_1}{2}\right)$	$a_nK=\dfrac{4E\cos\frac{n\pi}{K}}{\pi\left[1-\frac{4n^2}{K^2}\right]}$
 (b)Cosine-squared	$e=E\cos\frac{2\pi}{T_1}t$ $\left(-\frac{T_1}{2}<t<+\frac{T_1}{2}\right)$	$a_nK=\dfrac{E.K}{\pi.n}\cdot\dfrac{\sin\frac{n\pi}{K}}{\left[1-\frac{n^2}{K^2}\right]}$
 (c) Triangular	$e=E\left(1+\frac{2t}{T_1}\right)$ for $-\frac{T_1}{2}<t<0$ $e=E\left(1-\frac{2t}{T_1}\right)$ for $0<t<+\frac{T_1}{2}$	$a_nK=\dfrac{2E.K^2}{\pi^2.n^2}$
 (d) Rectangular	$e=E$ $\left(-\frac{T_1}{2}<t<+\frac{T_1}{2}\right)$	$a_nK=\dfrac{2E.K}{\pi.n}\sin\dfrac{n\pi}{K}$

Fig. 1-4 Pulse Waveforms, their Equations and Harmonic Amplitudes.

§ 2. Review of Network Theory

It is possible to classify any network or circuit from different viewpoint.

If a network does not contain any power source, it is called passive network while a network containing power source therein is called active network.

The terminology of distributed circuit is contrasted to lumped circuit, in which every element is idealized to be condensed into point without special extension and exhibit its electrical property only.

In nonlinear circuit, the value of elements is influenced by the current flowing or by the voltage across the element itself, while is not in linear circuit.

We confine ourselves largely to passive, linear, time-invariant and lumped circuits because such simplified networks are the most fundamental frameworks for more complicated analysis.

Among other things, approximation by linear system is prominently useful, because it enables to represent complex signal waveform in the form of a weighted sum of elementary components by superposition property.

Although not every time function possesses a Fourier transform, the very large class of useful functions which is Fourier transformable includes essentially all signal waveforms that can be generated in practice.

An important constraint to linear system is to be noted, the law of causality, that is, an output of linear network in the time domain never precedes the input.

Unit of Transfer Constant

In many problems when two power levels are to be compared, it is found very convenient to compare the relative powers on a logarithmic rather than on a direct scale. The unit of this logarithmic scale is called the bel and decibel (abbreviated db) is 1/10 bel.

If the power being compared are p_1 and p_2, the

$$\text{Decibels} = 10 \log p_2/p_1$$

It should be emphasized that the decibel denotes a power ratio. Consequently the specification of a certain power in decibels is meaningless unless a reference level is implied or is explicitly specified.

If the input and output impedances of a network are equal resistance, the power ratio can be translated into

$$\text{Decibles} = 20 \log E_2 / E_1 = 20 \text{ long } I_2 / I_1$$

Despite the fact that input and output resistances are not equal in general, this expression is adopted as a convenient definition in evaluating gain or loss of energy transmission.

The practical value of the decibel arises from its logarithmic nature. This permits the enormous range of power involved in engineering work to be expressed in terms of decibels without running into inconveniently large number.

The logarithmic character also makes it possible to express the ratio of input to output powers of a complicated circuit as the sum of the decibel equivalent of the ratios of the different parts of the circuit that are in cascade.

Finally, it is worth noting that in acoustics the degree of human perception is proportional to logarithm of strength of stimulus and the decibel is again effective in this regard.

2-1. Network Equations

Kirchhoff's Law

Most network equations are formulated from two simple laws given by Kirchhoff. The first law relates to the sum of the instantaneous voltages of the elements in a loop. It states that in any loop the sum of the voltage drops must equal the sum of the voltage rises.

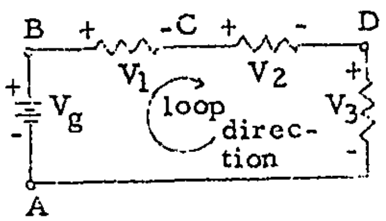


Fig. 2-1

The second law relates to the sum of instantaneous currents at a node. It states that the sum of currents flowing into the node equals the sum of currents flowing out.

Example

Consider the series circuit shown in Fig. 2-1. We see that there are voltage drops across the three passive elements and a voltage rise due to the battery.

According to Kirchhoff's voltage law,

$$V_1 + V_2 + V_3 = V_g$$

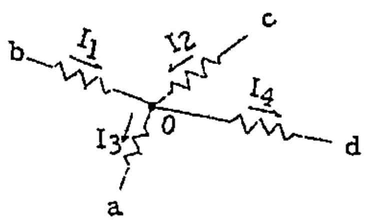


Fig. 2-2

A part of a network is shown in Fig. 2-2 with the direction of current shown for each branch attached to a particular node. At that node, the currents flowing into the node must equal those flowing out, or

$$I_1 + I_2 = I_3 + I_4$$

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On the basis of Kirchhoff's law, there are two ways in analyzing complex networks.

(1) Mesh Analysis

In the mesh analysis, which is based on the first of the two laws, cyclic or mesh currents are assumed for each of the meshes in the network to be analyzed.

Example

Consider the bridge network presented in Fig. 2-3. There are three loops or meshes, and three currents, namely, I_1, I_2, and I_3, have been assumed in the corresponding meshes.

The equations for this networks by Kirchhoff's first law are as follows:

For mesh 1:

$$E = I_1 (Z_1 + Z_2 + Z_3) - I_2 Z_2 - I_3 Z_5$$

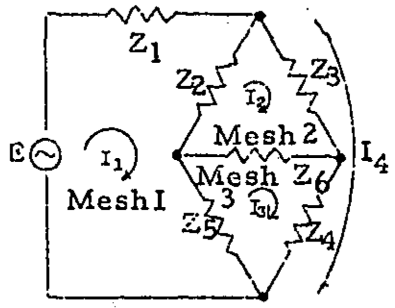

Fig. 2-3

For mesh 2:

$$0 = - I_1 Z_2 + I_2 (Z_2 + Z_3 + Z_6) - I_3 Z_6$$

For mesh 3:

$$0 = - I_1 Z_5 - I_2 Z_6 + I_3 (Z_4 + Z_5 + Z_6)$$

These three linear simultaneous equations must be solved to determine I_1, I_2, and I_3. The current directions and voltage polarities which are assumed are entirely arbitrary. From the very nature of the linear equations obtained in this manner, the initial current directions and voltage polarities will correct themselves as the solution proceeds.

(2) Nodal Analysis

This method has comparatively recently begun to achieve popularity and has distinct advantage over the mesh analysis for certain applications such as vacuum tube amplifier analysis.

In nodal analysis, one node is selected as the reference node and the combination of the reference node with any other node is called an independent node pair.

The selection of the reference node is optional, but the grounded or common side of a network, if any, is usually designated. The nodal analysis is performed by summing up and equating the current components at each node to zero.

Example

Consider Fig. 2-4, in which the nodes are numbered 1, 2, and 3, the third being selected as the reference node. Kirchhoff's second law is utilized o write a current equation for each node.

The complete current equation at node 1 is,

$$- E_1Y_1 + V_1 (Y_1 + Y_3 + Y_4 + Y_5) - V_2 (Y_4 + Y_5) = 0$$

and the corresponding equation for node 2 is

$$- E_2Y_2 - V_1 (Y_4 + Y_5) + V_2 (Y_2 + Y_4 + Y_5 + Y_6) = 0$$

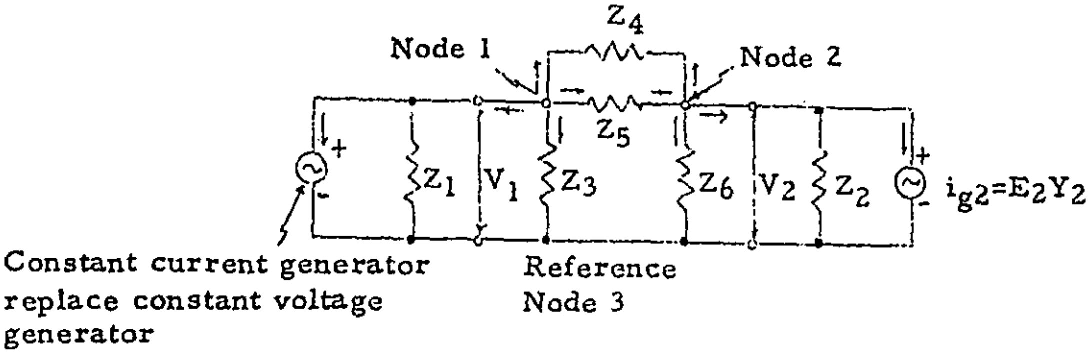

Fig. 2-4

Comparison of Two Methods

Of the two methods, the nodal method is particularly useful because vacuum tube networks lend themselves so readily to this form analysis. This would be expected since one side of such netowrk is quite cften grounded, yielding a convenient reference node, and the various load impedance and interelectrode capacitance appear in parallel.

A paramount feature of the nodal method is that, as long as the number of nodes in a network is not altered, the number of nodal equations will not increase as additional branches are added in parallel to elements already contained in the network. These additional parallel branches will appear merely as additional terms, so that a solution obtained by the nodal analysis may often be modified to take into account additional parallel branches such as interelectrode capacitance which were previously neglected, without requiring a reworking of the whole problem.

2-2. Transient Analysis of Networks

Whenever the conditions in a physical system in which there is an energy-storage element are changed, some time elapses before the system attains a new steady state and a field which deals with such process is called <u>transient analysis</u>.

From basic relationships, voltage drops in transient form in an electrical circuit are represented as

$$\text{Voltage developed across } R = iR$$

$$\text{Voltage developed across } C = Q/C = \frac{1}{C}\int_0^t i\,dt$$

$$\text{Voltage developed across } L = L\cdot\frac{di}{dt}$$

or in current terms

$$\text{Current in } R = e/R$$

$$\text{Current in } C = C\,de\,/\,dt$$

$$\text{Current in } L = \frac{1}{L}\int_0^t e\,dt$$

It is clear that the differential equations normally encountered in network applications are of the first or second order, are linear, and employ time as the independent variable. Even for nonlinear problems, the advantages that accrue by assuming linear performance usually outweigh the errors which result from nonlinearity.

A scheme for simplifying transient analysis by reducing it to algebraic form is called the operational method of differential equation solution and its mathematical foundation is based on Laplace transform.

The Laplace transformation and its inverse transformation are defined as

$$F(s) = \int_0^\infty f(t)\,e^{-st}\,dt$$

$$f(t) = \frac{1}{2\pi j}\int_{\sigma-j\infty}^{\sigma+j\infty} F(s)\,e^{st}\,ds$$

where $S = \sigma + j\omega$

When the above operation has been completed, $f(t)$ will have been converted into an expression which is a function of s alone because the variable t disappears when the integration is performed.
Listings of transform pairs of common functions are easily found in a number of literatures.

The concept of operational impedance can also be introduced, that is, it is obtained directly from the steady-state impedance expression by substitution of *s for* $j\omega$, and it may intuitively be expected that the transient response is a function of the applied excitation and the steady-state response.

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One may profitably study the action of networks either in the time or in the frequency domains, or in a combination of both. The choice of a particular approach depends on the nature of the problem at hand.

