ELECTRONIC COMMUNICATION

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Level
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Electronic Eng. / Communication Eng. Dep.

1. Communication Systems

All electronic communication systems have a transmitter, a communication channel or medium, and a receiver. These basic parts are shown in Fig. 1-1. The process of communication begins when a human being generates some kind of message, data, or other intelligence (information) that must be received by others.

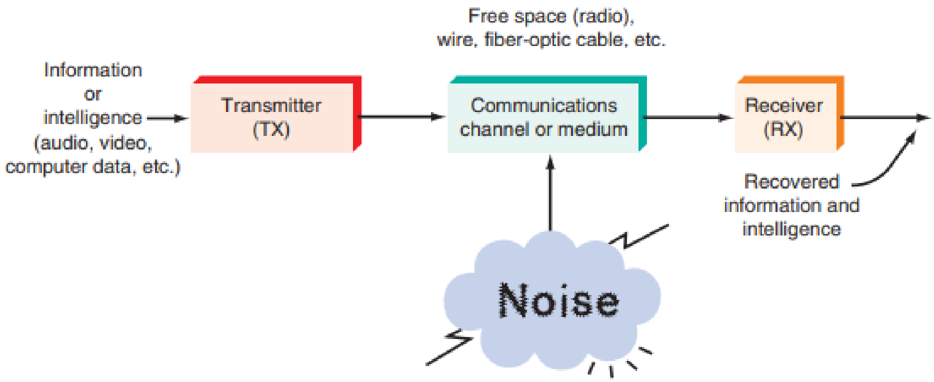


Fig. 1-1 A general model of all communication systems.

1.1. Signals

Information needs to be represented by signals. Signals are divided into two forms:

1. **Analog Signal**: Continuous variations of voltage and current. For instance, human voice is an analog signal.
2. **Digital Signal**: Discrete stepwise values such as 0 and 1. A human voice recorded on an mp3 becomes a digital signal.

1.2. Transducer

Transducer is a device that converts one form of energy into another form of energy.

The transducer, in electronic communication, is a device that converts one form of physical variables such as sound, video, temperature, pressure and force etc. into a corresponding electrical signal and vice versa. The simplest example of a transducer is the microphone of our phone, our voice in the form of sound waves is provided to the phone which in turn converts it into an electrical signal and transmits it.

1.3. Transmitter

The transmitter is an electronic circuit, designed by a collection of electronic components, to convert the electrical signal to a suitable form for transmission over a given communication channel (medium).

Transmitters are made up of oscillators, amplifiers, tuned circuits, filters, modulators, frequency mixers, frequency synthesizers, and etc.

1.4. Communication Channel

The communication channel is the medium by which the electronic signal is sent from one place to another. Many different types of medium are used in communication systems, including wire conductors, fiber-optic cable, and free space.

1.5. Receiver

The receiver is an electronic circuit, designed by a collection of electronic components, to receive the transmitted signals from the channel and converts it back to a form understandable by humans.

Receivers are made up of oscillators, amplifiers, tuned circuits, filters, demodulators, and etc.

1.6. Attenuation

Signal attenuation refers to the reduction in the strength of the analog or digital signal as it is transmitted over a communication medium or channel. Attenuation often occurs when signals are transmitted over long distances.

1.7. Amplitude

An amplitude of a signal refers to the strength of the signal.

1.8. Amplification

Sometimes when the distance between the sender of the signal and the receiver of the signal is too large, the amplitude of the signals drops significantly. To remedy the problem of weak signals, amplification of the signals is carried out to rejuvenate their strength.

Amplification is the process to strengthening the amplitude of the signals using an electronic amplifier circuit.

1.9. Bandwidth

Bandwidth describes the range of frequency over which a signal has been transmitted.

1.10. Modulation

Modulation refers to the act of adding information to an electronic or optical waveform. The information may be added by altering the frequency phase of the waveform, its amplitude or more.

1.11. Demodulation

Demodulation reverses modulation. It takes a modulated signal and extracts the original message out of it.

1.12. Noise

Any random electrical signal that interferes with the information signal is known as noise. It is the bane of all electronic communications. It can come from a variety of sources in the environment such as the rain, hailstorms or thunderstorms etc. Noise is always present in the system, it can be diminished but it can never be completely eliminated. In some instances, noise may even be generated by the receiver and hinder the demodulation process.

2. NOISE

Noise is an electronic signal that is a mixture of many random frequencies at many amplitudes that gets added to a radio or information signal as it is transmitted from one place to another or as it is processed. Noise is not the same as interference from other information signals.

White noise: The amplitude varies over a wide range, as does the frequency. One can say that noise in general contains all frequencies, varying randomly.

Fig. 2-2 White noise

2.1. Noise sources and types:

1- External Noise

External noise comes from sources over which we have little or no control— industrial, atmospheric, or space.

A- Industrial Noise
Produced by manufactured equipment, such as automotive ignition systems, electric motors, and generators. Any electrical equipment that causes high voltages or currents to be switched produces transients that create noise.

B- Atmospheric Noise
The electrical disturbances that occur naturally in the earth's atmosphere are another source of noise. Atmospheric noise is often referred to as static. Static usually comes from lightning, the electric discharges that occur between clouds or between the earth and clouds.

C- Extraterrestrial Noise

Extraterrestrial noise, solar and cosmic, comes from sources in space. One of the primary sources of extraterrestrial noise is the sun, which radiates a wide range of signals in a broad noise spectrum.

2- Internal Noise

Noise is generated in all equipment. Both passive components (like resistors and cables) and active devices (like diodes, and transistors) are major sources of internal noise. Internal noise, although it is low level, is often great enough to interfere with weak signals.

A- Thermal Noise

Most internal noise is caused by a phenomenon known as thermal agitation, the random motion of free electrons in a conductor caused by heat. The thermal noise is a white noise.

Noise power: The electric power that it is generate from the noise signal.

……………………………………(2.1)

Where:

P_N = noise power in watt.
k = Boltzman's constant (1.38×10^{-23} J/K)
T = absolute temperature in kelvins (K). This can be found by adding 273 to the Celsius temperature.
B = noise power bandwidth in Hz.

Example 2.1 A receiver has a noise power bandwidth of 10 kHz. A resistor that matches the receiver input impedance is connected across its antenna terminals. What is the noise power contributed by that resistor in the receiver bandwidth if the resistor has a temperature of 27 C ?

The noise power directly proportional to the T and B. The only way to reduce it is to decrease the temperature or the bandwidth.

Noise voltage: The voltage appears across a resistor.

Power in resistive circuit is given by the equation;

……………………………………(2.2)

From this

……………………………………(2.3)

The amount of open-circuit **noise voltage** appearing across a resistor or the input impedance to a receiver can be calculated according to Johnson's formula

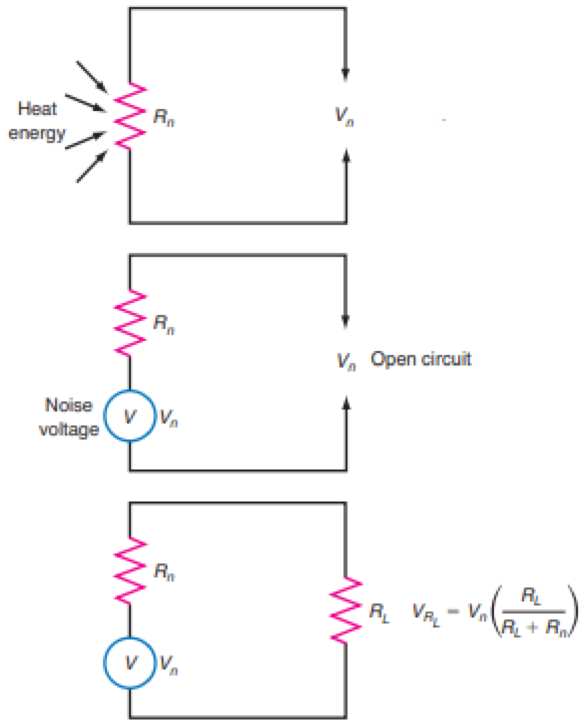


Fig. 3-1

……………………………………(2.4)

By using the noise power equation 2.1

……………………………………(2.5)

For the matching circuit $R_n = R_L$ and $V_L = V_n/2$ therefore;

……………………………………(2.6)

………………………………………(2.7)

Example 2.2 A 300ohm resistor is connected across the 300 ohm antenna input of a television receiver. The bandwidth of the receiver is 6 MHz and the resistor is at room temperature of 293 K (20°C or 68°F). Find the noise power and noise voltage applied to the receiver input.

Example 2.3 What is the open-circuit noise voltage across a 100k ohm resistor over the frequency range of direct current to 20 kHz at room temperature (25°C)?

Example 2.4 The bandwidth of a receiver with a 75ohm input resistance is 6 MHz. The temperature is 29°C. What is the input thermal noise voltage?

B- **SHOT NOISE**: shot noise is due to random variation in current flow in active devices such as transistor and diodes.

C- **Excess Noise:** is rarely a problem in communication circuits because it decline with increasing frequency and is usually insignificant above 1kHz

D- **High frequency effects:** Many junction devices cause a noise in the frequency approach their cutoff frequency.

A-1 Addition of noise from different source

If the system has more than single noise source in such a case, the total noise voltage or current can be found by the taking the square root of the sum of the squares of the individual voltages.

$$V_{nt} = \sqrt{V_{n1}^2 + V_{n2}^2 + V_{n3}^2 + \cdots} \text{......................................(3.7)}$$

EXAMPLE 3.5 A circuit with two resistors in series (100ohm and 200ohm), at two different temperatures (300k and 400k) respectively. Find the total noise voltage and noise power produced at the load, over a bandwidth of 100 kHz.

2.2. Signal to noise ratio

It is the ratio of Signal to Noise power that is important rather than the noise power alone. This signal-to-noise ratio (S/N) are usually expressed in decibels (dB).

Two different equations bellow are used for power ratio and voltage ratio respectively;

$$\left(\frac{S}{N}\right) dB = 10 \log \frac{P_S}{P_N} \text{......................................(2.8)}$$

$$\left(\frac{S}{N}\right) dB = 20 \log \frac{V_S}{V_N} \text{......................................(2.9)}$$

Although the signal-to-noise ratio is a fundamental characteristic of any communication system, it is often difficult to measure. For instance, it may be possible to measure the noise power by turning off the signal, but it is not possible to turn off the noise in order to measure the signal power alone. Consequently, a variant of S/N, called (S+N/N) is often found in receiver specifications. This stands for the ratio of signal-plus-noise power to noise power alone.

EXAMPLE 2.6 A receiver produces a noise power of 200 mW with no signal. The output level increases to 5 W when a signal is applied. Calculate (S+N)/N as a power ratio and in decibels.

2.3. Noise Figure

Noise figure (abbreviated NF or just F) is a figure of merit, indicating how much a component, stage, or series of stages degrades the signal-to-noise ratio of a system.

The noise figure is, by definition:

$$NF = \frac{(S/N)_i}{(S/N)_o} \quad \text{.......................................(2.10)}$$

where

(S/N)i = input signal-to-noise power ratio (not in dB)
(S/N)o = output signal-to-noise power ratio (not in dB)

Very often both S/N and NF are expressed in decibels, in which case we have:

$$NF(dB) = (S/N)_i\,(dB) - (S/N)_o\,(dB) \quad \text{.......................................(2.11)}$$

Since

$$NF(dB) = 10 \log NF \quad \text{.......................................(2.12)}$$

EXAMPLE 2.7 The signal power at the input to an amplifier is 100 uW and the noise power is 1 uW. At the output, the signal power is 1 W and the noise power is 30 mW. What is the amplifier noise figure, as a ratio?

EXAMPLE 2.8 The signal at the input of an amplifier has an S/N of 42 dB. If the amplifier has a noise figure of 6 dB, what is the S/N at the output (in decibels)?

3. Radio Frequency Amplifies

3.1 Narrow Band Amplifiers

1- RC coupling amplifier

Consider the bipolar common emitter amplifier shown in fig 3-1 below. It shows a conventional RC coupling amplifier.

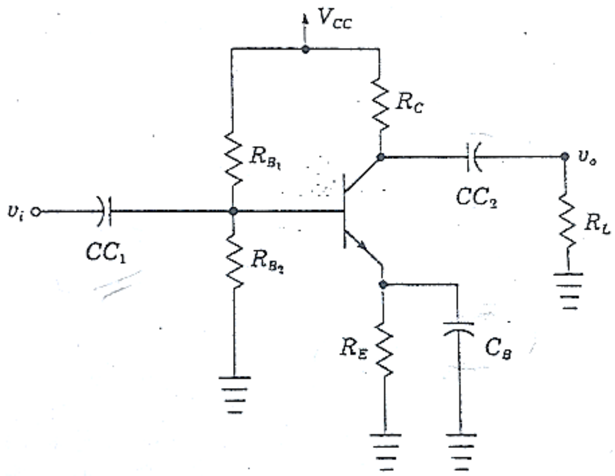


Fig. 3-1 Low frequency RC coupled amplifier

The voltage Gain is given approximately by;

$$Av = \frac{-(R_C || R_L)}{r_e} \quad (3.1)$$

Where

Av : Voltage Gain as ratio of Vo/Vi

RC||RL : Parallel combination of the collector and the load resistors

re : ac emitter resistance of the transistor in ohm

$$R_C || R_L = \frac{R_C * R_L}{R_C + R_L} \quad (3.1 - a)$$

The value of re depends on the bias current it is given by;

$$r_e = \frac{26mV}{I_E} \text{..(3.2)}$$

where
IE = dc emitter current in Amperes

$$I_E = \frac{V_E}{R_E} \text{..(3.2 - a)}$$

$$V_E = V_B - 0.7 \text{..(3.2 - b)}$$

$$V_E = V_{cc} * \frac{R_{B2}}{R_{B2}+R_{B1}} \text{..(3.2 - c)}$$

Example 3.1 : Calculate the Av of an low frequency RC coupled amplifier circuit similar to Fig. 3-1, with (R_{B1} = 56kohm, R_{B2} = 10kohm, R_E=1.2kohm, R_c=2kohm, R_L =300ohm and Vcc = 15V).

2- Tuned amplifier

The collector tuned circuit will be parallel resonant at a frequency that is given approximately by;

$$f_o = \frac{1}{2\pi\sqrt{L_1 C_1}} \quad \text{.......................................(3.3)}$$

Where

F_0 : resonant frequency in hertz

L1 : primary inductance in henrys

C1 : primary capacitance in farad

The gain is still equal to the ratio of collector-circuit impedance to emitter-circuit impedance, but the collector-circuit impedance is now very much a function of frequency.