(1) Frequency Domain

The transfer function, which is in general complex, may be described in terms of its magnitude and its phase angle. If we write the transfer function of any network in polar form:

$$Y(f) = A(f)\, e^{j\phi(f)}$$

we see that the magnitude and the angle represent, respectively, the amplitude and phase characteristics of the network. In linear system there exists some interrelationship between amplitude response and phase response as well as between real component and imaginary component. In other words, they are not independent and one can be derived from the other.

One then views filtering as a decomposition of an input waveform into its Fourier cosine harmonics. Amplitudes and phase angles of these harmonics are modified by multiplication with the filter's transfer function.

This operation is described by the following general equation

$$E_o(f) = Y(f) \cdot E_i(f) \quad \text{--------------------} \quad (2.1)$$

where

E_i (f) = frequency spectrum of input signal

Y (f) = complex transfer function

After algebraic manipulation, inverse transforms are carried out to recover the time varying solution, if necessary.

Laplace transform offers more versatile means in solving many problems than Fourier transform does, for the former is inherently defined against transient functions.

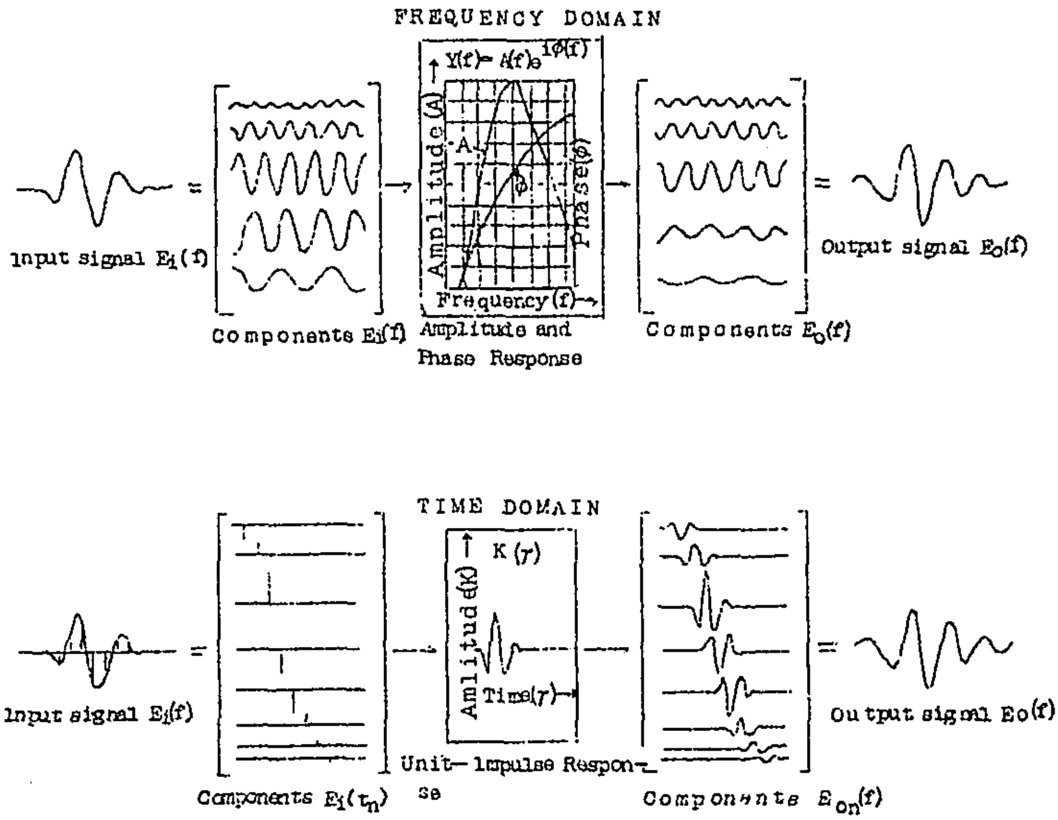


Fig. 2-5

(2) Time Domain

In the time domain the filter characteristics are given uniquely by the unit-impulse response K (τ). The filtering operation is visualized by considering that each ordinate of the input signal E_i (t) represents an impulse.

To define the output signal precisely it is necessary to consider an infinite number of input signal ordinates at infinitesimal intervals; that is, filtering is viewed as a transducer problem with an input E_i (t), and output E_o (t), and unit impulse response K (τ). (See Fig. 2-5). The convolution integral describes a linear time-invariant transformation of an input into an output.

$$E_o(t) = \int_0^t E_i(t-\tau)\, K(\tau)\, d\tau \quad \text{---------------} \quad (2\cdot2)$$

Equation (2·1) and (2·2) are mathematically equivalent. The time and frequency domains are related through Fourier integral transform pairs.

$$K(\tau) = \int_{-\infty}^{+\infty} Y(f)\, e^{j\omega\tau}\, dt$$

$$Y(f) = \int_{-\infty}^{+\infty} K(\tau)\, e^{-j\omega\tau}\, dt$$

Another useful signal form employed in standardizing the characteristics of networks, incidentally, is the step function and the output for this type of input is termed indicial response.

2-3. Network Functions

(1) Terminals and Terminal Pairs

In Fig. 2-6 is shown a symbolic representation of a one-terminal-pair (or two-terminal) netowrk. The terminal pair is customarily connected to a driving force and so is sometimes given the name driving point.

Fig. 2-7 shows a two-terminal-pair network. The terminal pair designated 1 is usually connected to a driving force (or input) while the terminal pair marked 2 is usually connected to a load (as an output). The number of terminal pairs in a network can increase without limit: Fig. 2-8 illustrates a representation of an n-terminal-pair network.

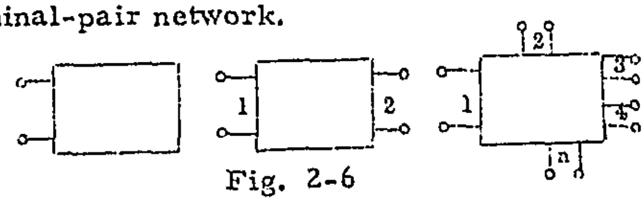


Fig. 2-6

(2) Driving-point Immittance

The impedance or admittance found at a given terminal pair is called a driving-point impedance (or admittance). Because of the similarity of impedance and admittance, the two quantities are assigned one name, immittance (or adpedance).

The driving point immittance of a network is found by combining impedance terms (Ls, R, and I/Cs) or corresponding admittance terms by adding, multiplying, or dividing. This algebraic combination of terms results in an immittance function in the form of a quotient of polynomials as

$$\frac{a_o s^n + a_1 s^{n-1} + \text{-------} + A_{n-1}\, s + a_n}{b_o s \;\; + b_1 s^{m-1} + \text{------} + b_{m-1}\, s + b_m}$$

which is a rational function of s(n and m are integers).

Example

Fig. 2-9 shows an RLC series one-terminal-pair network with transform impedance marked for each element. The driving-point impedance Z(s) is

$$Z(s) = R + Ls + \frac{1}{Cs} = \frac{Lcs^2 + Rcs + 1}{Cs}$$

$$Z(s) = \frac{S^2 + R/L \cdot S + 1/LC}{S}$$

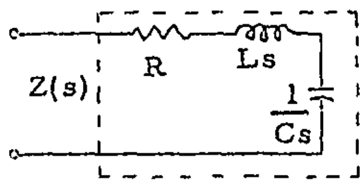


Fig. 2-9

(3) Transfer Functions

The concept of a transfer function is identified with networks having at least two terminal pairs. The function relating the transform of a quantity at one terminal pair to the transform of another quantity at another terminal pair is given the name transfer functions.

There are several forms for transfer functions in electric networks: for example, the ratio of one current to another voltage is called transfer admittance in mhos.

Again, the transfer function can always be reduced to a quotient of polynomials which has the same general form as the driving point immittance function.

Example

The two-terminal-pair network shown in Fig. 2-10 has marked $V_1(s)$ as the input voltage and $V_2(s)$ as the output terminals, the voltage equations are

$$RI(s) + \frac{1}{CS} I(s) = V_1(s)$$

$$\frac{1}{CS} I(s) = V_2(s)$$

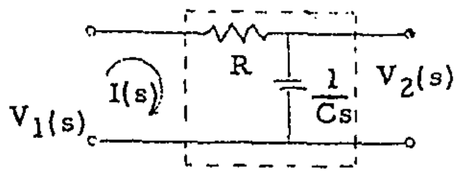


Fig. 2-10

the ratio of these equations is

$$G(s) = \frac{V_2(s)}{V_1(s)} = \frac{1/RC}{S + 1/RC}$$

for this network. This transfer function has a numerator polynomial of zero order and a denominator polynomial of first order.

(4) Poles and Zeros

All network functions have the form of a quotient of polynomials as described above.

If the numerator polynomial is factored into its n roots, and the denominator polynomial is factored into its m roots, the equation can be written in the form

$$H \frac{(s-s_1)(s-s_2) \text{ --- } (s-s_n)}{(s-s_a)(s-s_b) \text{ --- } (s-s_m)}$$

where $H = a_o / b_o$ is a constant known as the scale factor, and the roots s_1, s_2, ----- s_a, s_b, ----- are complex frequencies. When the variable s has the values s_1, s_2, s_n, the network function vanishes. Such complex frequencies are called zeros of the network function.

When s has the values s_a, s_b, ----- , s_m, the network function becomes infinite. These complex frequencies are called poles of the network function.

Poles and zeros are important concepts in network theory because a transfer function is perfectly specified by its poles, zeros, and the scale factor, which provide a great deal of insight into the nature of the response.

(5) Two-terminal-pair (four-pole) Networks

By far the largest percentage of the problems encountered in engineering are of the cause-and-effect type. In an electrical circuit, our main interest does not lie in what is in the black box but rather in what features define the behaviour of this box.

The specialized network configuration which is probably of the greatest importance is the two-terminal-pair, or four-pole, network. It can be shown that four independent parameters (which are generally all a function of frequency) can be employed to specify the performance of a linear four-pole structure.

In Fig. 2-11, a number of sets of equations can be written representing the relations between V_1, I_1, V_2, and I_2.

From the mesh analysis,

$$E_1 = Z_{11} I_1 + Z_{12} I_2$$

$$E_2 = Z_{21} I_1 + Z_{22} I_2$$

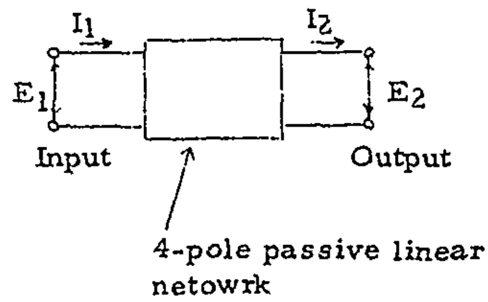


Fig. 2-11

and, from the nodal analysis,

$$I_1 = Y_{11} E_1 + Y_{12} E_2$$

$$I_2 = Y_{21} E_1 + Y_{22} E_2$$

If the voltage and current at the input end of the network are expressed in terms of the voltage and current at the output, then

$$E_1 = AE_2 + BI_2$$

$$I_1 = CE_2 + DI_2$$

Each expression has its own physical meanings.
For example, the parameter A, B, C, and D can be determined by measurement as follows, substituting each condition in original equation:

$A = \dfrac{E_1}{E_2}$ with $I_2 = 0$ (output terminals open)

$B = \dfrac{E_1}{I_2}$ with $E2 = 0$ (output terminals shorted)

$C = \dfrac{I_1}{E_2}$ with $I_2 = 0$ (output terminals open)

$D = I_1 / I_2$ with $E_2 = 0$ (output terminals shorted)

If the structure is passive, then only three of the parameters are independent, therefore, it can be shown that

$$AD - BC = 1$$

****

In addition to the three parameters just treated, what follows may be used on occasion.

a) Image parameters (for design of filtering circuit)

b) h parameters (for analysis of transistor circuit)

****

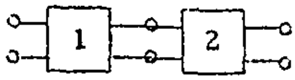
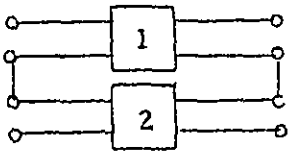
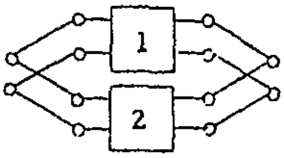
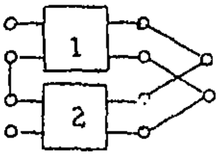
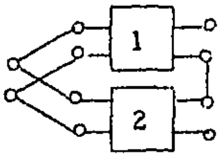
These parameters are all convertible one another and are chosen most profitably according to each given problem.