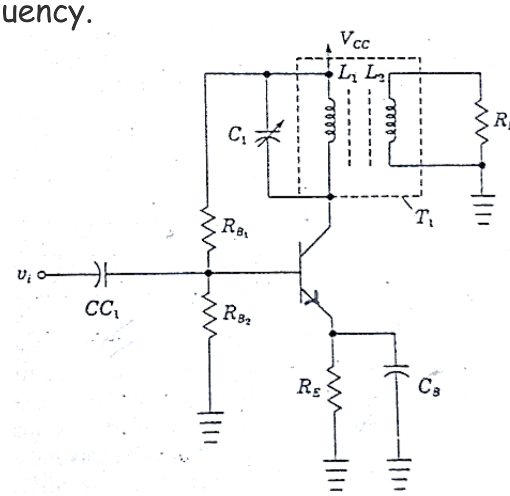


Fig. 3-2 High frequency transformer coupled tuned amplifier

$$R_L^{`} = \left(\frac{N_1}{N_2}\right)^2 R_L \quad \text{.......................................(3.4)}$$

RL : actual load resistance

$R_L^{`}$: equivalent load resistance

N1 : Number of turns in primary winding

N2 : Number of turns in secondary winding

$$A_{v_o}^{`} = \frac{R_L^{`}}{r_e} \quad \text{.......................................(3.5)}$$

$$A_{v_o}^{`} = \frac{R_L\, N_1^2}{r_e\, N_2^2} \quad \text{.......................................(3.6)}$$

The gain must include the voltage step up or step down effect of the transformer turns ratio.

$$A_{v_o} = \frac{A\grave{}_{v_o}\, N_2}{N_1} \quad \text{.......................................(3.7)}$$

$$A_{v_o} = \frac{R_L\, N_1^2}{r_e\, N_2^2}\,\frac{N_2}{N_1} \quad \text{.......................................(3.8)}$$

$$A_{v_o} = \frac{R_L}{r_e}\,\frac{N_1}{N_2} \quad \text{.......................................(3.9)}$$

$$Q = \frac{R\grave{}_L}{X_L} = \frac{R\grave{}_L}{W_0 L} \quad \text{.......................................(3.10)}$$

Where

Q: quality factor at resonance

$R\grave{}_L$: Load resistance transformed into the primary

X_L: reactance of the inductor at resonance

W_o: radian frequency at resonance ($2\pi f_0$)

and

$$B = \frac{f_o}{Q} \quad \text{.......................................(3.11)}$$

where

B: Bandwidth

F_0 : resonance frequency

3.2 Miller Effect

The Miller effect appears at high frequencies where the internal transistor capacitances are important. The internal capacitance C_{bc}, between the base and collector on the input, and the internal capacitance C_{be}, between the base and emitter, causes trouble by reduce the gain of the circuit at high frequencies. The effect of these capacitance called Miller effect.

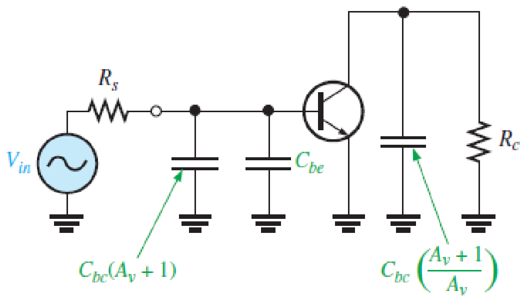


Fig. 3-3 Amplifier internal capacitances and effective Miller capacitances.

$$C_{in} = C_{bc}(A_{Vo} + 1) \qquad (3.11)$$

$$C_{out} = C_{bc}\left(\frac{A_{Vo}+1}{A_{Vo}}\right) \qquad (3.12)$$

Cin and Cout : Miller capacitance
C_{bc} : Capacitance between base and collector
A_{vo} : The gain of the amplifier ignoring the Miller effect.

Example 3.3 If C_{bc} 6 pF and the amplifier gain is 50, what is the Cin(Miller)?

Miller effect reduces by using;

1- Transformer coupling in the input as well as the output

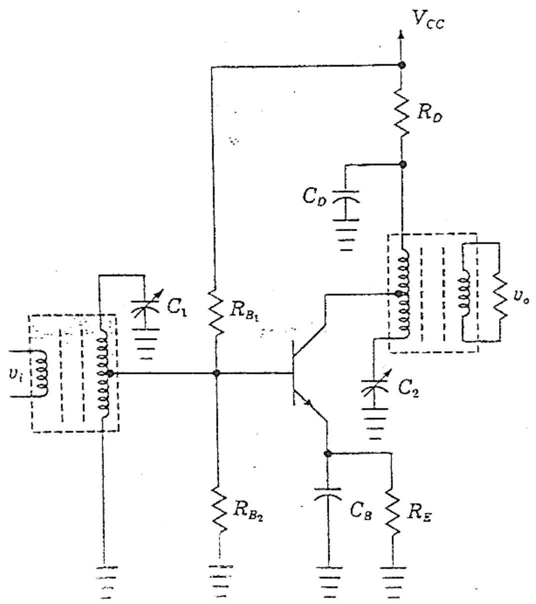


Fig. 3-4 HF transformer coupled input and output tuned amplifier

2- A common base amplifier

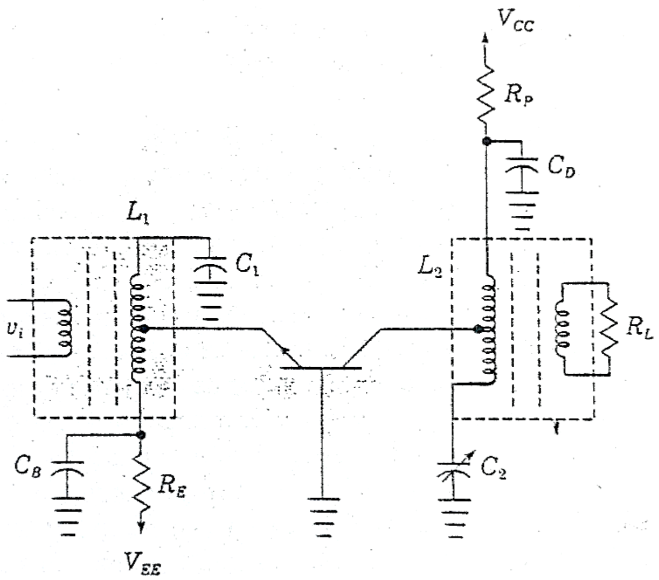


Fig. 3-5 Common base RF amplifier

3.3 RF OP-Amplifier

3.3 Wideband (Broadband) Amplifiers

The amplifiers designed for high frequency operation, where the response is required to extend over a relatively wide range of frequencies. For example, an amplifier for a cable-TV system might require to amplify frequencies from 50MHz to 400MHz.

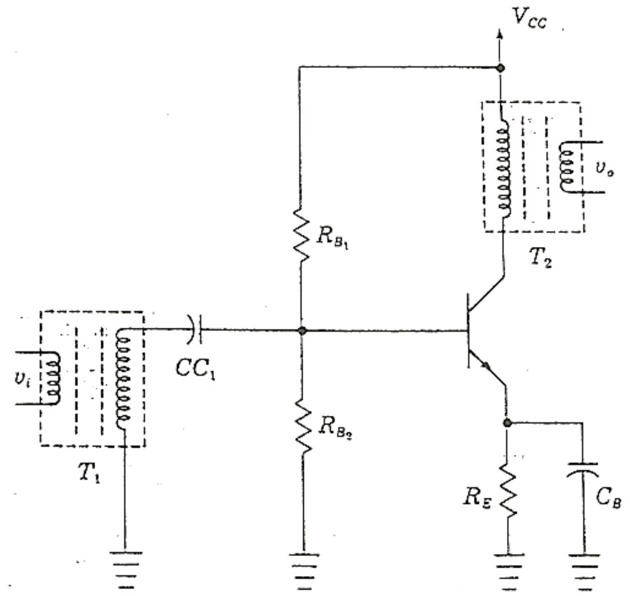


Fig. 3-6 Broadband RF amplifier

Wideband RF amplifiers, like their narrowband counterparts, typically use transformer coupling. For RF amplifiers they retain some advantages, however:

1- Transformers are very suitable for impedance matching.
2- The isolation between input and output is useful in helping to keep unwanted signals at frequencies greatly different from the desired signal frequency from propagating through the system.
3- Transformer coupling also makes it easy to couple balanced inputs or loads to the amplifier.

The transformers used in wideband amplifiers need careful design to avoid:
1- Self-resonance
2- To maintain relatively constant gain across the frequency range of interest.

Figure 3-6 is the circuit of a broadband RF amplifier using a bipolar transistor. Notice that it is the same as Figure 3-2, except that C1, the tuning capacitor, is gone, and the input as well as the output is transformer-coupled. Therefore the Av of the circuit is:

$$A_V = \frac{R_L}{r_e}\frac{N_1}{N_2} \quad \text{...(3.14)}$$

The difference is that the gain (A_{VO}) for this amplifier remains relatively constant over a wide frequency range. Also, the gain must be multiplied by the turn's ratio of the input transformer.

Example 3.4 An amplifier has the circuit of Figure 3-7 Calculate its voltage gain Av, assuming ideal transformers and ignoring any loading by the transistor.

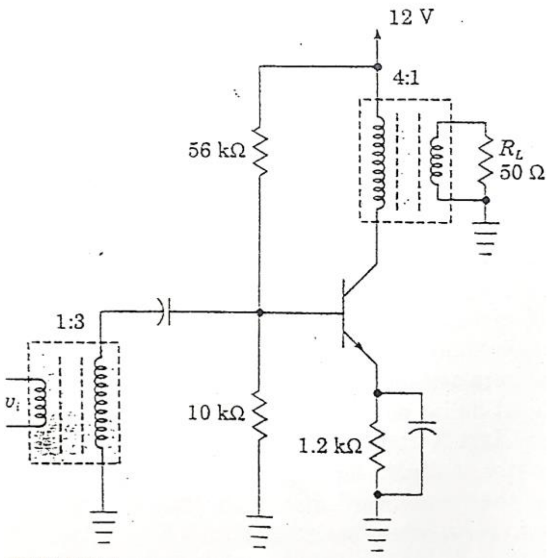


Fig. 3-7

3.4 Amplifier Classes

Amplifiers are classified according to the portion of the input cycle during (**when excited by a sinusoidal input signal**) which the active device conducting current. This called the conduction angle and it is expressed in degree.

Not all amplifiers are the same and there is a clear distinction made between the way their output stages are configured and operate. The main operating characteristics of an ideal amplifier are linearity, signal gain, efficiency and power output but in real world amplifiers there is always a tradeoff between these different characteristics.

1- Class A amplifier

When an amplifier is biased such that the transistor conducts current at all time (always operates) for a conduction angle 360 in the leaner region, where the output signal is an amplified replica of the input signal.

Features of class A amplifiers:

1- The most common type of amplifier topology.
2- Use just one output switching transistor within their amplifier design.
3- The linearity, this single output transistor is biased around the Q-point within the middle of its load line and so is never driven into its cut-off or saturation regions thus allowing it to conduct current over the full 360 degrees of the input cycle.
4- Then the output transistor of a class-A topology never turns “OFF” which is one of its main disadvantages.

Advantages of Class A:

1- Linear
2- Less distortion
3- Quite stable

Disadvantages of Class A:

1- The maximum efficiency of a class A is thus to 25%.

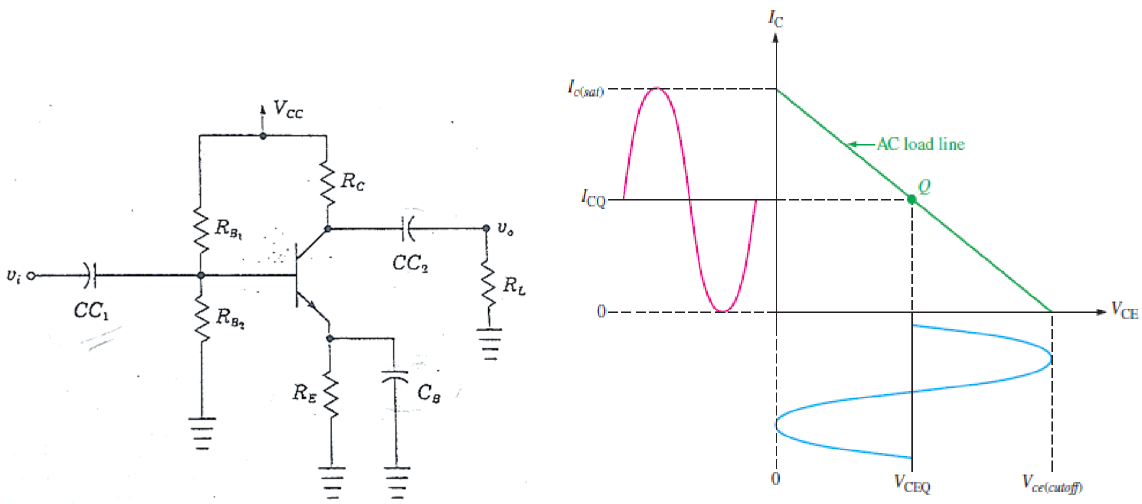


Fig. 3-8 Class A amplifier circuit and its load line curve

2- Class B amplifier

When an amplifier is biased at cutoff (I_{CQ}=0 and V_{CEQ}=$V_{CEcutoff}$) so that it operates in the linear region of the input cycle at cutoff point for 180° conduction degree.

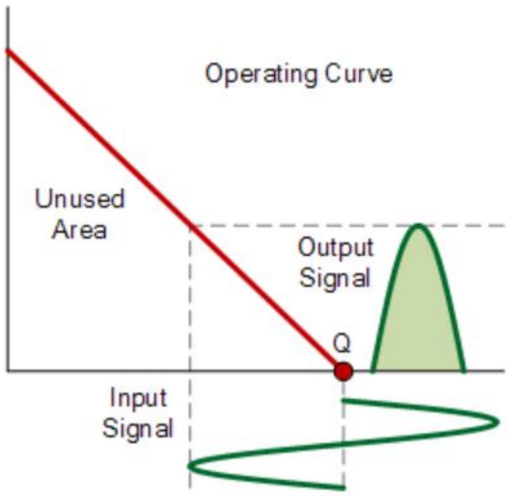


Fig. 3-9 Class B amplifier load line curve

The basic class B amplifier uses two complimentary transistors for each half of the waveform with its output stage configured in a "push-pull" type arrangement, so that each transistor device amplifies only half of the output waveform.

There are two common approaches for using push-pull amplifiers to reproduce the entire waveform. The first approach uses transformer coupling. The second uses two complementary symmetry transistors; these are a matching pair of npn/pnp:

A- Complementary Symmetry Transistors

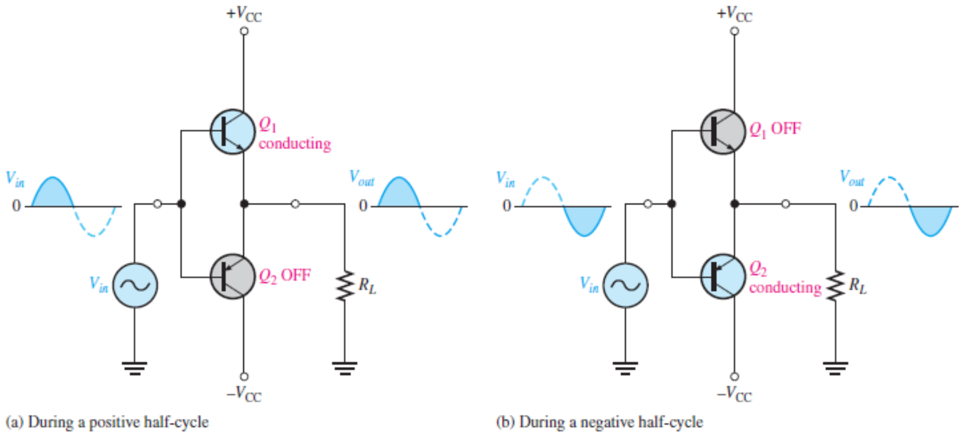


Fig. 3-10 Class B amplifier complementary symmetry transistors circuit

B- Transformer Coupling

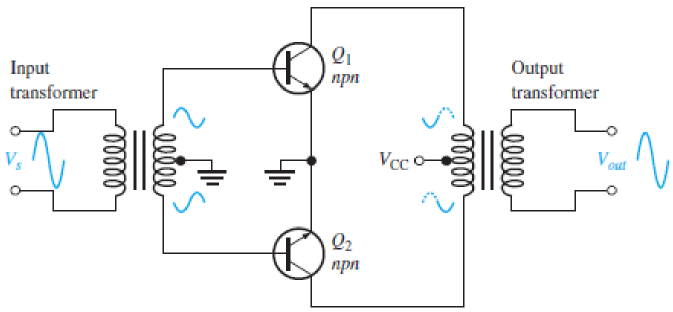


Fig. 3-11 Class B amplifier transformer coupling circuit

Advantages of Class B:
More efficient than Class A, at about 50%.

Disadvantage of Class B:
Its design creates distortion at the zero-crossing point of the waveform due to the transistors dead band of input base voltages from -0.7V to +0.7 and it is called **Crossover Distortion.**

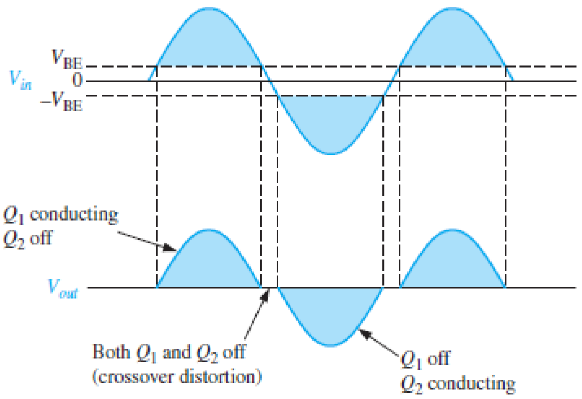


Fig. 3-11 Illustration of crossover distortion in a class B push-pull amplifier

3- Class AB amplifier

The class AB amplifier is a variation of a class B amplifier as described above, except that both transistors are allowed to conduct at the waveforms crossover point (more than cutoff point degree 180°) eliminating the crossover distortion problems of the previous class B amplifier.

To overcome crossover distortion, the biasing is adjusted to just overcome the V_{BE} of the transistors. In class AB operation, the push-pull stages are biased into slight conduction, even when no input signal is present.

This can be done with a voltage-divider and diode arrangement, as shown in Figure 3–12. When the diode characteristics of D1 and D2 are closely matched to the characteristics of the transistor base-emitter junctions, the current in the diodes and the current in the transistors are the same; this is called a current mirror. This current mirror produces the desired class AB operation and eliminates crossover distortion.

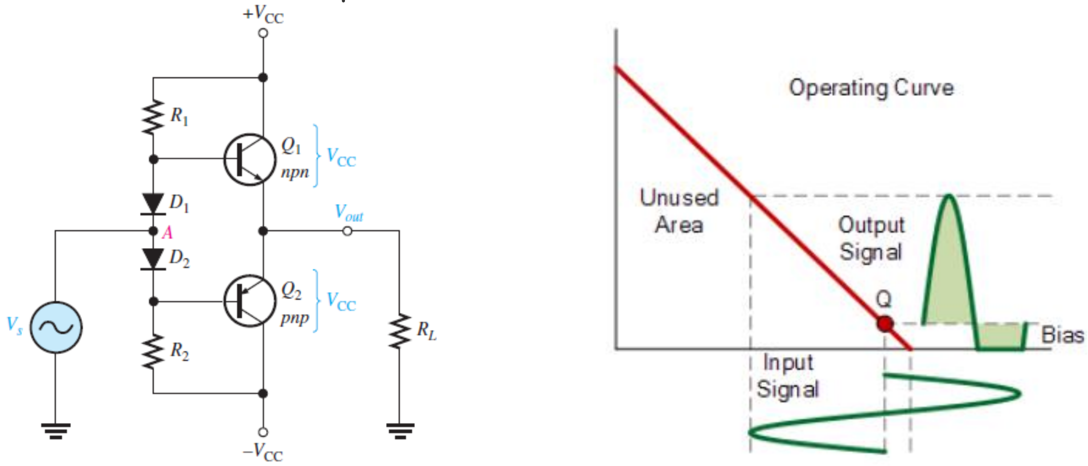


Fig. 3-12 Class AB amplifier as complementary symmetry transistors (Biasing the push-pull amplifier with current mirror diode) and load line curve.

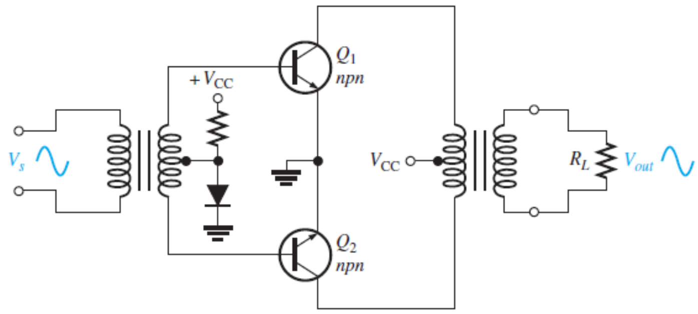


Fig. 3-13 Class AB amplifier eliminating crossover distortion in a transformer-coupled push-pull amplifier.

The diode current and I_{CQ} can be found by applying Ohm's law to either R1 or R2 as follows:

$$I_{CQ} = \frac{V_{CC} - 0.7}{R_1} \quad \text{......................................(3.15)}$$

AC Operation:
Consider the ac load line for Q1 of the class AB amplifier in Figure 3–12. The Q-point is slightly above cutoff. (In a true class B amplifier, the Q-point is at cutoff.) The ac cutoff voltage for a two-supply operation is at V_{CC} with an I_{CQ} as given earlier. The ac saturation current for a two-supply operation with a push-pull amplifier is The ac load line for the npn transistor is as shown in Figure 3–12. The dc load line can be found by drawing a line that passes through V_{CEQ} and the dc saturation current, $I_{C(sat)}$.

$$I_{C(sat)} = \frac{V_{CC}}{R_L} \quad \text{......................................(3.16)}$$

Example 3.5 Determine the ideal maximum peak output voltage and current for the circuit shown in Figure 3-12. When R1 = R2 = 430ohm, RL= 150ohm and Vcc=20 volt.

Example 3.6 What is the maximum peak output voltage and current if the supply voltages are changed to +15 V and -15 V. Analyze the circuit using the Multisim Software.

Single-Supply Push-Pull Amplifier:
Push-pull amplifiers using complementary symmetry transistors can be operated from a single voltage source as shown in Figure 3–14. The circuit operation is the same as that described previously, except the bias is set to force the output emitter voltage to be $V_{CC}/2$ instead of zero volts used with two supplies. Because the output is not biased at zero volts, capacitive coupling for the input and output is necessary to block the bias voltage from the source and the load resistor. Ideally, the output voltage can swing from zero to V_{CC}, but in practice it does not quite reach these ideal values.

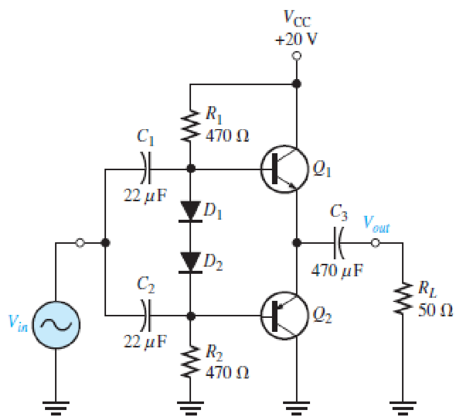


Fig. 3-14 Class AB amplifier with single supply push-pull amplifier.

The Power of Class B/AB:

Maximum Output Power you have seen that the ideal maximum peak output current for both dual-supply and single-supply push-pull amplifiers is approximately Ic(sat), and the maximum peak output voltage is approximately V_{CEQ}. Ideally, the maximum average output power is, therefore,

$$P_{out} = I_{out(rms)} V_{out(rms)} \qquad (3.16)$$

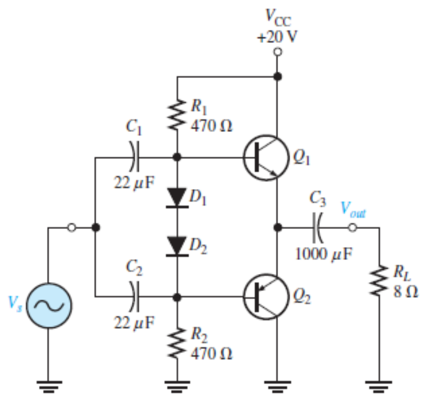
$$I_{out(rms)} = 0.707\, I_{out(peak)} = 0.707\, I_{C(sat)} \qquad (3.17)$$

Since each transistor draws current for a half-cycle, the current is a half-wave signal with an average value of so,

$$I_{CC} = \frac{I_{C(sat)}}{\pi} \quad \text{.......................................(3.21)}$$

Efficiency An advantage of push-pull class B and class AB amplifiers over class A is a much higher efficiency. This advantage usually overrides the difficulty of biasing the class AB push-pull amplifier to eliminate crossover distortion. Recall that efficiency, is defined as the ratio of ac output power to dc input power.

Example 3.7 Find the maximum ac output power and the dc input power of the amplifier in the following circuit.



Advantages of Class AB:

1- The class AB has linear behavior
2- The design of this amplifier is very simple

Disadvantages of Class AB:

The class AB amplifier is a good compromise between class A and class B in terms of efficiency, with conversion efficiencies reaching about 80%

4- Class C amplifier

The class C amplifier designed to has a great efficiency and poor linearity. In other words, the conduction angle for the transistor is significantly less than 180 degrees, and is generally around the 90 degrees area.

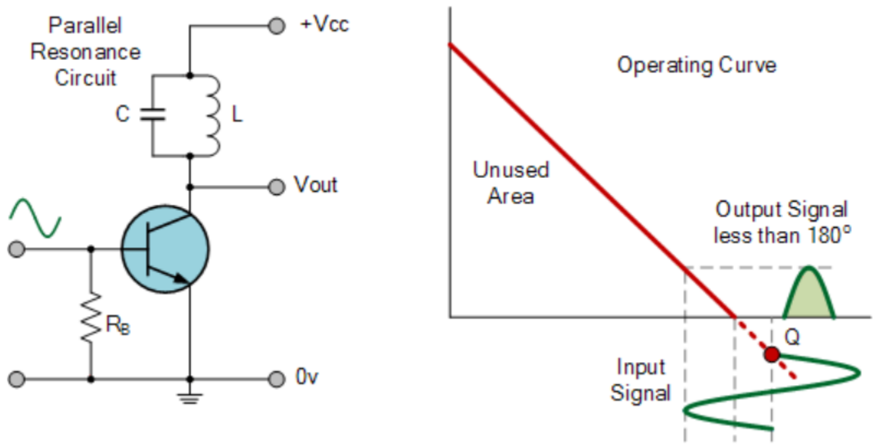


Fig. 3-15 Class C amplifier and its load line curve

Power Dissipation

The power dissipation of the transistor in a class C amplifier is low because it is on for only a small percentage of the input cycle.

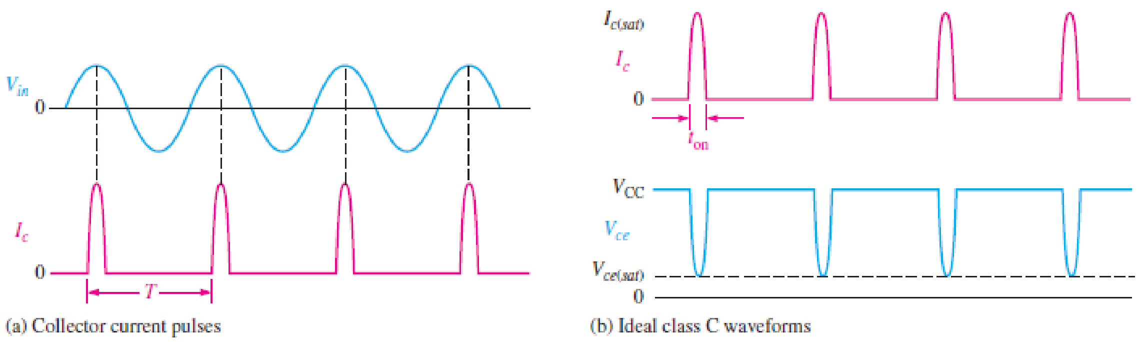


Fig. 3-16 Class C amplifier collector current and the waveform

The transistor is On for a short time, t_{on}, and Off for the rest of the input cycle. Therefore, assuming the entire load line is used, the power dissipation averaged over the entire cycle is:

$$P_{D(avg)} = (\frac{t_{on}}{T}) I_{C(sat)} V_{CE(sat)} \quad \text{.......................................(3.19)}$$

Example 3.8 A class C amplifier is driven by a 200 kHz signal. The transistor is On for 1 us, and the amplifier is operating over 100 percent of its load line. If $I_{C(sat)}$ is 100 mA and $V_{CE\ (sat)}$ is 0.2 V, what is the average power dissipation of the transistor?

Maximum Output Power

Since the voltage developed across the tank circuit has a peak-to-peak value of approximately $2V_{CC}$, the maximum output power can be expressed as

$$P_{out} = \frac{(0.707 V_{CC})^2}{R_C} \quad \text{.......................................(3.20)}$$

R_C is the equivalent parallel resistance of the collector tank circuit at resonance and represents the parallel combination of the coil resistance and the load resistance. It usually has a low value. The total power that must be supplied to the amplifier is:

$$P_T = P_{out} + P_{D(avg)} \quad \text{.......................................(3.21)}$$

Therefore, the efficiency is:

$$\eta = \frac{P_{out}}{P_{out} + P_{D(avg)}} \quad \text{.......................................(3.22)}$$

When $P_{Out} >> P_{D(avg)}$,the class C efficiency closely approaches 1 (100 percent).

Example 3.9 Suppose the class C amplifier described in example 4-8 has a V_{CC} equal to 24V and the R_C is 100 ohm determine the efficiency.

Advantages of Class C:

The efficiency of class C amplifier is high 85% to 90%

Disadvantages of Class C:

The linearity of class C amplifier is low. It introduces a very heavy distortion of the output signal. Therefore, class C amplifiers are not suitable for use as audio amplifiers.

4. RF Oscillators

An **Oscillator** is a circuit which produces a continuous, repeated, alternating waveform without any input. Oscillators basically convert unidirectional current flow from a DC source into an alternating waveform which is of the desired frequency, as decided by its circuit components.

Any amplifier can be made to oscillate if a portion of the output is feedback to the input according the Barkhausen criteria. The figure bellow shows a generic oscillator consist of an amplifier with gain **A** and a feedback with **B** (B will be less than 1, only a part of output will be fed back to the input).

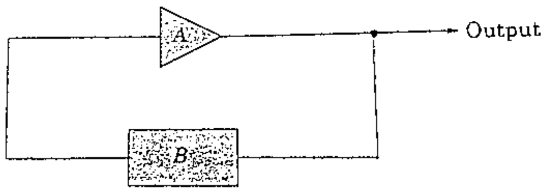


Fig. 4-1 Typical oscillator

According to the Barkhausen criteria, when oscillations are in progress, we will have:

$$AB = 1 \quad \text{...(4.1)}$$

4.1. LC Oscillators

The oscillators whose frequency is controlled by a resonant circuit using inductance and capacitance.