The conversion table in matrix form is presented in table 2-1.

TABLE 2-1

	$[A]$	$[Z]$	$[Y]$	$[H]$
$[A]$ Transfer matrix	$\begin{bmatrix} A & B \\ C & D \end{bmatrix}$	$\begin{bmatrix} \frac{Z_{11}}{Z_{21}} & -\frac{\lvert Z\rvert}{Z_{21}} \\ \frac{1}{Z_{21}} & \frac{-Z_{22}}{Z_{21}} \end{bmatrix}$	$\begin{bmatrix} \frac{-Y_{22}}{Y_{21}} & \frac{1}{Y_{21}} \\ \frac{-\lvert Y\rvert}{Y_{21}} & \frac{Y_{11}}{Y_{21}} \end{bmatrix}$	$\begin{bmatrix} \frac{-\lvert H\rvert}{H_{21}} & \frac{H_{11}}{H_{21}} \\ \frac{-H_{22}}{H_{21}} & \frac{1}{H_{21}} \end{bmatrix}$
$[A]^{-1}$	$\begin{bmatrix} \frac{D}{\lvert A\rvert} & \frac{-B}{\lvert A\rvert} \\ \frac{-C}{\lvert A\rvert} & \frac{A}{\lvert A\rvert} \end{bmatrix}$	$\begin{bmatrix} \frac{Z_{22}}{Z_{12}} & -\frac{\lvert Z\rvert}{Z_{12}} \\ \frac{1}{Z_{12}} & \frac{-Z_{11}}{Z_{12}} \end{bmatrix}$	$\begin{bmatrix} \frac{-Y_{11}}{Y_{12}} & \frac{1}{Y_{12}} \\ \frac{-\lvert Y\rvert}{Y_{12}} & \frac{Y_{22}}{Y_{12}} \end{bmatrix}$	$\begin{bmatrix} \frac{1}{H_{12}} & \frac{-H_{11}}{H_{12}} \\ \frac{H_{22}}{H_{12}} & -\frac{\lvert H\rvert}{H_{12}} \end{bmatrix}$
$[Z]$	$\begin{bmatrix} \frac{A}{C} & -\frac{\lvert A\rvert}{C} \\ \frac{1}{C} & -\frac{D}{C} \end{bmatrix}$	$\begin{bmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{bmatrix}$	$\begin{bmatrix} \frac{Y_{22}}{\lvert Y\rvert} & \frac{-Y_{12}}{\lvert Y\rvert} \\ \frac{-Y_{21}}{\lvert Y\rvert} & \frac{Y_{11}}{\lvert Y\rvert} \end{bmatrix}$	$\begin{bmatrix} \frac{\lvert H\rvert}{H_{22}} & \frac{H_{12}}{H_{22}} \\ \frac{-H_{21}}{H_{22}} & \frac{1}{H_{22}} \end{bmatrix}$
$[Y]$	$\begin{bmatrix} \frac{D}{B} & -\frac{\lvert A\rvert}{B} \\ \frac{1}{B} & \frac{-A}{B} \end{bmatrix}$	$\begin{bmatrix} \frac{Z_{22}}{\lvert Z\rvert} & \frac{-Z_{12}}{\lvert Z\rvert} \\ \frac{-Z_{21}}{\lvert Z\rvert} & \frac{Z_{11}}{\lvert Z\rvert} \end{bmatrix}$	$\begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix}$	$\begin{bmatrix} \frac{1}{H_{11}} & \frac{-H_{12}}{H_{11}} \\ \frac{H_{21}}{H_{11}} & \frac{\lvert H\rvert}{H_{11}} \end{bmatrix}$
$[H]$	$\begin{bmatrix} \frac{B}{D} & \frac{\lvert A\rvert}{D} \\ \frac{1}{D} & \frac{-C}{D} \end{bmatrix}$	$\begin{bmatrix} \frac{\lvert Z\rvert}{Z_{22}} & \frac{Z_{12}}{Z_{22}} \\ \frac{-Z_{21}}{Z_{22}} & \frac{1}{Z_{22}} \end{bmatrix}$	$\begin{bmatrix} \frac{1}{Y_{11}} & \frac{-Y_{12}}{Y_{11}} \\ \frac{Y_{21}}{Y_{11}} & \frac{\lvert Y\rvert}{Y_{11}} \end{bmatrix}$	$\begin{bmatrix} H_{11} & H_{12} \\ H_{21} & H_{22} \end{bmatrix}$
$[H]^{-1}$	$\begin{bmatrix} \frac{C}{A} & \frac{\lvert A\rvert}{A} \\ \frac{1}{A} & \frac{-B}{A} \end{bmatrix}$	$\begin{bmatrix} \frac{1}{Z_{11}} & \frac{-Z_{12}}{Z_{11}} \\ \frac{Z_{21}}{Z_{11}} & \frac{\lvert Z\rvert}{Z_{11}} \end{bmatrix}$	$\begin{bmatrix} \frac{\lvert Y\rvert}{Y_{22}} & \frac{Y_{12}}{Y_{22}} \\ \frac{-Y_{21}}{Y_{22}} & \frac{1}{Y_{22}} \end{bmatrix}$	$\begin{bmatrix} \frac{H_{22}}{\lvert H\rvert} & \frac{-H_{12}}{\lvert H\rvert} \\ \frac{-H_{21}}{\lvert H\rvert} & \frac{H_{11}}{\lvert H\rvert} \end{bmatrix}$

TABLE 2-2

Circuit configuration	Block diagram	Matrix equation
Two networks in cascade		$[A_{1,2}] = [A_1] \cdot [A_2]$
Two networks with inputs and outputs in series		$[Z_{1,2}] = [Z_1] + [Z_2]$
Two networks with inputs and outputs in parallel		$[Y_{1,2}] = [Y_1] + [Y_2]$
Two networks with inputs in series and outputs in parallel		$[H_{1,2}] = [H_1] + [H_2]$
Two networks with inputs in parallel and outputs in series		$[H_{1,2}]^{-1} = [H_1]^{-1} + [H_2]^{-1}$

The use of the various *forms* given in the table is essential in the calculation of the matrix equations for interconnected four-pole netowrks. As indicated by the column designated matrix equation in table 2-2, which serves to demonstrate the fundamental ways in interconnecting a pair of four-terminal networks, the combined matrix of each of these can be secured by multiplication or addition of the appropriate matrices. One of the most useful cases shown is that of networks in cascade.

2-4. Filtering

Of the great variety of four-terminal networks, filtering circuits occupy extremely significant position.

We classify the functions of filters in accordance with their objectives and characteristics:

- Low pass filter
- High pass filter
- Band pass filter
- Band elimination filter

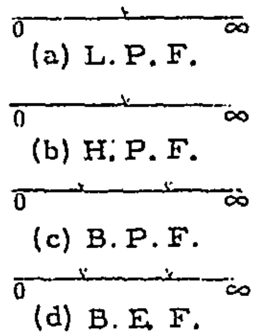


Fig. 2-12

From the standpoint of designing, another classification is made:

- Butterworth (Wagner) filter
- Tchebycheff filter

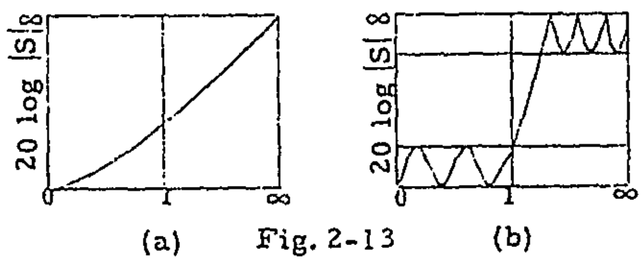


Fig. 2-13

Also in synthesizing filter, it may be viewed in terms of image parameter

- Constant-K filter
- M-derived filter

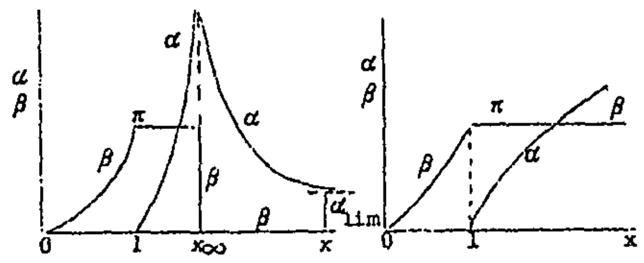


Fig. 2-14

Almost all networks containing nondissipative elements yield, to some extent, frequency-selective characteristics inevitably and hence every network may be considered to perform filtering action in broad sense. Accordingly, the name of filter is here given to those networks designed especially for frequency discriminating function.

It this textbook we confine ourselves to conventional filters, whose behaviour is mainly denoted in the frequency domain. It is expected that the knowledge of the so-called digital filtering which is playing now a vital role in modern data processing technique will be given in the course of seismic method.

The network theory can be divided into two major categories, analysis and synthsis. In analysis, the network is given, and the problem is to determine certain aspects of its performance, while synthesis is broadly defined as methods used in finding an electric circuit meeting certain prescribed specifications.

The general pattern of approach to the synthesis problem is to consider purely reactive networks as the first approximation. Following the establishment of the reactive network pattern, corrections are made for the irremovable resistance effects of the physical counterpart of the purely reactive network.

Although much of the classical design techniques are those of the image parameter theory, it should be stressed that the image parameter does not show true response of that filter and the real value thereof exists in its simplicity in synthesizing desired characteristics as a patchwork.

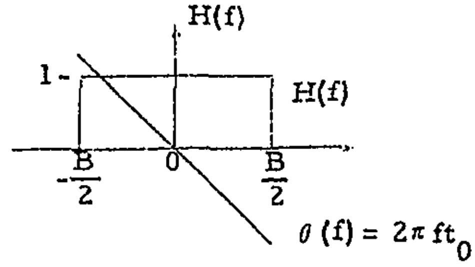
Ideal Filter

It is often useful to review the so-called "ideal low-pass filter" described by the following function

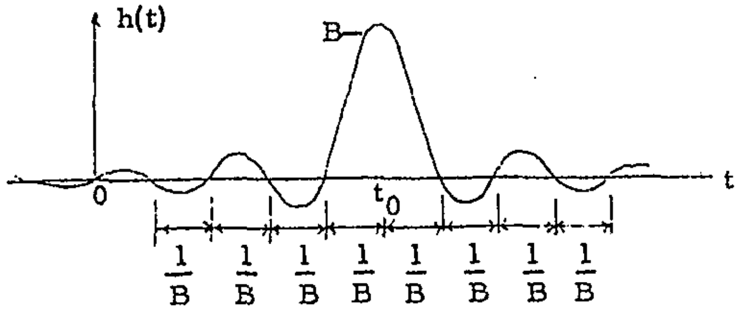
$$H_f = \begin{cases} 1, & |f| \leq B/2 \\ 0 & |f| > B/2 \end{cases}$$

The impulse response of this fictitious filter is

$$I(t) = \int_{-B/2}^{B/2} e^{-j\omega t}\, df = B \cdot \frac{\sin \pi Bt}{\pi Bt}.$$



(a) Filter frequency function



(b) Unit impulse response

Fig. 2-15

As indicated in Fig. 2-15, it is apparent that "ideal filter" is impossible to realize physically without infinite delay because the response precedes the input, namely it does not satisfy the law of causality.