1- Hartley Oscillator

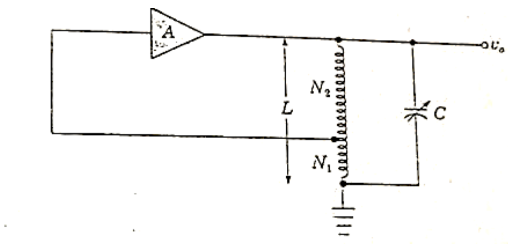
This oscillator type can be recognized by its use of a tapped inductor, part of a resonant circuit, to provide feedback. Fig. 4-2(a) shows a Hartley oscillator using noninverting amplifier and Fig. 4-2(b) shows the same Oscillator using an inverting amplifier.

For the non-inverting circuit of Fig. 4-2(a):

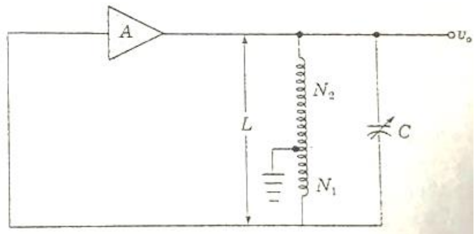
$$B = \frac{N_1}{N_1+N_2} \quad \text{...(4.2)}$$

For the inverting circuit of Fig. 4-2(b):

$$B = \frac{-N_1}{N_2} \quad \text{...(4.3)}$$



(a) Using Noninverting Amplifier



(b) Using Inverting Amplifier

Fig. 4-2 Hartly Oscillators

Example 4.1: A Hartley oscillator is shown in Figure 5-3 uses a junction field - effect transistor (JFET) in the common-source configuration. Determine whether it will oscillate, and if so, calculate the operating frequency.

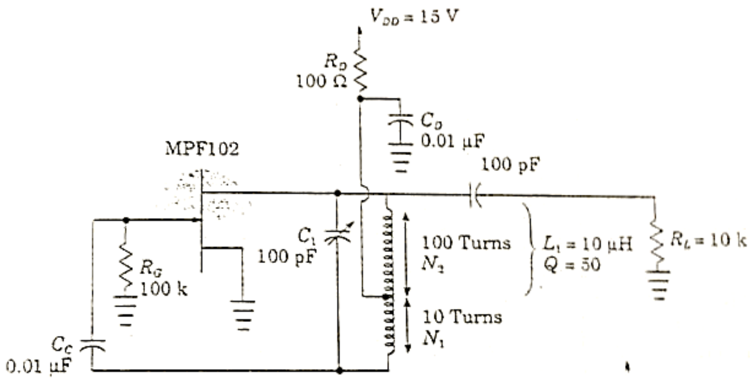


Fig. 4-3

2- Colpitts Oscillator

The Colpitts oscillator uses a capacitive voltage divider. Once again, the feedback depends on whether the amplifier is non-inverting, as in Fig. 4-4(a), or inverting, as in Fig. 4-4 (b).

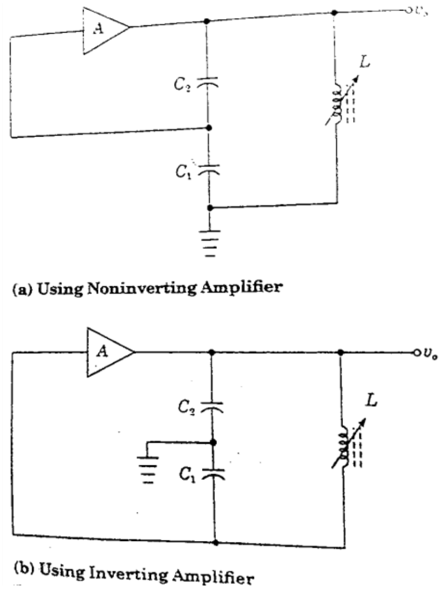


Fig. 4-4 Colpitts Oscillators

The operating frequency is determined by the inductor and the series combination of C1 and C2.

$$f_o = \frac{1}{2\pi\sqrt{LC_T}} \quad \text{.......................................(4.4)}$$

Where

$$C_T = \frac{C_1 C_2}{C_1 + C_2} \quad \text{.......................................(4.5)}$$

The feedback fraction for the non-inverting version of the Oscillator, the output is across the series C_1 and C_2 , which corresponds to C_T in Eq.(4.4), and the input is the voltage across C_1. It is easy to see that the feedback fraction is;

$$B = \frac{X_{C1}}{X_{CT}} = \frac{\frac{1}{2\pi f C_1}}{\frac{1}{2\pi f C_T}} \quad \text{.......................................(4.6)}$$

$$= \frac{C_T}{C_1} \quad \text{.......................................(4.6a)}$$

$$= \frac{\frac{C_1 C_2}{C_1+C_2}}{C_1} \quad \text{..(4.6b)}$$

$$B = \frac{C_2}{C_1+C_2} \quad \text{..(4.7)}$$

The feedback fraction is even easier to determine for the inverting circuit of Fig 4-4b. Since the output is applied across C_2, and the input is taken across C_1, the feedback fraction is;

$$B = -\frac{X_{C1}}{X_{C2}} \quad \text{..(4.8)}$$

$$B = -\frac{C_2}{C_1} \quad \text{..(4.9)}$$

Example 4.2: Determine the feedback fractions and operating frequencies for the oscillators whose circuits are shown in Fig. 4-5.

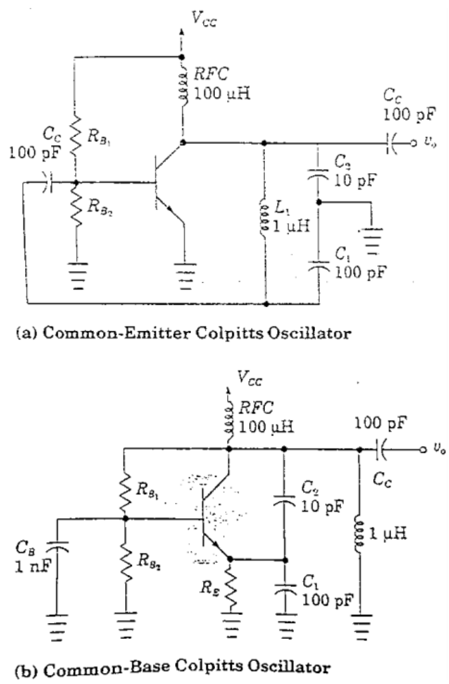


Fig. 4-5

3- Clapp Oscillator

In the Clapp oscillator, the frequency of oscillation is determined by the inductor and the series combination of C_1, C_2 and C_3.

In practice, the total effective capacitance of the three capacitors in series is given by;

$$C_T = \frac{1}{\frac{1}{C_1}+\frac{1}{C_2}+\frac{1}{C_3}} \quad \text{..(4.10)}$$

The operating frequency can easily be found by using the Eq (4.4) as for the Colpitts oscillator.

The feedback fraction of the non-inverting and inverting circuit is found in the same way as for Colpitts oscillator, since Clapp oscillator is so similar to the Colpitts.

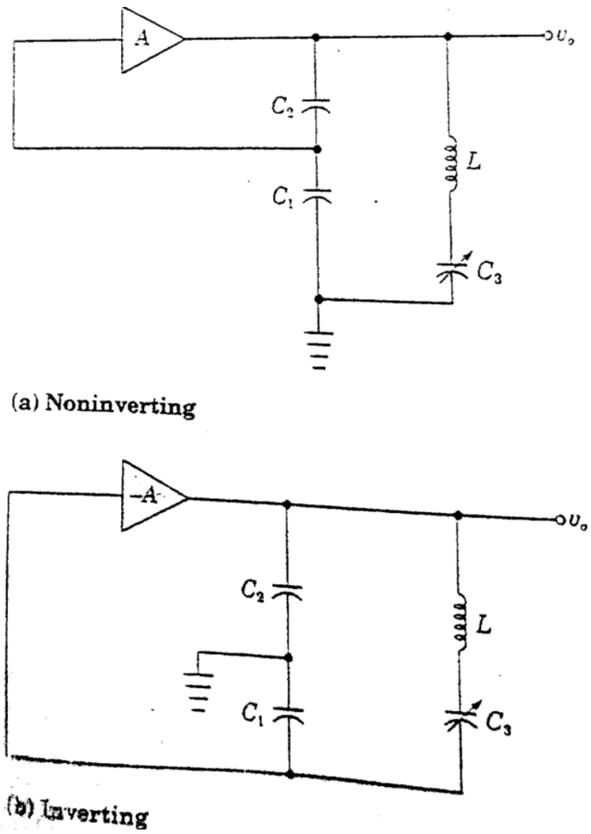


Fig. 4-6 Clapp Oscillators

Example 4.3: Calculate the feedback fraction and oscillation frequency for the circuit Fig. 4-7.

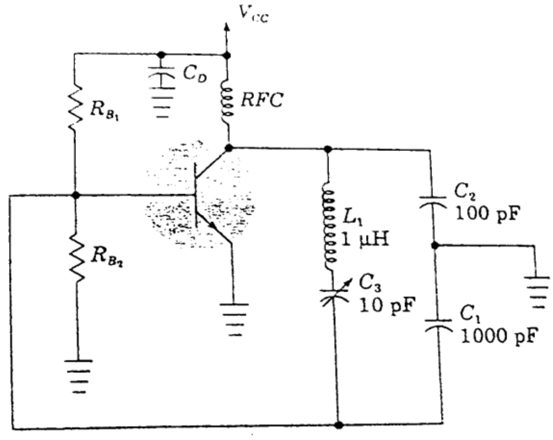


Fig. 4-6

4.2. RC Oscillator

The oscillators whose frequency is controlled by a resonant circuit using resistor and capacitance.

Wein Bridge Oscillator

The Wein bridge oscillator is the most widely type of RC feedback osc. to generate signals for frequencies 1Hz to 1MHz.

The Wein bridge oscillator uses the OP-AMP circuit with lead-lag network as shown in Fig.4-8.

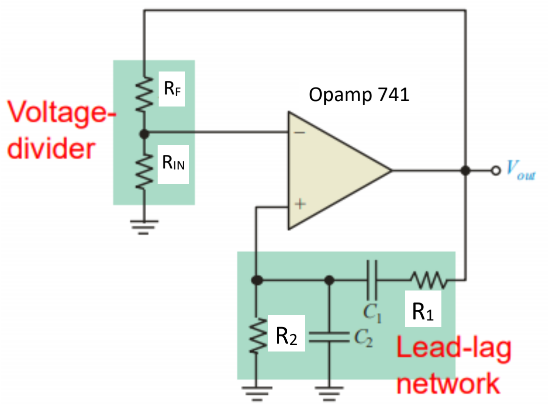


Fig.4-8 Wein Bridge

The basic Wien Bridge uses the lead-lag network as shown in Fig.4-8 to get feedback fraction and select a specific frequency that is amplified.

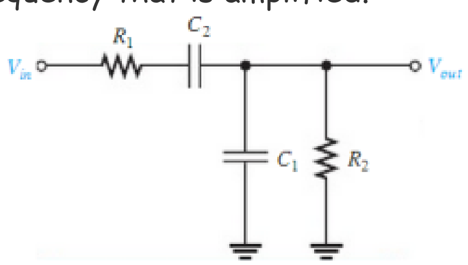


Fig.4-9 Lead-Lag circuit

The lead-lag circuit gives 0 phase shift and 1/3 (Vout/Vin) attenuation at resonant frequency.

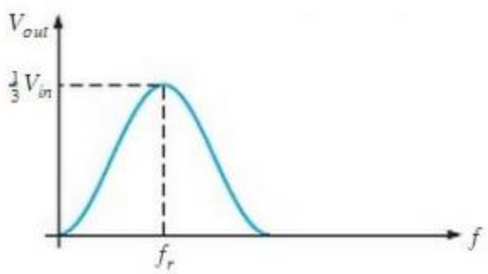


Fig.4-10 Lead-Lag circuit output attenuation at the resonant frequency

To calculate the feedback fraction and the resonant frequency f_0;

$$Z_1 = R_1 - j\frac{1}{WC_1} \qquad \text{...(4.11)}$$

$$Z_2 = \frac{R_2 * \frac{-j}{WC_2}}{R_2 - j\frac{1}{WC_2}}$$

In order Vout to have the same phase of the Vin in the lead-lag circuit. The feedback fraction or ratio must be purely real value, the imaginary part must be **Zero.**

From Eq. (4.14) if the $R_1 = R_2 = R$ and $C_1 = C_2 = C$ then;

The feedback fraction is;

The resonant frequency f_0 is ;

$WR_1 R_2 C_2 - 1/W C_1 = 0$ The imaginary part

For Op-Amplifier

The non-inverting amplifier must have a gain of exactly 3.0 as set by R_F and R_{IN} of OP-Amp to make the magnitude of the loop gain unity (AB = 1).

$$A = 1 + \frac{R_F}{R_{IN}} \quad \text{.......................................(4.16)}$$

$$1 + \frac{R_F}{R_{IN}} = 3$$

The Positive feedback condition for oscillator choose;

$$R_F = 2R_{IN} \quad \text{.......................................(4.17)}$$

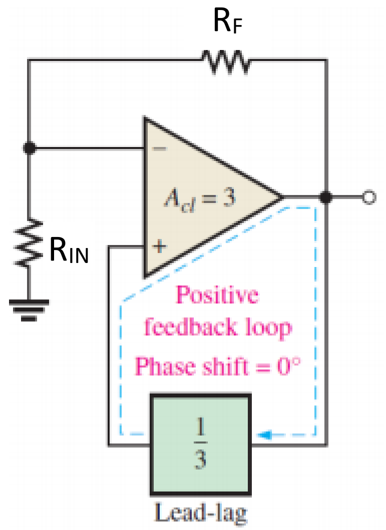


Fig.4-11 The phase shift is 0 and the loop gain is 1 around the loop.

Example 4.4: Calculate the oscillation frequency for the Wien bridge osc. in Fig. 4-12.

Example 4.5: Design a Wein bridge osc. as in Fig. 4-12 for 10KHz operation frequency.

4.3. Voltage Control Oscillator (VCO)

Instead of tuning the osc. circuits for resonant frequency by using variable capacitors or inductors, that it is mechanical bulk component, we can use voltage control tuning way to automatic frequency control.

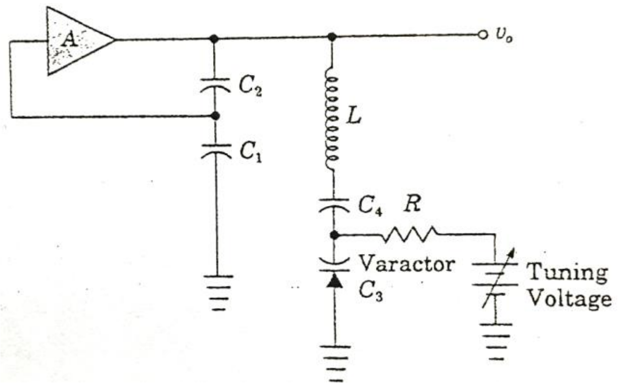


Fig.4-13 Voltage control oscillator circuit

The non-inverting Clapp oscillator has been adapted for use as a VCO by using a Varactor with junction capacitance.

Varactor is a reverse-biased Silicon diode.

When the reverse voltage of Varactor increases, so does the width of the diode's depletion layer, then the junction capacitance decrease.

The variation of capacitance with voltage is not linear for a varactor. It is given approximately by;

$$C = \frac{C_0}{\sqrt{1+2V}} \quad \text{...(4.18)}$$

Where C = capacitance at reverse voltage V
C_0 = capacitance with no reverse voltage (Zero)

In practice, a variation of about 5:1 in capacitance is quite practical, so;

$$\frac{C_0}{C} = 5 \quad \text{...(4.19)}$$

From Eq.(4.18)

$$\frac{C_0}{C} = \sqrt{1 + 2V}$$

By substitute Eq.(4.19)

$$5 = \sqrt{1 + 2V}$$

$$V = 12 \text{ volt} \quad \text{............................ (4.20)}$$

Example 4.6: A varactor has a maximum capacitance of 80 pF and is used in a tuned circuit with a 100 μH inductor.
(a) Find the resonant frequency with no tuning voltage applied.
(b) Find the tuning voltage necessary for the circuit to resonate at double the frequency found in Part (a).

4.4. Crystal Controlled Oscillator

Crystal oscillator achieve greater stability using a quartz as a mechanical resonance of a vibrating crystal of piezoelectric material, in place of an LC tuned circuit. The crystal osc. can be model using the following circuit;

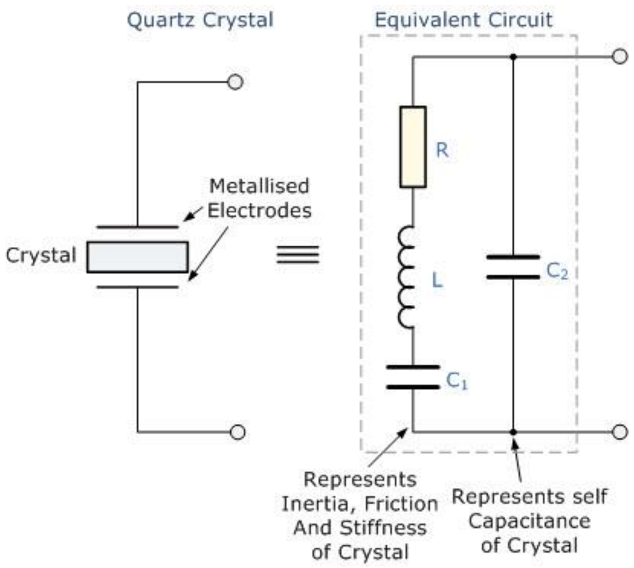


Fig.4-14 Crystal oscillator equivalent circuit (modeling circuit)

The equivalent impedance of the crystal has a series resonance where Cs resonates with inductance Ls at the crystal operating frequency. This frequency is called crystal series frequency fs. There is a second frequency point as a result of parallel resonance resonates with the parallel capacitor Cp.

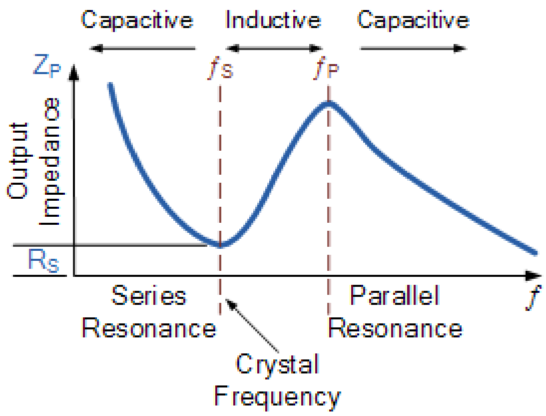


Fig.4-15 Crystal oscillator output impedance and resonant frequencies

The mathematical representation of the crystal osc. equivalent circuit is;

$$R_s = R \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (4.21)$$

$$X_{Ls} = 2\pi\ f\ L_s \text{ ……………………… (4.22)}$$

$$X_{cs} = \frac{1}{2\pi\ f\ C_s} \text{ ……………………… (4.23)}$$

$$X_{cp} = \frac{1}{2\pi\ f\ C_p} \text{ ……………………… (4.24)}$$

$$Z_s = \sqrt{R_S^2 + (X_{Ls} - X_{cs})^2} \text{ ……………………… (4.25)}$$

$$Z_P = \frac{Z_S\ X_{Cp}}{Z_S + X_{Cp}} \text{ ……………………… (4.26)}$$

Series resonant frequency;

$$f_s = \frac{1}{2\pi\ \sqrt{L_S C_S}} \text{ ……………………… (4.27)}$$

Parallel resonant frequency;

$$f_P = \frac{1}{2\pi\ \sqrt{L_S\ (\frac{C_P\ C_S}{C_P + C_S})}} \text{ ……………………… (4.28)}$$

Both capacitors are very small (Pico-farad or less than), with the parallel capacitance Cp much larger than the series Cs, and the resistor Rs is small therefore the Q factor of the crystal is very high.

$$Q = \frac{X_{Ls}}{R_s} \text{ ……………………… (4.26)}$$

Example 4.7 A quartz crystal osc. has the following values Rs = 6.4 ohm, Cs = 0.09972pF, Ls = 2.546mH and Cp = 28.62pF. Calculate the oscillating resonant frequencies of the crystal and the Q factor.

Example 4.8 A quartz crystal has the following values Rs = 1 Kohm, Cs = 0.05pF, Ls = 3H and Cp = 10pF. Calculate the series and parallel oscillating frequencies.

Some types of Crystal oscillator circuit;

1- Colpitts Crystal Oscillator

This type of Crystal Oscillators are designed around a common collector (emitter-follower) amplifier as shown in fig 4-16.

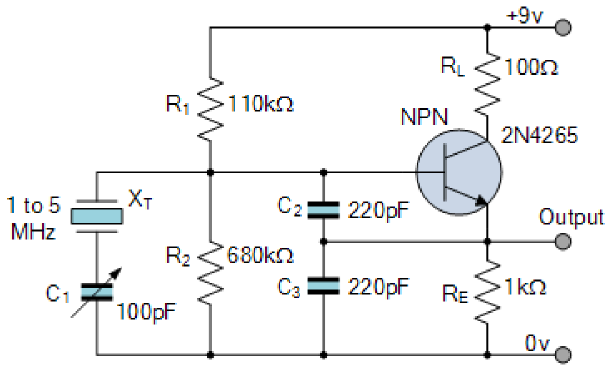


Fig. 4-16 Colpitts Crystal oscillator circuit

- The R1 and R2 resistor network sets the DC bias level on the base.
- Emitter resistor RE sets the output voltage level.
- Resistor R2 is set as large as possible to prevent loading to the parallel connected crystal.
- The capacitors, C1 and C2 shunt the output of the transistor which reduces the feedback signal. Therefore, the gain of the transistor limits the maximum values of C1 and C2.

The output amplitude should be kept low in order to avoid excessive power dissipation in the crystal otherwise could destroy itself by excessive vibration.

2- Pierce Oscillator

The Pierce oscillator is primarily a series resonant tuned circuit as shown in Fig. 14-17.

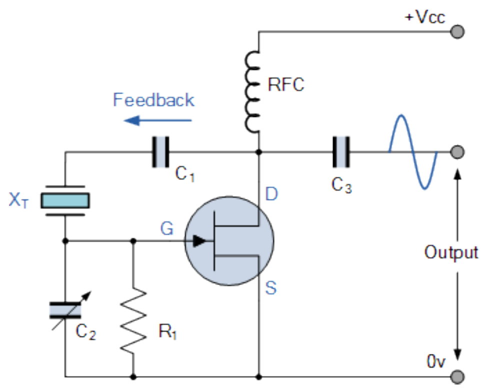


Fig. 4-17 Pierce Crystal oscillator circuit

- It uses a JFET for its main amplifying device as FET's provide very high input impedances with the crystal connected between the Drain and Gate via capacitor C1.

- The crystal determines the frequency of oscillations and operates at its series resonant frequency, *ƒs* giving a low impedance path between the output and the input.

- Resistor, R1 controls the amount of feedback and crystal drive while the voltage across the radio frequency choke, RFC reverses during each cycle.

Most digital clocks, watches and timers use a Pierce Oscillator in some form or other as it can be implemented using the minimum of components.

3- Microprocessor Crystal Oscillator

Most microprocessors CPU and micro-controllers MC have two oscillator pins labelled OSC1 and OSC2 to connect to an external quartz crystal circuit as shown in Fig. 4-18.

In this type of microprocessor application the **Quartz Crystal Oscillator** produces a train of continuous square wave pulses whose fundamental frequency is controlled by the crystal itself.

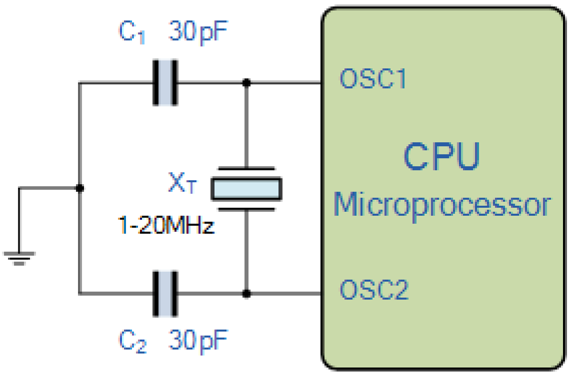


Fig. 4-18 Microprocessor Crystal oscillator circuit

This fundamental frequency regulates the flow of instructions that controls the processor device. For example, the master clock and system timing.

Report 4-1 IC 555 chip and using it as oscillator circuit to generate a binary output signal and a triangle signal.

5. Passive and active Filters

A filter is an electronic circuit capable of passing signals in a certain frequencies while attenuating other frequencies. Thus, a filter can extract important frequencies from signals that also contain undesirable or irrelevant frequencies.

In the field of electronics, there are many practical applications for filters. Examples include:

- Radio communications
- DC power supplies
- Audio electronics
- Analog-to-digital ADC convertors

There are four primary types of filters include:

1- Low-pass filter LPF
2- High-pass filter HPF
3- Band-pass filter BPF
4- Notch filter (or the band-reject or band-stop filter).

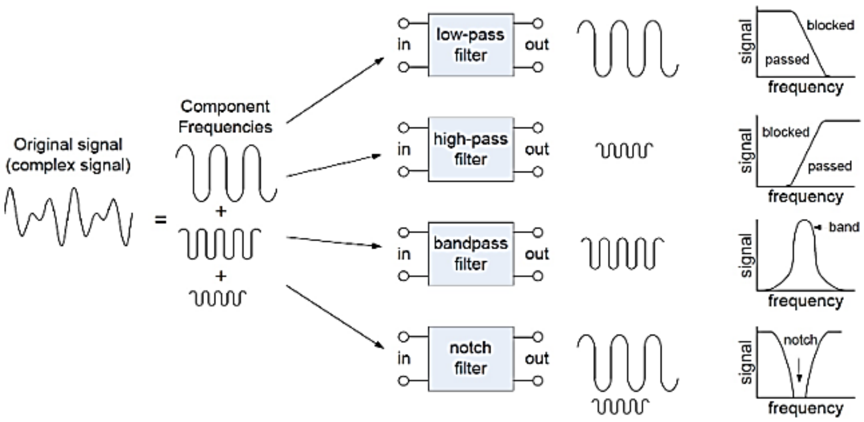


Fig. 5-1 Filter types

Filters can be placed in one of two categories:

1- Passive filters
2- Active filters

Passive filters include only *passive components*—**resistors, capacitors, and inductors**. In contrast, active filters use *active components*, such as **op-amps, in addition to resistors and capacitors, but not inductors.**

5.1. Useful terms in filters

- ***Response curve*** is simply a graph showing an attenuation ratio (V_{OUT} / V_{IN}) versus frequency of a filter then describe how a filter behaves as shown in Fig. 5-2.

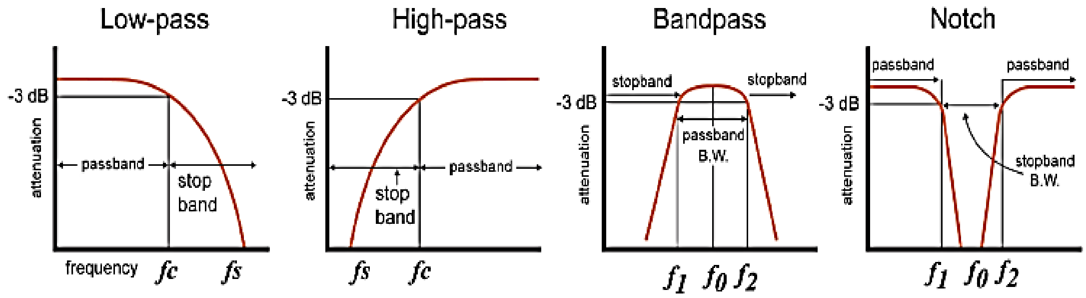


Fig. 5-2 *Response curves for the four major filter types*

Attenuation is commonly expressed in units of decibels (dB). Frequency can be expressed in two forms either Hz or ω.

Finally, filter response curves may be plotted in linear-linear, log-linear, or log-log form.

- ***-3dB Frequency*** : This term, pronounced "minus 3dB frequency", corresponds to the input frequency that causes the output signal to drop by -3dB relative to the input signal. The -3dB frequency is also referred to as the cutoff frequency, and it is the frequency at which the output power is reduced by one-half (which is why this frequency is also called the "half-power frequency"), or at which the output voltage is the input voltage multiplied by $1/\sqrt{2}$. For low-pass and high-pass filters there is only one -3dB frequency. However, there are two -3dB frequencies for band-pass and notch filters—these are normally referred to as f1 and f2.

- ***Center frequency (f_0).*** The center frequency, a term used for band-pass and notch filters, is a central frequency that lies between the upper and lower cutoff frequencies.

- ***Bandwidth (B.W.).*** The bandwidth is the width of the passband, and the passband is the band of frequencies that do not experience significant attenuation when moving from the input of the filter to the output of the filter.

- ***Stopband frequency (f_s).*** This is a particular frequency at which the attenuation reaches a specified value.

For low-pass and high-pass filters, frequencies beyond the stopband frequency are referred to as the stopband.
For band-pass and notch filters, two stopband frequencies exist. The frequencies between these two stopband frequencies are referred to as the stopband.

5.2. Passive filter types

1- Passive Low Pass filter LPF

A Low Pass Filter is a circuit that can be designed to modify, reshape or reject all unwanted high frequencies of an electrical signal and accept or pass only those signals wanted by the circuit's designer.

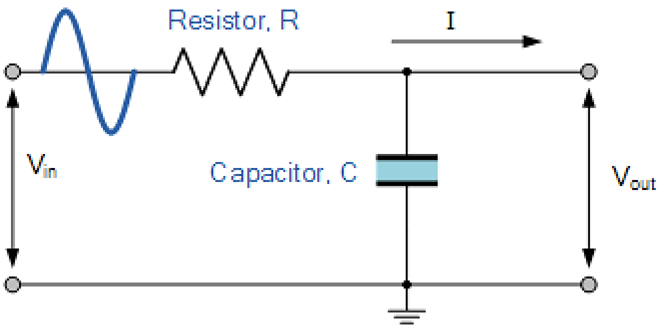


Fig. 5-3 Passive RC low pass filter circuit

RC Low Pass Filter, can be easily made by connecting together in series a single Resistor R with a single Capacitor C as shown in Fig. 5-3. The input signal (V_{IN}) is applied to both the R and C together but the output signal (V_{OUT}) is taken across the capacitor only.