Because frequency response of R, L, and C, are continuous with the exception of resonance points, a network containing these elements can not be made to cut off abruptly as depicted in ideal filter's response. Instead, we can realize low pass filter which have the magnitude characteristics approximated as close as ideal response.

With respect to this a highly refined technique has been developed in approximating desired frequency response.

§3. Some Aspects of Electronic Systems

It is expected that transistors will play an increasing important role in electronic circuitry as the years go by. Nowadays the positions occupied by vacuum tubes have been almost replaced by semiconductors except for the use in special purposes.

Nevertheless, examples are frequently illustrated here with vacuum tube representation because many of the classical vacuum tubes configurations may be adapted to transistors by direct analogy although basic differences between them do exist, and it gives rather easier insight of the problem.

Fig. 3-1 serves to demonstrate the corresponding configurations between these two elements.

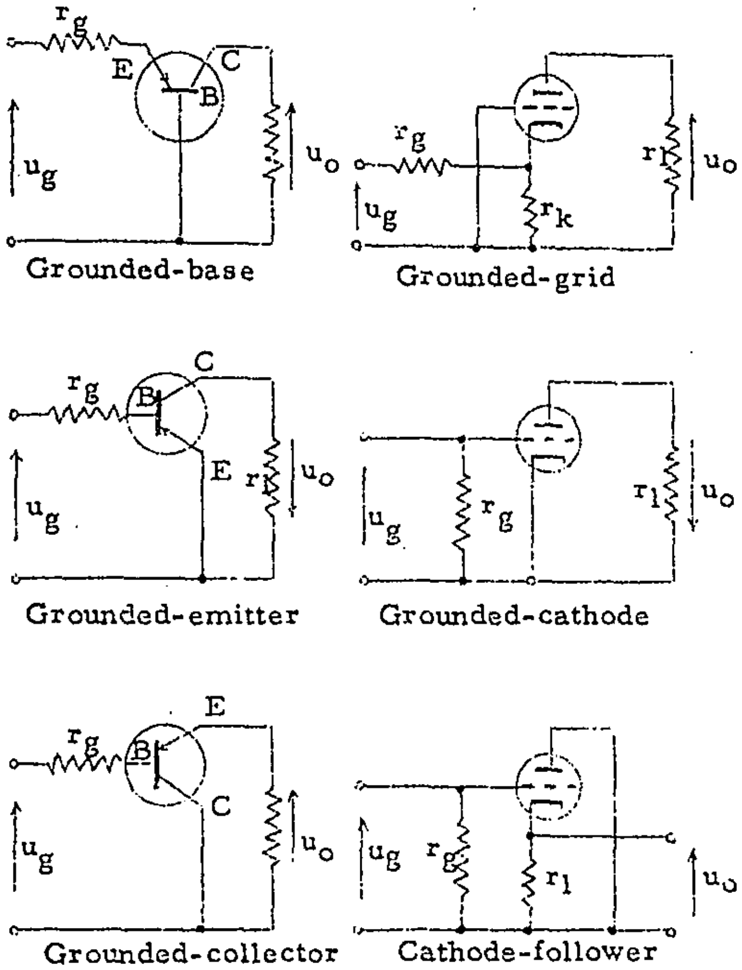


Fig. 3-1

Signal to Noise Ratio

All electrical components generate noise, owing to microscopic fluctuation phenomena associated with their macroscopic properties.

A familiar example is the resistor which generates the thermal agitation noise due to heat energy of electrons.

$$\bar{e}_n^2 = 4\,KTBR\,(V^2)$$

K = Boltzman constant

T = absolute temperature

B = frequency band

R = resistance

Electrical signals in some geophysical instrument have an extremely low level, falling within the range from $1\mu v$ to 50mv. Obviously, these signal must be amplified in order to raise the signal to usable levels. Unfortunately, while an amplifier performs its function, it also creates noise. There is no way to reduce this noise to an absolute zero value, all that can be done is to minimize it.

As some geophysical signals have a level of only 1 μv, the system or amplifier noise must be less than this so as to identify the signal. Typically, good instrumentation has an equivalent noise level (referred to as the input) of 0.2 μv or less, while system malfunction can raise the equivalent noise level by factors of 10 to 100 or even 1000 in serious cases. Under these circumstances it becomes impossible to recognize low level geophysical signals.

It is the relative value between useful signal and noise level that determine the quality of any transmission system, not absolute one. Thus a criterion of the signal to noise ratio is defined as:

$$S/N = \frac{\text{Peak instantaneous output signal power}}{\text{output noise power}}$$

and the concept of SN ratio is broadly extended to other categories of data acquisition system. Improvement of SN ratio often forms a focal point of engineering practice in a field or in a laboratory.

3-1. Amplifier and Linear Distortion

Of all the applications of vacuum tube or transistor circuitry, amplification of weak signal is considered the most elementary one and involves many important concepts which are, to a large extent, associated with other branches.

Amplifiers are classified according to the frequency range, the method of interstage coupling and bias condition. For example, they may be classed as direct-coupled amplifiers, audio-frequency amplifiers, video amplifiers if some indication of the frequency of operation is desired. Also, the position of the quiescent point to which an input is swung will provide the classification such as class A, class B, or class C, as is well known in classical radio engineering.

It is frequently necessary to achieve a higher gain in an amplifier than is possible with a single amplifier stage. In such cases, the amplifier stages are cascaded to accomplish this higher gain. Because of the several features that play a part in amplifier design, however, gains in excess of about 120 db potential gain, are extremely difficult to achieve.

A variety of coupling networks between the cascaded stages are possible, but the resistance-capacitance coupled amplifier is, in particular, one of the more common and more useful amplifier circuits.

The circuit of Fig. 3-2 is representative of a complete stage of R-C coupled vacuum tube amplification in that it includes the elements of a single amplifer and the elements required to couple this stage to the next.

The detailed analysis of the characteristics of this circuit shows that the gain is substantially constant over a range of frequencies and falls off at both the low and the high frequencies. A typical frequency-response curve has the form sketched in Fig. 3-3.

An outstanding feature to be mentioned is that while there is an increase in in the bandwidth of R-C coupled amplifier, the gain is thereby reduced; one is obtained at the expense of the other for a given tube and a given circuit configuration. In other words, the gain-bandwidth product is a constant such that

$$GB = \frac{g'_m}{2\pi C}$$

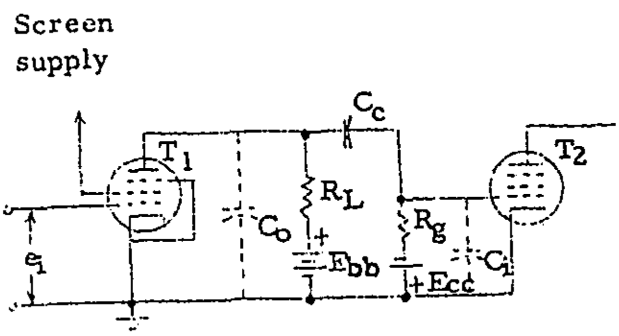


Fig. 3-2

where C indicates effective capacitance which is the sum of output capacitance of the final stage and input capacitance to next stage.

From above expression, a quantity M is defined as

$$M \equiv g_m / C$$

M is known as the Figure of Merit of the vacuum tube.

Obviously, for service requiring a large gain-bandwidth product, the tube should possess a large transconductance in proportion to the input plus output electrode capacitance, and this is the generally the case for a broad-band amplifer.

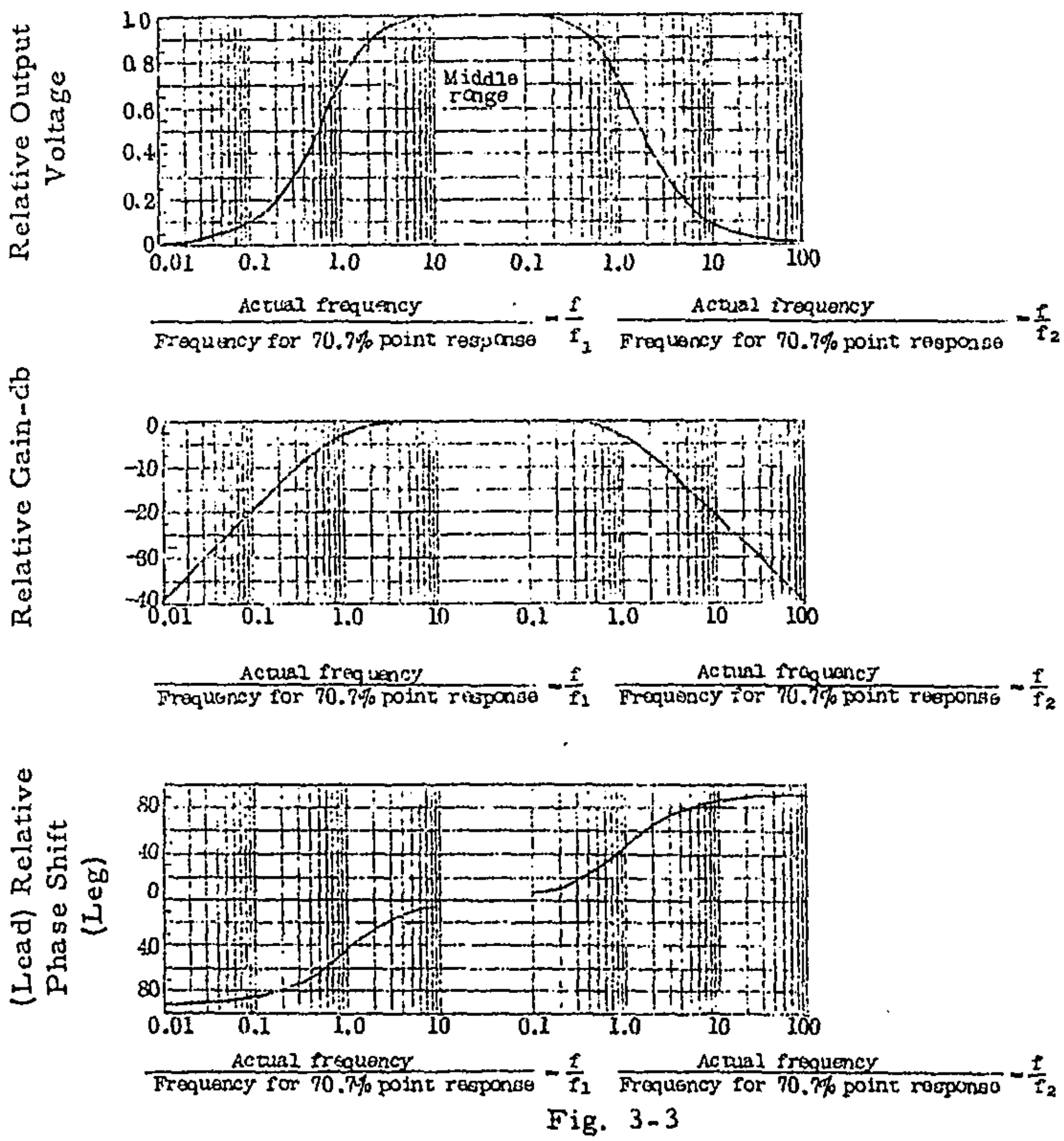


Fig. 3-3

Linear Distortion

Frequently the need arises in electric systems for transmitting a signal with a minimum of distortion.