The reactance of a capacitor X_C varies inversely with frequency, while the value of the resistor R remains constant as the frequency changes.

At low frequencies the capacitive reactance, (X_C) of the capacitor will be very large compared to the resistive value of the resistor R. This means that the voltage potential, V_C across the capacitor will be much larger than the voltage drop, V_R developed across the resistor. At high frequencies the reverse is true with V_C being small and V_R being large due to the change in the capacitive reactance the output signal value and causes the cutoff value.

The LPF circuit impedance is calculated as:

$$Z = \sqrt{R^2 + {X_c}^2} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (5.1)$$

The Voltage output:

The V_{out} across the *C* according to voltage divider equation is:

$$V_{out} = V_{in} \frac{X_c}{\sqrt{R^2 + {X_c}^2}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (5.2)$$

Example 5.1 A LPF circuit consisting of a resistor of 2 kΩ in series with a capacitor of 0.1uf is connected across a 4V sinusoidal supply. Calculate the output voltage (V_{OUT}) at a frequency of 100Hz and again at frequency of 10kHz.

Frequency Response of LPF (Cut-off frequency, Phase shift and Gain);

Cut-off frequency is the frequency at which the output signal drop -3db.

$$f_c = \frac{1}{2\pi RC} \text{ .... } (5.3)$$

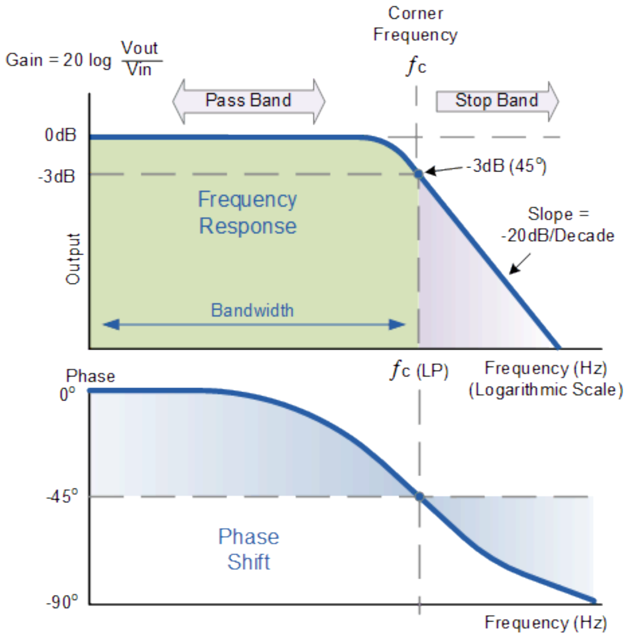


Fig. 5-4 Frequency response (Bode plot) and phase shift of LPF

When the frequency of the signal applied to LPF is increasing the phase shift between the input signal and output signal increases, in the cut-off frequency the phase shift is -45° as shown in Fig. 5-4.

Phase shift is;

$$phase\ shift\ \emptyset = -\arctan(2\pi f RC) \text{............................ } (5.4)$$

The gain of the input signal, at the frequencies less than the cut-off frequency, is unity and it is -3db at the cut-off frequency then after the cut-of frequency decreases to zero at a slope of -20dB/ Decade.

The gain of the LPF is;

$$Gain = 20\log(\frac{V_{out}}{V_{in}}) \text{ .... } (5.5)$$

Example 5.2 Then for our simple example of a LPF circuit above in Example 5.1, calculate the cut-off frequency (f_c) and the phase shift.

2- Passive High Pass filter HPF

The passive High Pass Filter HPF circuit only passes signals above the selected cut-off point (f_c) and eliminating any low frequency signals from input waveform.

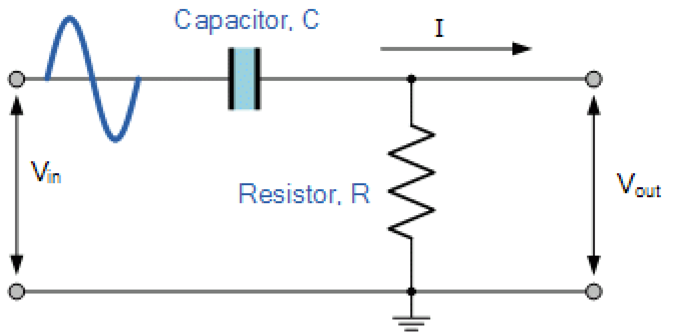


Fig. 5-5 Passive CR High pass filter circuit

RC Low Pass Filter, can be easily made by connecting together in series a single Capacitor C with a single Resistor R as shown in Fig. 5-5. The input signal (V_{IN}) is applied to both the C and R together but the output signal (V_{OUT}) is taken across the resistor only.

The reactance of a capacitor X_C varies inversely with frequency, while the value of the resistor R remains constant as the frequency changes.

In this circuit arrangement, the reactance of the capacitor (X_C) is very high at low frequencies so the capacitor acts like an open circuit and blocks any input signals at V_{IN} until the cut-off frequency point (f_C) is reached. Above this cut-off frequency point the reactance of the capacitor (X_C) has reduced sufficiently as to now act more like a short circuit allowing all of the input signal V_{IN} to pass directly to the output V_{out} across the resistor.

The voltage output and the circuit impedance of HPF is the same as the LPF in Eq. (5.1) & Eq. (5.2).

Example 5.3 A HPF circuit consisting of a resistor of 2kohm in series with a capacitor of 0.1uf is connected across a 4V sinusoidal supply. Calculate the output voltage (V_{OUT}) at a frequency of 100Hz and again at frequency of 10kHz.

Frequency Response of HPF (Cut-off frequency, Phase shift and Gain);

Cut-off frequency is the frequency at which the output signal drop -3db.

$$f_c = \frac{1}{2\pi RC} \quad \text{.............................} \quad (5.6)$$

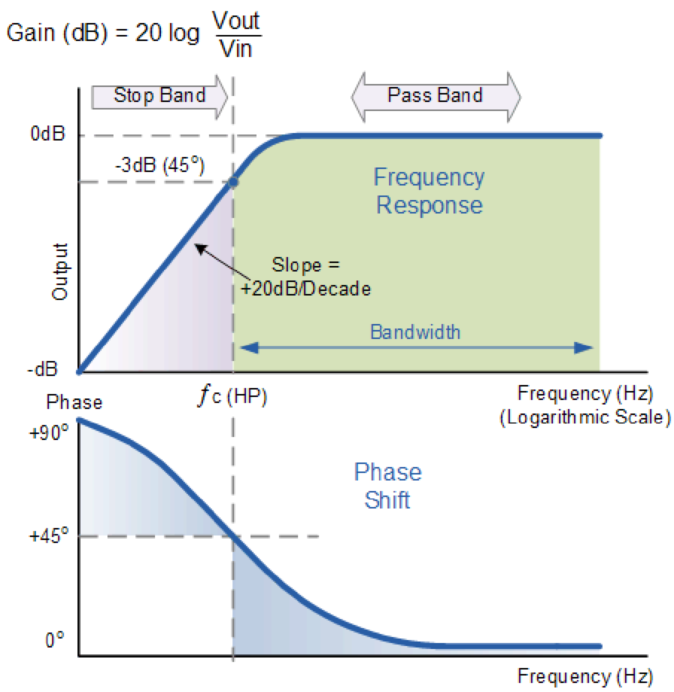


Fig. 5-6 Frequency response (Bode plot) and phase shift of LPF

The phase shift between the output signal and input signal applied to HPF is 45° at the cut-off frequency. When the frequency of the signal is increasing the phase shift decreases, and it is near to zero at the pass band.

Phase shift is;

$$phase\ shift\ \emptyset = \arctan(\frac{1}{2\pi f RC}) \quad \text{.......................... (5.7)}$$

The gain of the input signal, at the frequencies more than the cut-off frequency, is unity and it is -3db at the cut-off frequency then before the cut-of frequency decreases to zero at a slope of -20dB/ Decade.

The gain of the LPF is;

$$Gain = 20\log(\frac{V_{in}}{V_{out}}) \quad \text{.......................... (5.8)}$$

Example 5.4 Calculate the cut-off or "breakpoint" frequency (f_c) for a simple passive high pass filter consisting of an 82pF capacitor connected in series with a 240kΩ resistor.

3- Passive Band Pass filter BPF

The Band Pass Filter BPF circuit only passes signals within a certain "band" or "spread" of frequencies between the two cut-off frequencies, low (f_{CL}) and high cut-off (f_{CH}) frequency. This band of frequencies can be any width and is commonly known as the filters **Bandwidth B.W.**

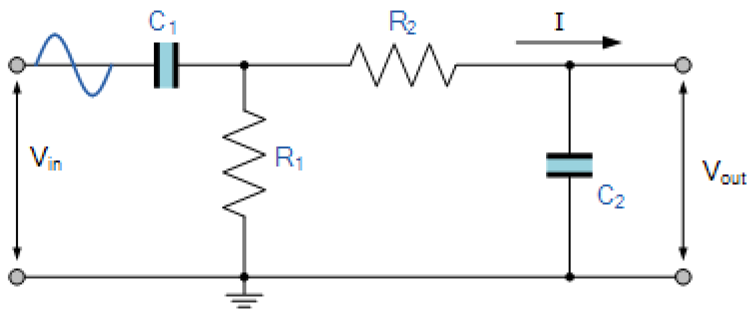


Fig. 5-7 Passive RC and CR Band pass filter circuit

CR-RC Band Pass Filter, By connecting or "cascading" together a single **High Pass Filter** circuit with a **Low Pass Filter** circuit, we can produce another type of passive RC filter called **Band Pass Filter** circuit. That passes a selected range or "band" of frequencies that can be either narrow or wide while attenuating the output signal voltage (V_{OUT})for all those outside of this range of frequency.

In this circuit arrangement, in the first stage of cascaded circuit (HPF) the reactance of the capacitor (X_{C1}) is very high at low frequencies so the capacitor acts like an open circuit and blocks any input signals at V_{IN} until the cut-off frequency point (f_L) is reached. Above this cut-off frequency point the reactance of the capacitor (X_{C1}) has reduced sufficiently as to now act more like a short circuit allowing all of the input signal V_{IN} to pass directly to the second stage circuit across the resistor R_1.

In the second stage of cascaded circuit (LPF) the reactance of the capacitor (X_{C2}) will be very large compared to the resistive value of the resistor R_2. This means that the voltage potential, V_C across the capacitor will be much larger than the voltage drop, V_R developed across the resistor R_2. Since any input signals in the second stage will be appear at V_{OUT} until the cut-off frequency point (f_H) is reached. Above this cut-off frequency point the reactance of the capacitor (X_{C2}) has reduced sufficiently as to now act more like a short circuit rejecting all of the input signal V_{IN} in the second stage to pass directly to the output.

Frequency Response of BPF (Cut-off frequency, Phase shift and Gain);

$$f_c = \frac{1}{2\pi RC} \quad \text{..........................} \quad (5.9)$$

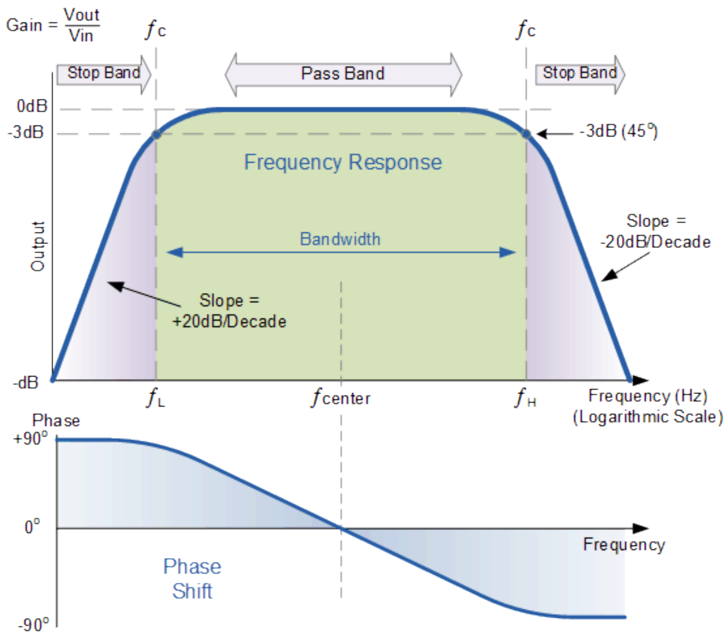


Fig. 5-8 Frequency response (Bode plot) and phase shift of BPF

The Bode Plot or frequency response curve above shows the signal is attenuated at low frequencies with the output increasing at a slope of +20dB/Decade until the frequency reaches the "lower cut-off" point f_L. At this frequency the output voltage V_{OUT} is again 1/√2 = 70.7% of the input signal V_{IN} value or -3dB of the input.

The output continues at maximum gain until it reaches the "upper cut-off" point f_H where the output decreases at a rate of -20dB/Decade attenuating any high frequency signals.

$$f_L = \frac{1}{2\pi R_1 C_1} \quad \text{.........................} \quad (5.10)$$

$$f_H = \frac{1}{2\pi R_2 C_2} \quad \text{.........................} \quad (5.11)$$

Bandwidth, The range of frequencies between the f_L and f_H.

$$BW = f_H - f_L \quad \text{.........................} \quad (5.12)$$

Centre Frequency (Resonant Pea), The point of maximum output gain and it is generally the geometric mean of the two -3dB value between the lower and upper cut-off points. This geometric mean value is calculated as;

$$f_r = \sqrt{f_H \ * f_L} \quad \text{.........................} \quad (5.13)$$

BPF is a second-order (two-pole) type filter because it has two reactive components within its circuit structure, then the phase angle will be twice that of the previously seen first-order filters, ie, 180o. The phase angle of the output signal between +90° and -90°, also zero at the f_r.

The gain of the BPF is;

$$Gain = 20\log(\frac{V_{out}}{V_{in}}) \quad \text{.........................} \quad (5.14)$$

Example 5.5 A BPF is to be constructed using RC components that will only allow a range of frequencies to pass above 1kHz and below 30kHz. Assuming that both the resistors have values of 10kΩ, calculate the values of the two capacitors required.

The main disadvantage of passive filters:

1- The amplitude of the output signal is less than the input signal
2- The gain is never greater than unity
3- The load impedance affects the filters characteristics.

5.3 Active filter types

Active Filters contain active components such as operational amplifiers (Op-Amp), and transistors within their circuit design. They draw their power from an external power source and use it to boost or amplify the output signal.

1- Active Low Pass filter

By combining a basic **RC passive Low Pass Filter** circuit with an **(Op-Amp)**, for amplification and gain control, we can create an Active Low Pass Filter circuit complete with amplification.

A- Active LPF with unity gain

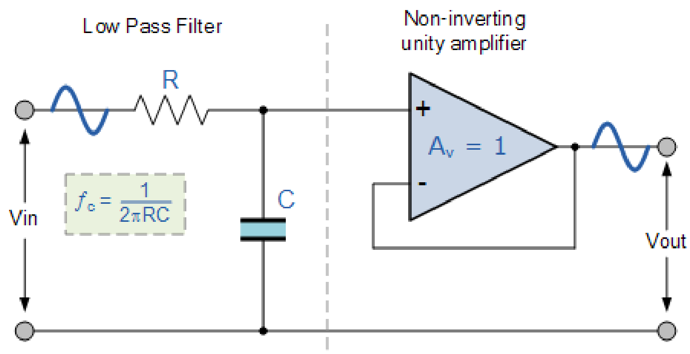


Fig. 5-9 Active LPF circuit with unity gain

For a **non-inverting** Op-Amp circuit, the magnitude of the voltage gain is given as;

$$DC\ gain = 1 \quad \text{.........................} \ (5.15)$$

The advantage of this configuration is that the op-amps high input impedance prevents excessive loading on the filters output while its low output impedance prevents the filters cut-off frequency point from being affected by changes in the impedance of the load.

The cut-off frequency of the LPF calculate according to the RC circuit and it is;

$$f_c = \frac{1}{2\pi RC} \quad \text{.........................} \ (5.16)$$

B- Active LPF with Amplification

The frequency response of the circuit will be the same as that for the passive RC filter (Eq. 5.16)

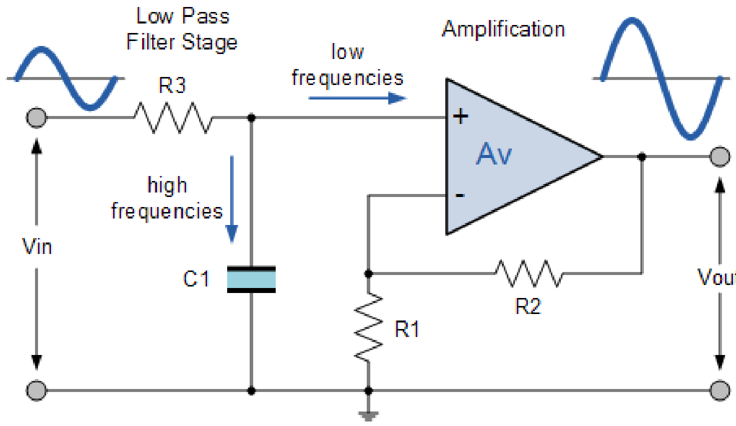


Fig. 5-10 Active LPF circuit with amplification

The amplitude of the output is increased by the pass band gain (DC gain A_F) of the (Op-Amp) amplifier. For a non-inverting amplifier circuit, the magnitude of the voltage gain is given as:

$$DC\ gain,\ A_F = 1 + \frac{R_2}{R_1} \quad \text{.........................} \ (5.17)$$

Therefore, the gain of an active low pass filter as a function of frequency will be:

$$Voltage\ Gain\ (Av) = \frac{V_{OUT}}{V_{IN}} = \frac{A_F}{\sqrt{1+\left(\frac{f}{fc}\right)^2}} \quad \text{.........................} \ (5.18)$$

Where:

A_F = the pass band gain of the filter, (1 + R2/R1)
f = the frequency of the input signal in Hz
f_c = the cut-off frequency in Hz

Thus, the operation of a low pass active filter can be verified from the frequency gain equation above as:

1- At very low frequencies, $f < fc$ $\quad \frac{V_{OUT}}{V_{IN}} \cong A_F$

2- At the cut-off frequency, $f = fc$ $\quad \frac{V_{OUT}}{V_{IN}} = \frac{A_F}{\sqrt{2}}$

3- At very high frequencies, $f > fc$ $\quad \frac{V_{OUT}}{V_{IN}} < A_F$

Thus, the **Active Low Pass Filter** has a constant gain A_F from 0Hz to the high frequency cut-off point, f_C. At f_C the gain is $0.707A_F$, and after f_C it decreases at a constant rate as the frequency increases. That is, when the frequency is increased tenfold (one decade), the voltage gain is divided by 10.

Magnitude of Voltage Gain in (dB);

Example 5.6 Design a non-inverting active low pass filter circuit that has a gain of ten at low frequencies, a high frequency cut-off or corner frequency of 159Hz and an input impedance of 10KΩ.

C- Non-inverting and inverting active LPF with amplification

NOTE: If the external impedance connected to the input of the filter circuit changes, this impedance change would also affect the cut-off frequency (corner frequency) of the filter (components connected together in series or parallel).

One way of avoiding any external influence is to place the capacitor in parallel with the feedback resistor R_2 effectively removing it from the input but still maintaining the filters characteristics.

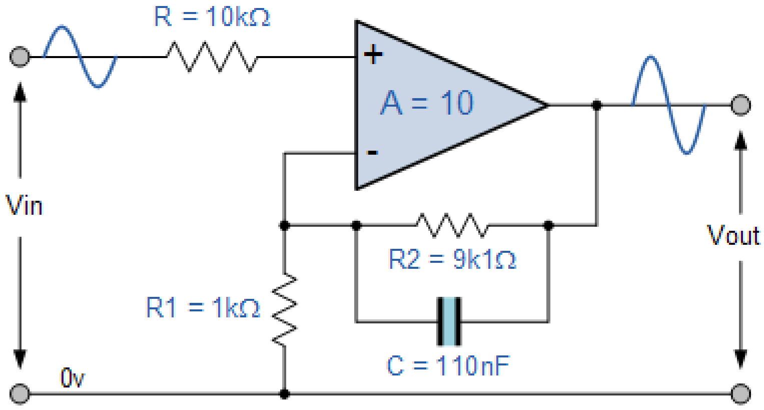


Fig. 5-11 Simplified non-inverting active LPF circuit with amplification

$$f_c = \frac{1}{2\pi R_2 C} \quad \text{.........................} \quad (5.19)$$

$$\frac{V_{OUT}}{V_{IN}} = 1 + \frac{R_2||X_C}{R_1} \quad \text{.........................} \quad (5.20)$$

The inverting LPF is given as;

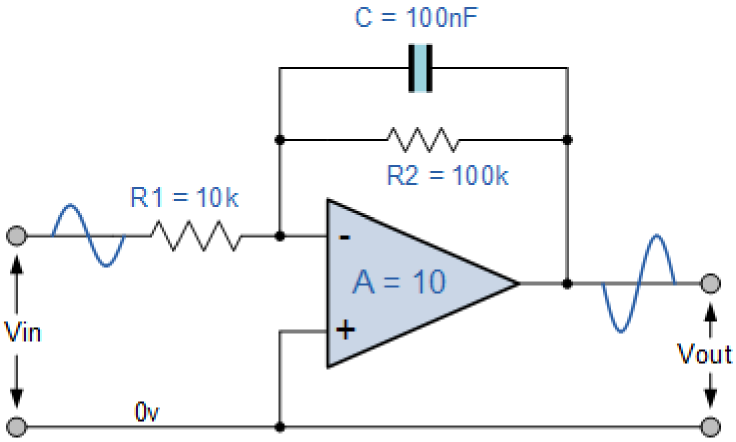


Fig. 5-12 Inverting active LPF circuit with amplification

$$\frac{V_{OUT}}{V_{IN}} = -\left(\frac{R_2||X_C}{R_1}\right) \quad \text{.........................} \quad (5.21)$$

2- Active High Pass filter HPF

By combining a basic **RC passive High Pass Filter** circuit with an **(Op-Amp)**, for amplification and gain control, we can create an Active High Pass Filter circuit complete with amplification.

A- Active HPF with unity gain

The frequency response of the circuit will be the same as that for the passive RC filter (Eq. 5.16). As in Eq.(5.15) for a non-inverting Op-Amp circuit, the magnitude of the DC voltage gain is one.

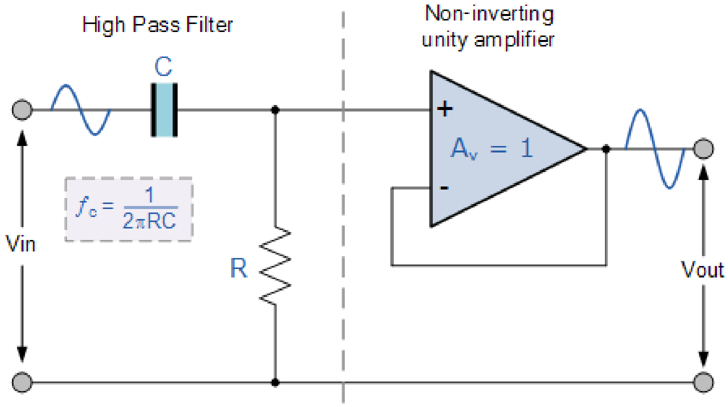


Fig. 5-13 Active HPF circuit with unity gain

B- Active HPF with Amplification

The frequency response of the circuit is the same as that of the passive RC filter (Eq. 5.16), except that the amplitude of the output signal is increased by the pass band gain (DC gain A_F) of the (Op-Amp) amplifier.

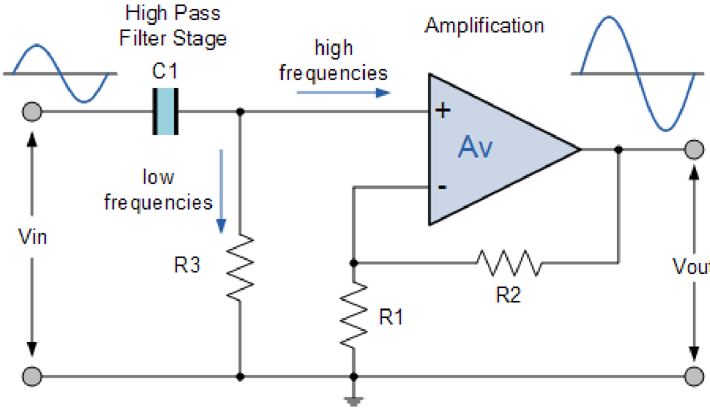


Fig. 5-14 Active HPF circuit with amplification

For a non-inverting amplifier circuit, the magnitude of the voltage gain is given as:

$$DC\ gain,\ A_F = 1 + \frac{R_2}{R_1} \quad \text{.........................} (5.22)$$

Therefore, the overall voltage gain of an active high pass filter as a function of frequency will be:

$$Voltage\ Gain\ (Av) = \frac{V_{OUT}}{V_{IN}} = \frac{A_F\left(\frac{f}{fc}\right)}{\sqrt{1+\left(\frac{f}{fc}\right)^2}} \quad \text{.........................} (5.23)$$

Where:

A_F = the pass band gain of the filter, (1 + R2/R1)
f = the frequency of the input signal in Hz
f_c = the cut-off frequency in Hz

Thus, the operation of a low pass active filter can be verified from the frequency gain equation above as:

1- At very low frequencies, $f < fc$ $\quad \frac{V_{OUT}}{V_{IN}} < A_F$
2- At the cut-off frequency, $f = fc$ $\quad \frac{V_{OUT}}{V_{IN}} = \frac{A_F}{\sqrt{2}}$
3- At very high frequencies, $f > fc$ $\quad \frac{V_{OUT}}{V_{IN}} \cong A_F$

Then, the **Active High Pass Filter** has a gain A_F that increases from 0Hz to the low frequency cut-off point, f_C at 20dB/decade as the frequency increases. At f_C the gain is 0.707*A_F, and after f_C all frequencies are pass band frequencies so the filter has a constant gain A_F with the highest frequency being determined by the closed loop bandwidth of the op-amp.

Magnitude of Voltage Gain in (dB);

Phase Shift of Active HPF is the same as that given for the passive RC HPF and leads that of the input signal. It is equal to +45° at the cut-off frequency f_C value and is given as:

$$phase\ shift\ \emptyset = \arctan\left(\frac{1}{2\pi fRC}\right) \quad \text{.........................} (5.24)$$

Example 5.7 A first order active high pass filter has a pass band gain of two and a cut-off corner frequency of 1kHz. If the input capacitor has a value of 10nF, calculate the value of the cut-off frequency determining resistor and the gain resistors in the feedback network. Also, plot the expected frequency response of the filter.

C- Inverting active HPF with amplification

The frequency response and the magnitude of the voltage gain of the inverting LPF circuit are given as:

$$f_c = \frac{1}{2\pi R_1 C_1} \quad \text{.......................... (5.25)}$$

$$A_V = -\frac{R_2}{R_1} \quad \text{.......................... (5.26)}$$

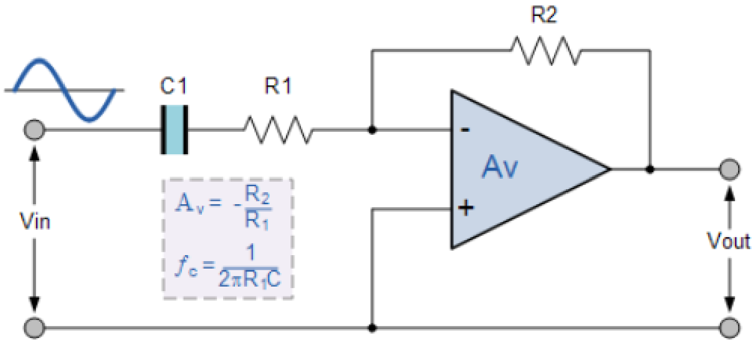


Fig. 5-15 Inverting active HPF circuit with amplification

Frequency Response Curve of HPF;

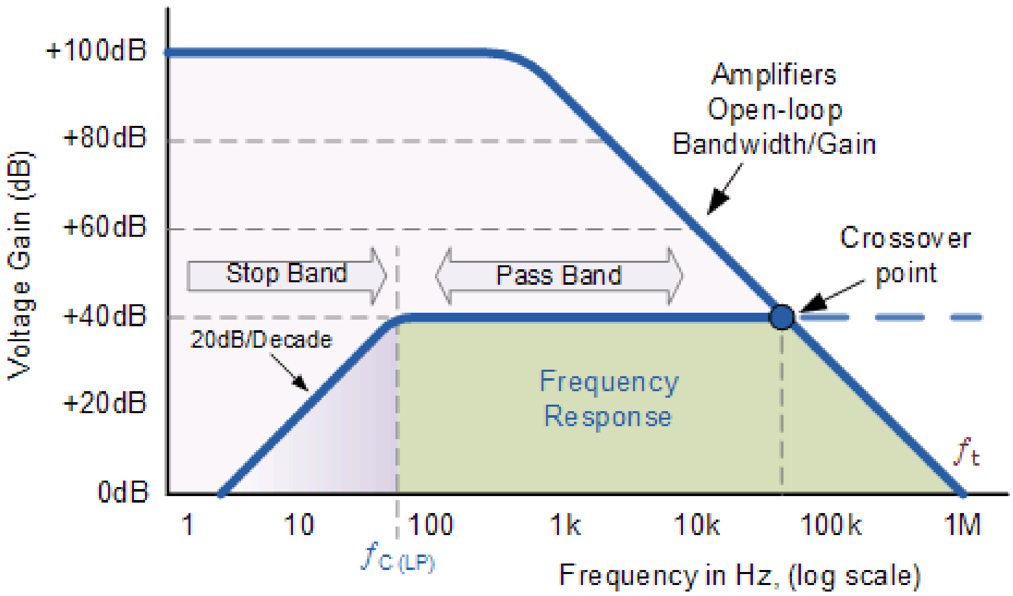


Fig. 5-16 Frequency Response Curve of HPF

3- Active Band Pass Filter BPF

Active Band Pass Filter can be easily made by cascading together a single Low Pass Filter with a single High Pass Filter as shown.