A criterion which may be used to compare one transmission network with another with respect to fidelity of reproduction of the input signal is suggested by the following considerations.

Any arbitrary waveforms of engineering importance may be resolved into Fourier spectrum. If the gain and time delay of a system are independent of the frequency, then the network must necessarily reproduce precisely the form of the input waveshape.

The necessity of constant gain over the required frequency range is self-evident, although it is seldom realized and causes, more or less, amplitude distortion. In addition to distortion created by the amplitude response, the input signal is further deteriorated by the phase response when the phase-versus-frequency curve is nonlinear (phase distortion.)

Now suppose a signal having wholly contained frequency spectrum within the pass-band of the network whose phase function is linear with slope . In this case the output signal is simply a delayed version of the input signal. Except for the time delay the signal has passed through the network without any modification or distortion. Repeating again, ideal signal transmission requires the network's magnitude function to be constant and the phase characteristics to be a linear function of frequency.

Any departure from these conditions leads to signal distortion. Thus the conditions for distortionless transmission are summarized as follows:

$$|H_f| = \text{constant}$$
$$\theta_f = \text{linear phase}$$

This holds over the entire band of signal frequencies.

In transmitting human voice or music, it is generally said that linearity requirements for phase characteristics are not exacting because our auditory sense is not capable of discerning the effect of variation of phase.

Pulse Testing of Linear Systems

The response of networks to pulses of various shapes can be theoretically computed. In experimental scheme, on the other hand, many electronic systems not primarily designed for pulse transmission, such as audio amplifiers, can be profitably tested with pulses.

If the pulse waveform is properly chosen, a transient measurement provides exactly the same information as a steady-state frequency response measurement but provides it in a different form.

A number of waveforms are conceived and one should choose most suitable one in accordance with each network configuration and specific point of view.

Fig. 3-4 gives general aspect of step response, wherein terminology of each part of distortion is also indicated.

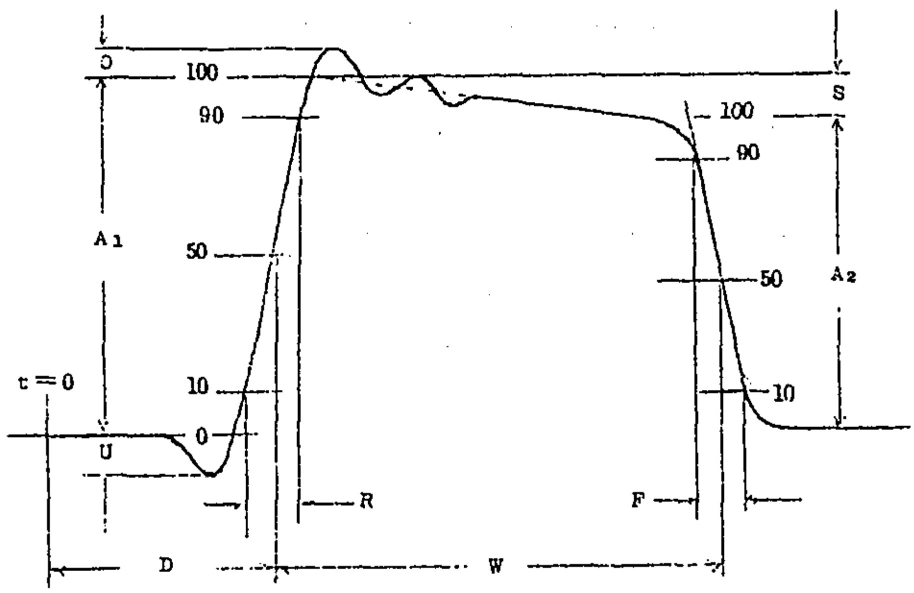


Fig. 3-4

R: rise time

F: fall time

D: delay time

W: pulse width or duration

S: sag

O: overshoot

U: undershoot

3-2. Linear Wave Shaping

The most important nonsinusoidal waveforms encountered in electrical ircuitry are the step, square wave, ramp, impulse, and exponential wave-orms. The response to these signals of linear networks is of great interest.

We take simple RC networks, which is in practice valuable, as a typical xample of such kind of shaping.

For the circuit of Fig. 3-5, the magnitude of the gain A and the phase angle θ are given by

$$|A| = \frac{1}{\left[1 + \left(\frac{f_1}{f}\right)^2\right]^{\frac{1}{2}}} \qquad \text{and } \theta = \arctan\frac{f_1}{f} \qquad \text{where } f_1 = 1 / 2\pi RC$$

At this frequency, f_1, the magnitude of the capacitive reactance is equal to the resistance and the gain is 0.707. this drop in signal level corresponds to a signal reduction of 3dbs and the circuit works as a high pass filter because it passes high frequencies readily, but attenuates low components.

From elementary considerations, the indicial response of the network is given by

$$e = E \cdot e^{-t/RC}$$

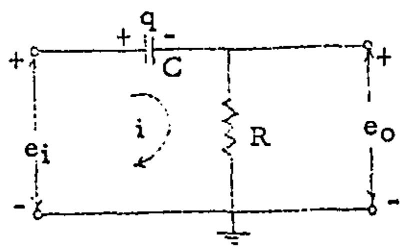


Fig. 3-5

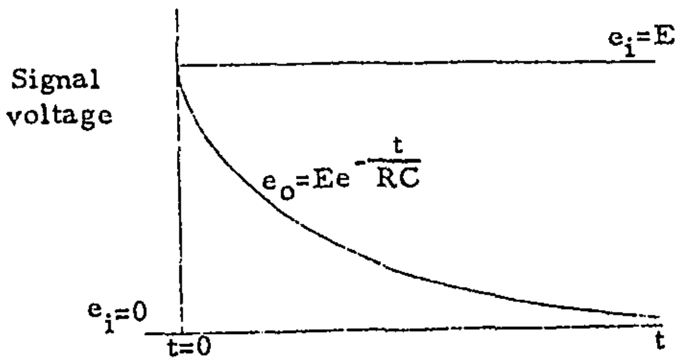


Fig. 3-6

The square pulse may be synthesized to be the sum of two voltages and its response is shown in Fig. 3-7.

There is a tilt to the top of the pulse and an undershoot at the end of pulse. If these distortion are to be minimized, then the time constant must be large compared with the pulse width. However, if the time constant is very small ($RC \ll t_p$), the output consists of a positive spike at the beginning and a negative spike of the same size at the end of the pulse. More generally, the response to a square wave train must have the appearance shown in Fig. 3-8. This process is called peaking or differentiation, where name arises from the fact that the output is proportional to the derivative of the input such that

$$e_o = RC \frac{de_i}{dt}$$

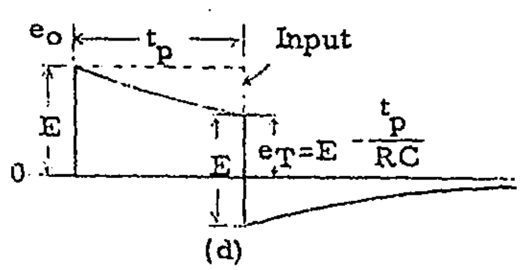


Fig. 3-7

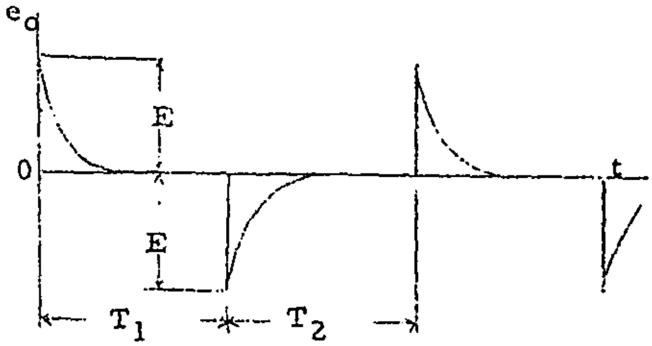


Fig. 3-8

As opposed to the preceding example, the RC network shown in Fig. 3-9 acts as a low pass filter. For sinusoidal input, the frequency response is given by

$$\left| A \right| = \frac{1}{\left[1 + \left(\frac{f}{f_2} \right)^2 \right]^{\frac{1}{2}}} \qquad \text{and} \quad \theta = - \arctan \frac{f}{f_2}$$

where $f_2 = 1/2\pi\, RC$. Again f_2 is called 3 db frequency. Also shown are step response and pulse response in Fig. 3-10 and Fig. 3-11.

In analogy to "differentiator", this circuit has a function of "integration" as

$$e_o = \frac{1}{RC}\int e_i \, dt$$

provided the time constant is very large in comparison with the time required for the input signal to make an appreciable change. Since the output is a small fraction of the input in both differentiator and integrator, the output will frequently have to be followed by a high-gain amplifier. Any drift in amplifier gain will, however, affect the level of the signal, and amplifier nonlinearity may affect the accuracy of differentiation (integration). These difficulties are avoided by using the operational differentiator (integrator) in which a feedback loop is incorporated in amplifier and stability thereof depends principally upon the constancy of R and C.

Note the foregoing equivalent characteristics can also be effected by the combination of a resistor and an inductor. However, the inductor is seldom used if a large time constant is called for because a large value of inductance may be manufactured only with an iron core which is physically bulky, heavy, and hence expensive relative to the cost of a capacitor for a similar application.

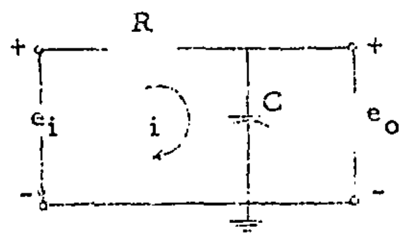


Fig. 3-9

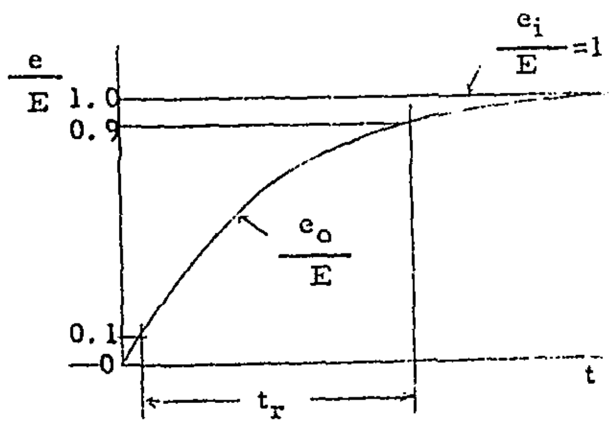


Fig. 3-10

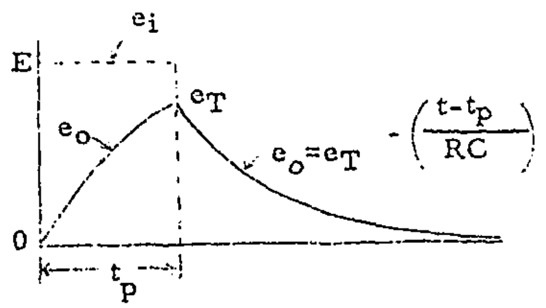


Fig. 3-11

3.3. Nonlinear Wave Shaping

In considering the process of wave shaping by nonlinear elements we idealize the characteristics of diode so that the ratio of applied voltage to the current, called the forward resistance, is zero when conducting and the same ratio, called the back resistance is infinite for negative voltage.

Clipping

Clipping circuits, also referred to as amplitude selector, are used when it is desired to select for transmission that part of an arbitrary waveform which lies above or below some particular reference level.

The most common selector is the diode-series resistance circuits that are illustrated in Fig. 3-12.

These circuits are essentially the same and differ only in the polarity with the diode is inserted and in whether the voltage is taken across the diode or resistor.

By combining two clippers, one may construct slicer, or limiter, whose output contains a slice of the input between the two reference levels, as presented in Fig. 3-13.

The process provides a means for generating waveforms including segments parallel to the horizontal

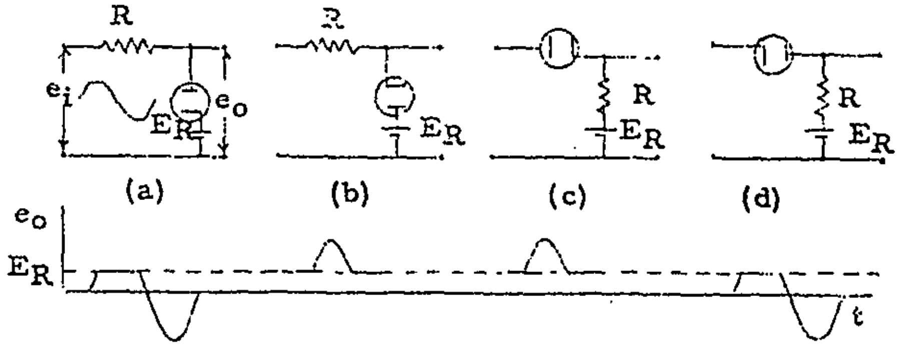


Fig. 3-12

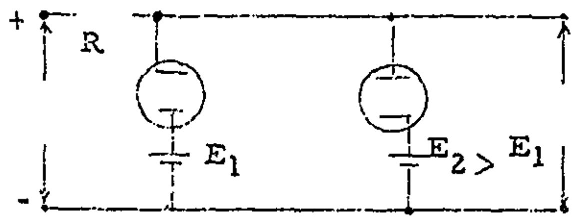
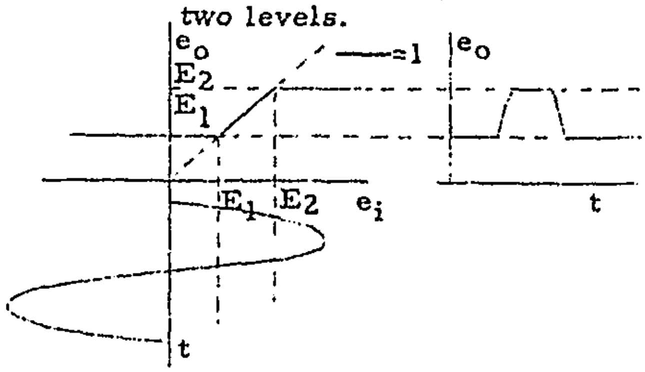


Fig. 4-10. A circuit for clipping at two levels.



axis, such as trapezoids or square waves.

The diode selector is extensively used. If the wave to be amplified, however, and if accurate boundary setting is of secondary importance, a grid tube is desirable. A triode, for example, limit a signal when the grid is driven beyond cutoff.

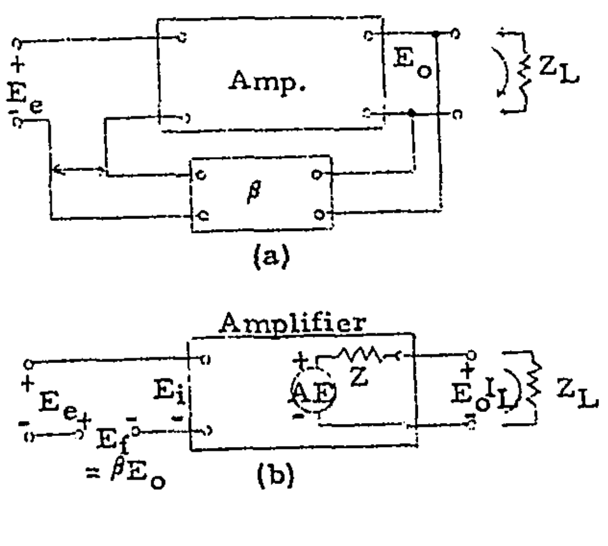
Power Supply

Notice that the foregoing circuits can be used as an rectifier in obtaining d-c potential from an a-c line. However, it is usually the requirement of a power supply to provide a relatively ripple-free source of d-c potential from an a-c line. It is customary therefore to include a filter between rectifier and the output to attenuate ripple components. Often electronic regulator is also included to stabilize the output.

3-4. Feedback System

Where accurate control of some desired quantity is required (such as temperature, flow, speed or position), it is necessary to compare the controlled quantity with some reference value, and to adjust the system in such a way that any difference between actual and desired output is reduced. The term servomechanism is often applied to such a system, whose underlying principle needs a function of feedback essentially. Moreover, many electronic systems other than servomechanism incorporate a feedback path by means of which a part of the output is reintroduced at the input.

In the feedback arrangement of Fig. 3-14, the signal at the input terminals to the amplifier is the sum of the externally impressed voltage E_e and a feedback voltage $E_f = \beta E_o$.



The feedback voltage is related to the output voltage by a factor β , which is determined by the feedback network. Let A be the forward gain without feedback (the open-loop gain). We may write

$$E_o = A\ E_i - I_L\ Z \qquad \text{and} \qquad E_i = E_e + \beta\ E_o$$

Eliminating E_i from these equations, we find

$$E_c = \frac{A}{1 - A\beta}\ E_e - \frac{Z}{1 - A_1\beta}\ I_L$$

Hence we conclude that the gain and output impedance with feedback are given by A_f (the closed-loop gain) and Z_f, respectively.
where

$$A_f = \frac{A}{1 - \beta\ A} \qquad\qquad Z_f = \frac{Z}{1 - \beta\ A}$$

The following properties of feedback amplifiers are to be noted.

a) Stability

If the feedback is negative, so that $|1 - \beta\ A| > 1$, the feedback will have served to improve the gain stability of the amplifier.

In particular, if $|\beta A| \gg 1$, then

$$A_f = \frac{A}{1 - \beta\ A} \cong -\frac{1}{\beta}$$

and the gain may be made to depend entirely on the feedback network.

In active networks or system under certain conditions this feedback may become positive and sometimes amplifier may change into oscillator.
The prevention of oscillation in feedback amplifier is a serious difficulty which oftenoffsets the advantages of negative feedback and care must be taken to prevent the system becoming unstable. It should be remembered, however, that the general condition for oscillation involves more complicated concepts.

b) Frequency Distortion

It follows from the equation $A_f \cong -1/\beta$ that if the feedback network does not contain reactive elements then the overall gain is not a function of frequency. Under these circumstances a substantial reduction in frequency and phase distortion is obtained. It is to be noted, however, that negative feedback improves frequency response only at the expense of gain.

****

The feedback principle may be employed with electronic amlifiers to make them perform mathematical operations, in particular, sign inversion, summation and integration. These operational amplifiers form the basic building blocks of electronic analog computers.

Cathode Follower

An example of a circuit which may profitably be viewed as a feedback amplifier is the cathode follower of Fig. 3-15.

The principal characteristics of the cathode follower may be summarized as:

1) Low output impedance
2) No inversion of the input signal
3) Gain is less than one but can be made almost equal to unity

A cathode follower is usually employed when a low output impedance is needed whenever it is required to transmit a signal over a relatively long distance, the capacitive loading of the long wires is minimized by taking advantage of the low output impedance of the cathode follower.

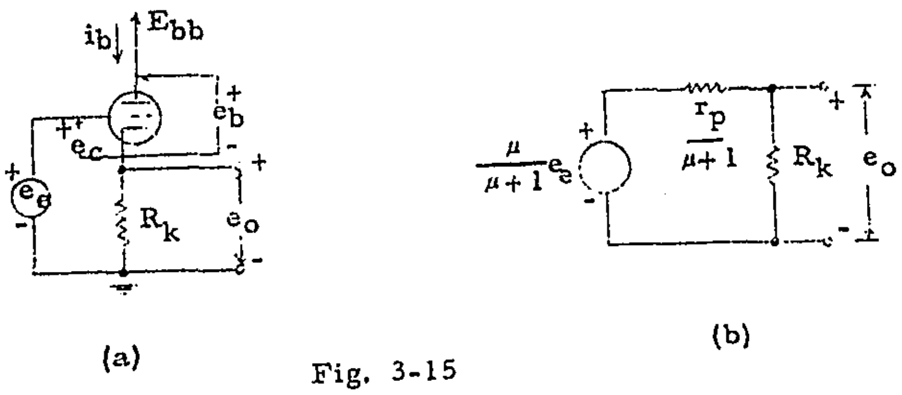


Fig. 3-15

3-5. Multivibrator

The multivibrator finds extensive application in pulse circuitry and its function is summarized in Fig. 3-20.

A circuit with zero stable state, called "astable" or "free-running" multivibrator, generates a continuous train of waves and requires no triggers to execute a complete cycle while a monostable multivibrator has one permanently stable state and quasi-stable state.

The basic application of the monostable multi. results from the fact that it may be used to establish a fixed time interval, the beginning and end of which are marked by an abrupt discontinuity in a voltage waveform.

The astable multivibrator is an oscillator and is used as a generator of "square waves."

Finally a bistable multivibrator, or "flip-flop", or "binary" has instead two stable states in either one of which it may remain permanently. It is employed not only for the generation of square waves but also for the performance of certain digital operations, such as counting.

In Fig. 3-21 are shown the examples of three types of multivibrators.

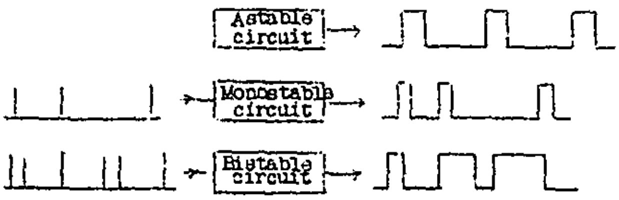


Fig. 3-20

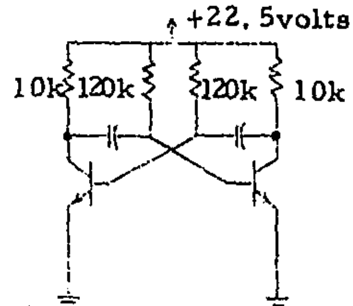


Fig. 3-21 Astable multi.

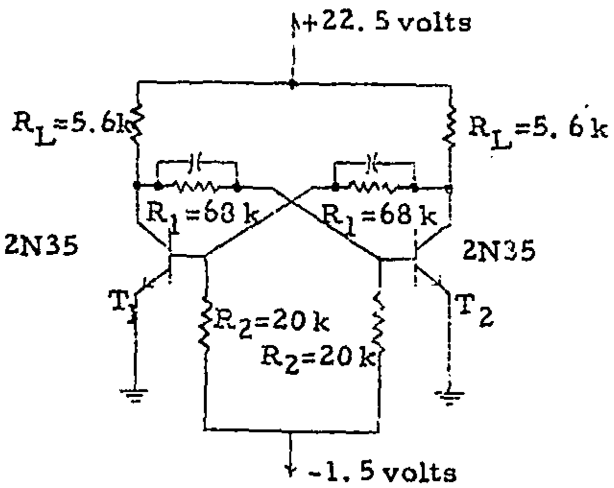


Fig. 3-21 Bistable multi.

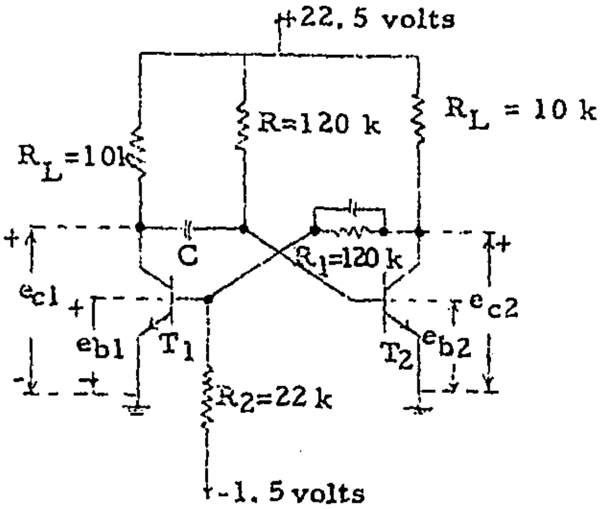


Fig. 3-21 Monostable multi.

3-6. Pulse Counting

Pulse counting technique has widespread application in many branches of geophysical instrumentation. One of the earlier applications arose in connection with the booming demand of the well-known Geiger counter or scintillation counter used popularly in the prospecting for radioactive minerals.

In a modern sophisticated proton magnetometer, whose principle lies in the measurement of free precession frequency of nuclear magnetic moment, high precision may be accomplished by a pulse counting system with a reference oscillator and an appropriate gating arrangement.

Similar technique is also adapted to portable refraction seismograph, combined with hammer percussion as a seismic energy source, mostly being employed in civil engineering.

(1) Binary Chain

Consider the cascade connection of four binaries as shown in Fig. 3-22-(a). A sequence of triggering pulses is applied to the first binary labelled B_0, whose output signal is in turn fed to the second binary B_1. Similarly, B_1 is coupled to B_2 and B_2 to B_3. A typical coupling arrangement is presented in Fig. 3-22-(b). Referring to Fig. 3-23, one may easily recognize that this binary chain can be used to divide the number of input pulses by a factor of 16. That is to say that each time 16 pulses are applied at the input, 1 pulse appears at the output. After 16 pulses the circuit will reset itself into its original state.

It should be clear that if n binaries are in tandem, division by the factor 2^n will be accomplished. The counting mechanism of the chain is intimately associated with the binary system of representing numbers in which the base is 2.

In the more familiar decimal system the base is 10, and ten numerals are called for to express an arbitrary number. A device with ten stable states will apparently be suitable for use in a decimal counting system and the most common of such technique involves feedback of pulses from succeeding stages to preceding stages of the chain (decimal counter).

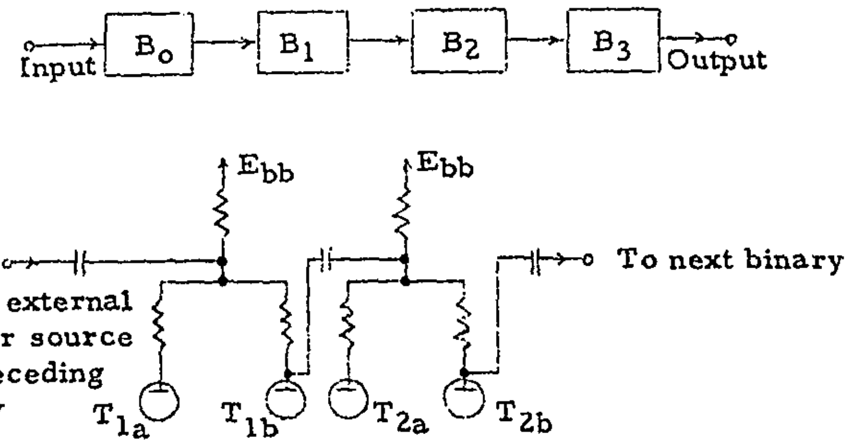


Fig. 3-22

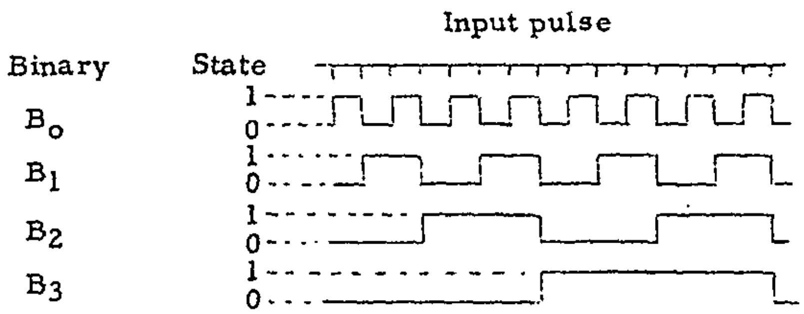


Fig. 3-23

(2) Storage Counter

A two-diode storage circuit, while operates on a principle basically different from the binary, is shown in Fig. 3-24. Each successive input pulse will cause a progressively smaller step in voltage at the output, the output approaching asymptotically the voltage $e_o = E$, as in Fig. 3-25, and the counter is completed by the addition of a circuit which operates as a switch shunted across C_2. this switch is normally open, but it closes when e_o attains some preestablished reference value.

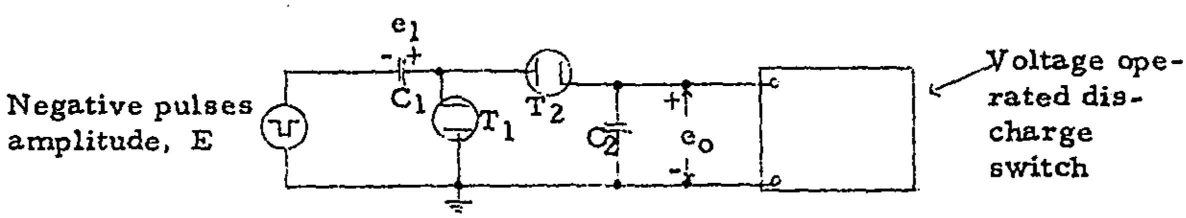


Fig. 3-24

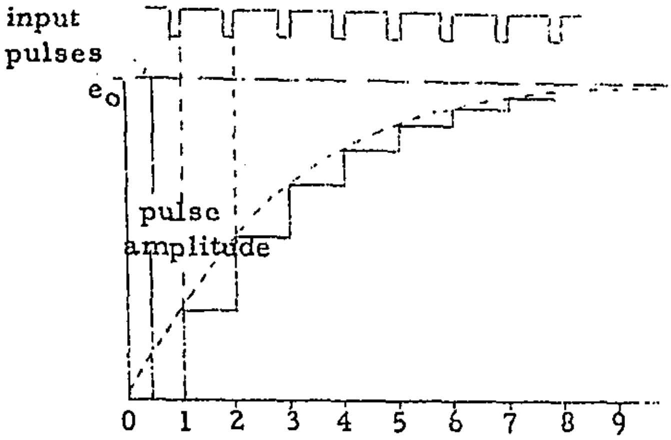


Fig. 3-25

(3) Counter Tube

The fundamental decade counter consisted of binaries requires a minimum of four double triode or equivalent. This complexity has naturally prompted investigations into the design of special tubes which would permit a greater economy.

It suffices here to cite several examples of special tubes developed for this purpose

Gas-filled counter tube : Dekatron

Vacuum type counter tube: Trochotron

: EIT

3-7. Modulation

The object of many electronic system is to convey intelligence of some type from an information source to an intended distant user. When converted to simplest electrical form, the intelligence usually has low-pass spectrum. However, to make efficient use of the available transmission media a narrow-band (bandpass) transmission at much higher frequency is required. The process of transforming intelligence into this form is termed modulation.

At the receiving terminal, after passing through the medium and possibly being corrupted by interference, the bandpass waveform is recovered from the medium to retrieve a reasonable facsimile of the original information.

All modulation techniques involve at least two quantities. One of these is the modulating waveform containing the message to be transmitted. The second quantity is the high frequency carrier wave, which is modulated by an appropriate variation of one or more of its parameters.

1) Frequency Division

In general, a continuous wave is written as

$$e\ (t) = E\ (t) \cdot \mathrm{COS}\ \phi\ (t)$$

and Momentary Angular Frequency is defined as

$$\omega_m = d\ \phi\ (t)\ /\ dt$$

Modulating wave can vary any of these quantities: $E\ (t)$, ω_m, or $\phi\ (t)$, and they called, amplitude modulation (AM), frequency modulation (FM) and phase modulation (PM), respectively.

(AM)

The distinguishing feature of AM is that the modulated carrier envelope is a true reproduction of the modulating wave. Conceptually, AM waves can be generated by adding a constant on average value to the modulating wave and multiplying the sum with a sinusoidal carrier. The resulting AM wave is

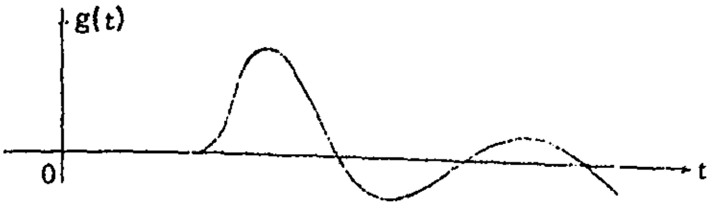
$$E\ (t) = E_m + g\ (t)$$

$$e\ (t) = E\ (1 + m \cdot g\ (t)\ )\ \cos\ \omega_o t$$

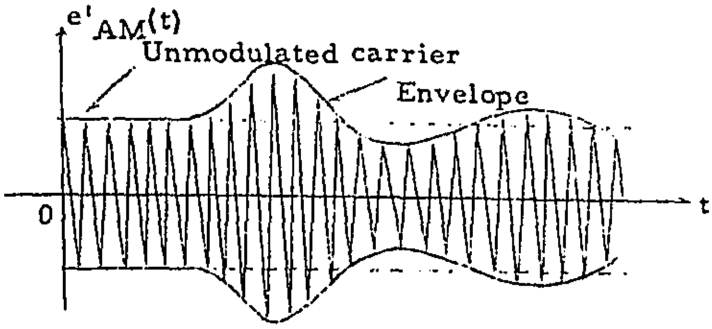
where

$$m = \text{modulating factor}$$
$$E \cos \omega_o t = \text{carrier wave}$$
$$g\ (t) = \text{modulating wave}$$

For a given bandpass spectrum, the AM signal spectrum E_{AM} (f) is expressed from the relationship

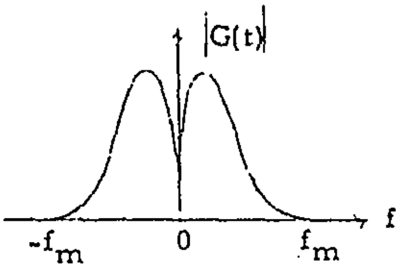


(a) Modulating wave

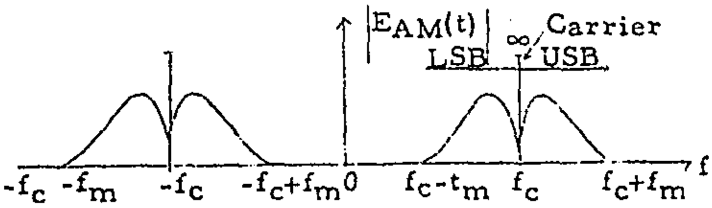


(b) Modulated carrier

$$E_{AM}(f) = \underbrace{\frac{1}{2}\left[E(f-f_c)+E(f-f_c)\right]}_{\text{carrier}} + \underbrace{\frac{m}{2}\left[G(f-f_c)+G(f+f_c)\right]}_{\text{sidebands}}$$



(a) Basebond spectrum



(b) AM spectrum

Fig. 3-26

This is shown in Fig. 3-26.

The AM signal spectrum contains a carrier components in addition to frequency translation of the baseband spectrum to the vicinity of the carrier.

Ordinarily, AM transmits both sidebands, either one of which fully describes the information. If the required transmission bandwidth can be compressed further, more information channels can be placed in a given frequency band.

This capability exists in a modified AM technique that halves the transmission bandwidth by eliminating the carrier and one sideband. This process is known as single-sideband (SSB).

(FM and PM)

Angle modulation is a generic term applied to both FM and PM, which are closely related technique. Angle modulation also involves frequency translation, but in contrast to AM, it transforms the message spectrum into an entirely new set of related frequency components, which is, generally, broader than the original signal spectrum.

Angle modulation often finds application where high fidelity or accuracy of message waveform reproduction is required, because these system more successfully overcome noise disturbances than do AM systems. However, a price is paid for this advantage, part of which lies in increased bandwidth requirements.

By reference to Fig. 3-27, in which schematic comparison between FM and AM is shown, it is evident that FM is less susceptible to disturbances.

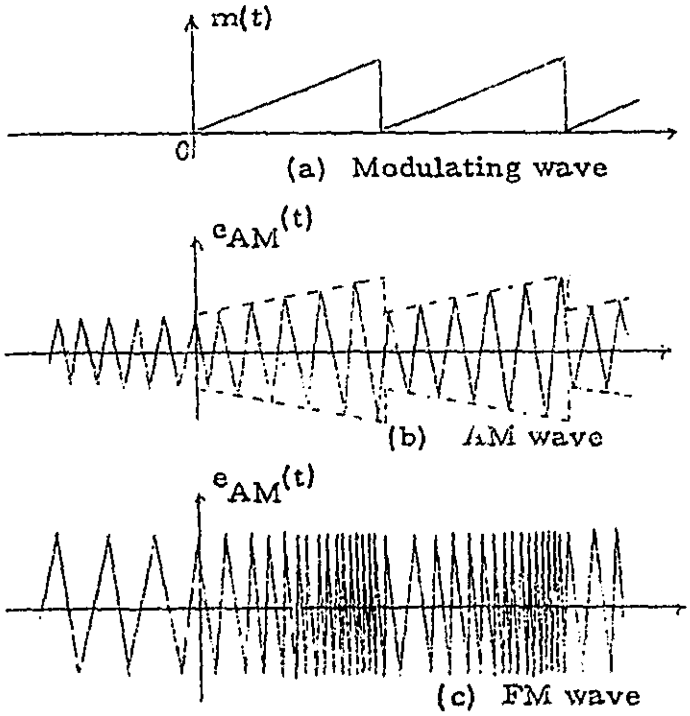


Fig. 3-27

2) Time Division

Another useful form of modulating system is pulse modulation, in which the carrier function is a uniform pulse sequence or pulse train, some parameter of which is modified in accordance with the modulating signal.

For example, in pulse amplitude modulation (PAM) systems, the modulation is carried out in variations of the individual pulse amplitudes. In pulse width modulation (PWM) and pulse position modulation (PPM) systems, the modulating signal changes, respectively, the pulse width and relative time of occurrence of the individual carrier pulses.

Sampling of the information source at a rate exceeding twice the highest frequency contained in the message waveform is a fundamental operation to all forms of pulse modulation for conveying analog waveforms.

Since pulse modulation involves in sampled form, it also provides the opportunity for message from several information sources to share common communications equipment and a single transmission channel by utilizing these "vacant" time intervals between samples of any one message.

Perhaps the most important aspect of pulse modulation and of all forms of digital data transmission is an inherent immunity to noise and disturbances similar to, and in some ways even better than, FM system. The degree to which this noise immunity is effective is dependent upon the individual techniques.

Quantized and coded systems, such as pulse code modulation (PCM), excel in this respect, and are employed in digital recording system of exploration seismology.

Fig. 3-28 serves to demonstrate various methods of pulse modulation system.

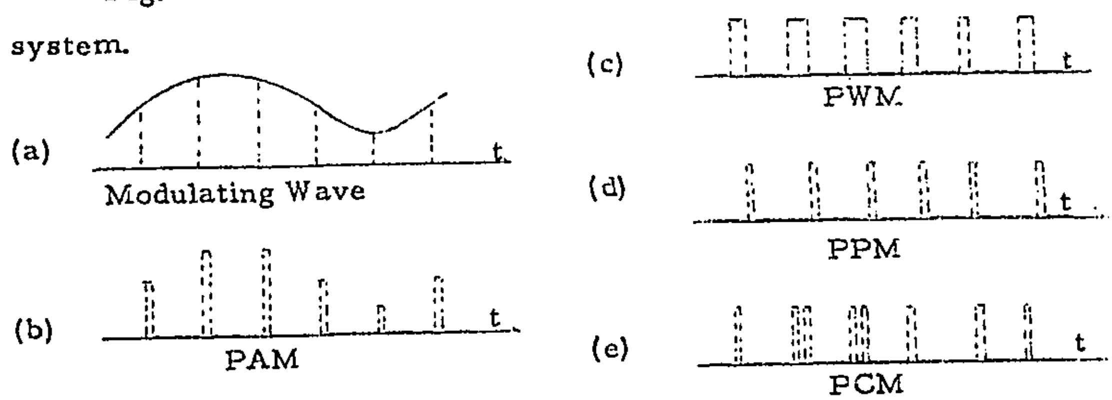


Fig. 3-28

3-8. Digital Computer Circuits

Switching or logical gate is a device having several inputs and one output, and finds wide applicability in all fields of pulse circuitry, but is used particularly in digital computers.

a) OR Circuit

If a pulse is applied to any one or more of the inputs of this circuit, a pulse appears at the output. The circuit derives its name from the fact that an output pulse appears when a pulse is applied at input 1 <u>or</u> input 2 <u>or</u> any other input.

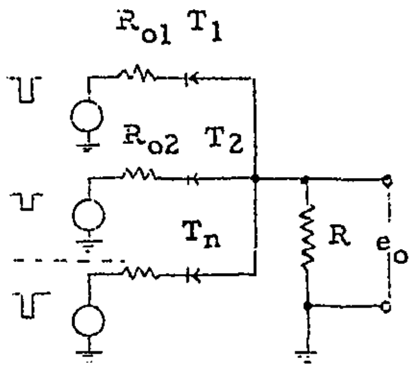


Fig. 3-16

b) AND Circuit

The AND circuit (also called a coincidence circuit) has a single output at which a pulse is applied simultaneously to all inputs. If the input pulses are not of the same time duration, the output pulse will appear during the time interval of the input pulses overlap.

The OR circuit may be modified for use as AND circuit by the simple expedient of adjusting the circuit.

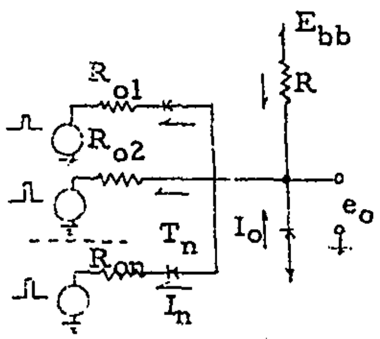


Fig. 3-17

c) NOT Circuit

A circuit which inverts the polarity of a pulse is termed an INVERTER or NOT circuit. A plate-loaded triode, for example, therefore constitutes a NOT circuit, however, a NOT circuit is strictly defined by the operation indicated in Fig. 3-18, that is, it inverts the wave form but keeps the variable operating between the same two limits, E_1 and E_2, as shown.

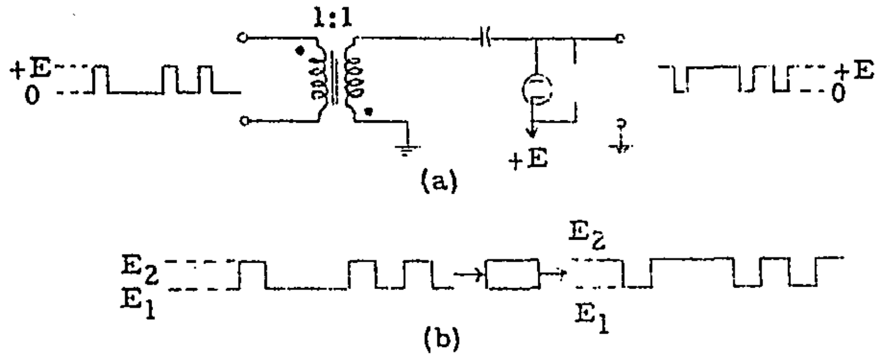


Fig. 3-18 (a) A NOT circuit. (b) Defining the operation of a NOT circuit.

d) INHIBITOR Circuit

Suppose that we have an AND circuit with N+I inputs and that the (N+I)st is preceded by a NOT circuit, as in Fig. 3-19. Such a circuit is called a NOT-AND circuit, an INHIBITOR, or an anticoincidence circuit and has the property that an output pulse will appear if and only if pulses are applied simultaneously to input 1 to n and no pulse is applied at the (n+I)st input. Frequently no special recognition is taken of the NOT component of the INHIBITOR and the circuit is represented as in Fig. 3-19.

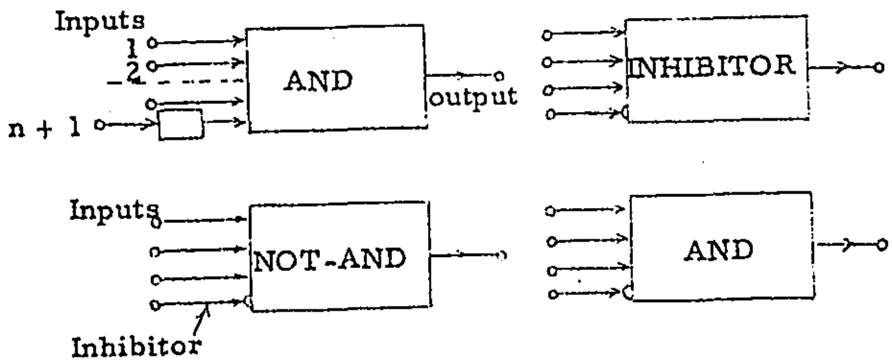


Fig. 3-19