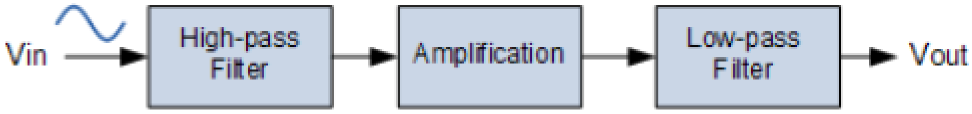


Fig. 5-17 Block diagram of active BPF

The cut-off frequency of the low pass filter (LPF) is higher than the cut-off frequency of the high pass filter (HPF) and the difference between the frequencies at the -3dB point will determine the **"bandwidth"** of the band pass filter while attenuating any signals outside of these points.

A- Non-inverting active Band pass filter with amplification

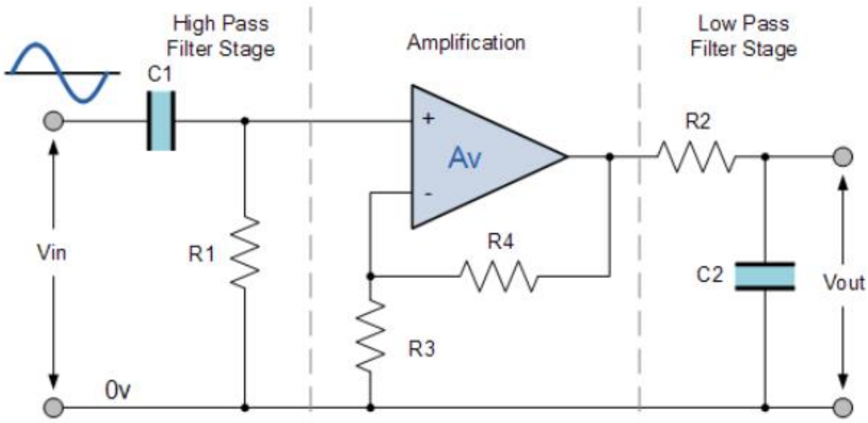


Fig. 5-18 Non-inverting active BPF circuit with amplification

Frequency Response of BPF (Cut-off frequency and Gain):

The Bode Plot or frequency response curve Fig.5.19 shows the signal is attenuated at low frequencies with the output increasing at a slope of +20dB/Decade until the frequency reaches the "lower cut-off" point f_L. At this frequency the output voltage V_{OUT} is again $1/\sqrt{2}$ = 70.7% of the input signal V_{IN} value or -3dB of the input.

The output continues at maximum gain until it reaches the "upper cut-off" point f_H where the output decreases at a rate of -20dB/Decade attenuating any high frequency signals.

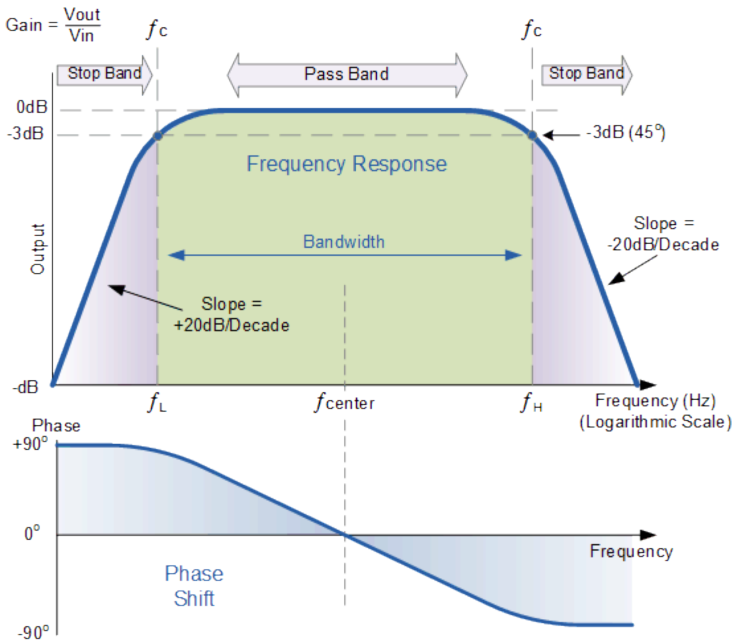


Fig. 5-19 Frequency response (Bode plot) and phase shift of active BPF

The Low and High cut-off frequencies is given as;

$$f_L = \frac{1}{2\pi R_1 C_1} \quad \text{......................... (5.27)}$$

$$f_H = \frac{1}{2\pi R_2 C_2} \quad \text{......................... (5.28)}$$

The gain of the active BPF is the overall of the circuit gain;

$$Voltage\ Gain\ (Av) = \frac{V_{OUT}}{V_{IN}} = \frac{A_F\left(\frac{f}{fc}\right)}{1+\left(\frac{f}{fc}\right)^2} \quad \text{......................... (5.29)}$$

$$Gain(dB) = 20\log\left(\frac{V_{out}}{V_{in}}\right) \quad \text{......................... (5.30)}$$

NOTE: This cascading together of the individual low and high pass passive filters produces a low "Q-factor" type filter circuit which has a wide pass band. While the above passive tuned filter circuit will work as a band pass filter, the pass band (bandwidth) can be quite wide and this may be a problem if we want to isolate a small band of frequencies

B- Inverting active Band Pass Filter

The frequency response and the magnitude of the voltage gain of the inverting BPF circuit are given as:

$$f_L = \frac{1}{2\pi R_1 C_1} \text{ .... (5.31)}$$

$$f_H = \frac{1}{2\pi R_2 C_2} \text{ .... (5.32)}$$

$$Voltage\ Gain\ (Av) = \frac{V_{OUT}}{V_{IN}} = -\frac{R_2}{R_1} \text{ .... (5.33)}$$

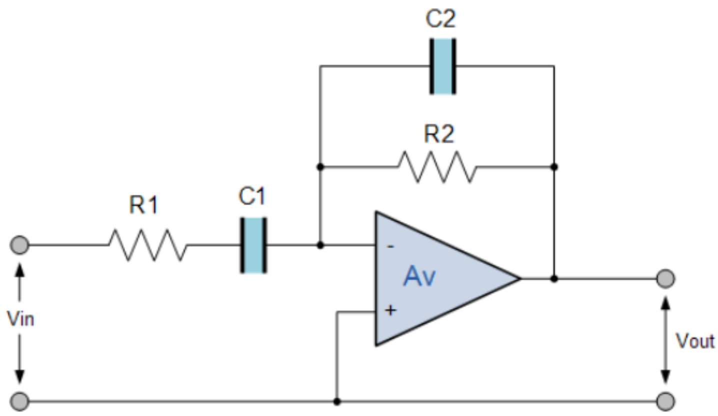


Fig. 5-20 Iverting active BPF circuit with amplification

Bandwidth of the BPF, The range of frequencies between the f_L and f_H.

$$BW = f_H - f_L \text{ .... (5.34)}$$

Centre Frequency of the BPF, The point of maximum output gain and it is generally the geometric mean of the two -3dB value between the lower and upper cut-off points. This geometric mean value is calculated as;

$$f_r = \sqrt{f_H * f_L} \text{ .... (5.35)}$$

BPF is a second-order (two-pole) type filter because it has two reactive components within its circuit structure, then the phase angle will be twice that of the previously seen first-order filters, ie, 180°. The phase angle of the output signal between +90° and -90°, also zero at the f_r.

6- Mixers

A mixer is a nonlinear electronic circuit that combines two signals are applied to a mixer, and it produces new signals at the sum and difference of the original frequencies.

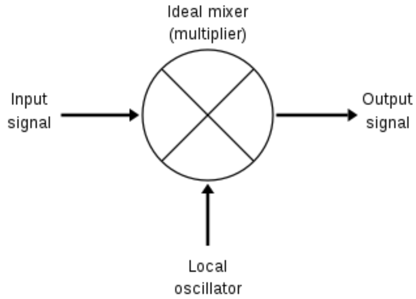


Fig. 6-1 Frequency mixer symbol

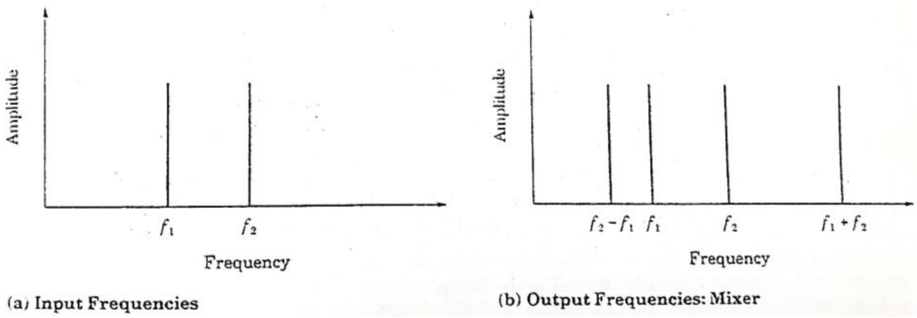


Fig. 6-2 Output spectra mixer

6-1 Square Law Mixers

The square law mixer is the simplest to understand mathematically, its output is given by:

$$V_o = A\, V_i + B\, V_i^2 \quad \text{.........................} \; (6.1)$$

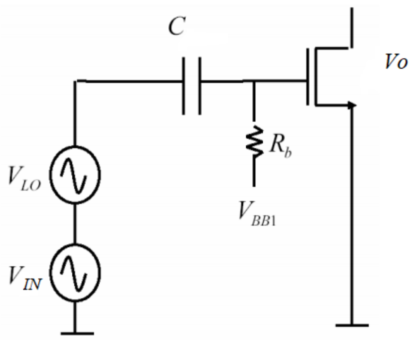


Fig. 6-3 Simple square law mixer circuit

Apply two signals, the signals will be ***sine*** wave different frequencies (f_1 and f_2) and amplitudes are 1 volt, then:

$$V_i = sin(w_1 t) + sin(w_2 t) \text{ .... } (6.2)$$

Substitute 6.2 in 6.1;

$$V_o = A(sin(w_1 t) + sin(w_2 t)) + B(sin(w_1 t) + sin(w_2 t))^2$$

$$V_o = A\,sin(w_1 t) + A\,sin(w_2 t) + B\,sin(w_1 t)^2 + B\,sin(w_2 t)^2 + 2B\,sin(w_1 t)\,sin(w_2 t) \text{ ... } (6.3)$$

In Eq. (6.3) there are many terms, first term is;

$$A\,sin(w_1 t)$$

Second terms is;

$$A\,sin(w_2 t)$$

Where

$$sin^2 A = \frac{1}{2} - \frac{1}{2}\cos(2A)$$

$$sinA\ sinB = \frac{1}{2}[\cos(A-B) - \cos(A+B)]$$

Then third term becomes;

$$B\,sin(w_1 t)^2 = \frac{B}{2} - \frac{B}{2}\cos(2w_1 t)$$

And the fourth term is;

$$B\,sin(w_2 t)^2 = \frac{B}{2} - \frac{B}{2}\cos(2w_2 t)$$

Then final term is;

$$2B\,sin(w_1 t)\,sin(w_2 t) = B\,[\cos(w_1 - w_2)t - \cos(w_1 + w_2)t]$$

6-2 Balanced Mixers

A balanced mixer is one in which the input frequencies do not appear at the output. This is easy to show mathematically by;

$$V_o = A\,V_{i\,1}\,V_{i\,2} \text{ .... } (6.4)$$

When

$$V_{i\,1} = sin(w_1 t)$$

$$V_{i\,2} = sin(w_2 t)$$

Then the output will be

$$V_o = A\,sin(w_1 t)\ \ sin(w_2 t) \text{......................... } (6.5)$$

Where

$$sinA\ sinB = \frac{1}{2}[\cos(A-B) - \cos(A+B)]$$

So the output is;

$$V_o = \frac{A}{2}\,[\cos(w_1 - w_2)t - \cos(w_1 + w_2)t] \text{ .... } (6.6)$$

As predicted, this mixer produces only the sum and difference of the put frequencies.

6-3 Transistor Mixers

1- Bipolar transistor mixers

The bipolar transistor mixer of figure below resembles conventional tuned amplifier, except that is has two inputs

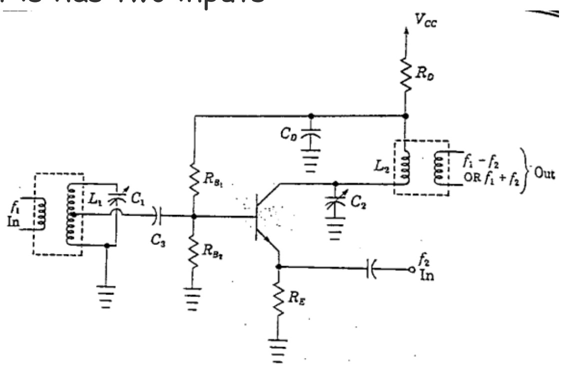


Fig. 6-4 Bipolar mixer circuit

- The output may be tuned either to the sum or the difference of the two input signal frequencies
- The resonant circuits L_1C_1 and L_2C_2 are tuned to f1 and the required output frequency
- RB_1, RB_2, and RE from the usual voltage divider bias circuit

2- JFET mixer

The JFET circuit of figure below illustrates the technique of summing the two input signals

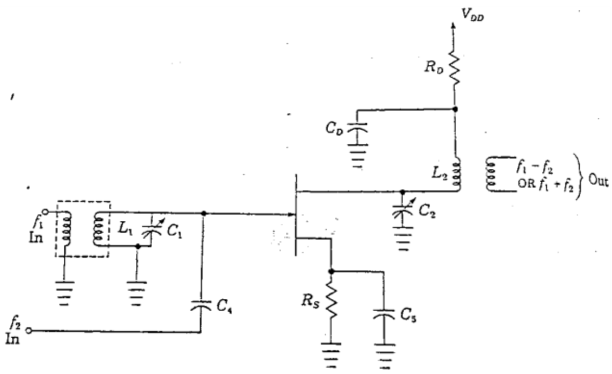


Fig. 6-5 JFET mixer circuit

- Both f_1 and f_2 are applied to the gate
- The tuned circuit L_1-C_1 and L_2-C_2 have the same functions as before
- C_4 couples the f_2 input to the circuit

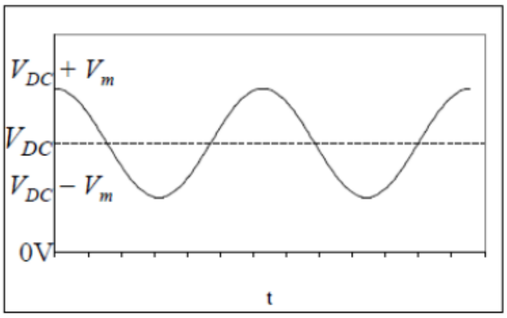
7- Amplitude Modulation Circuits

7-1 AM Modulation

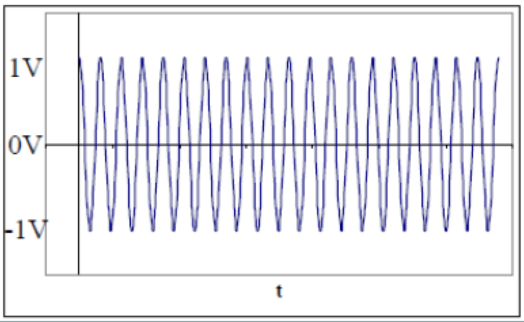
1- The signal

Amplitude modulation (AM) is a modulation technique used in electronic communication, most commonly for transmitting information via a radio carrier wave.

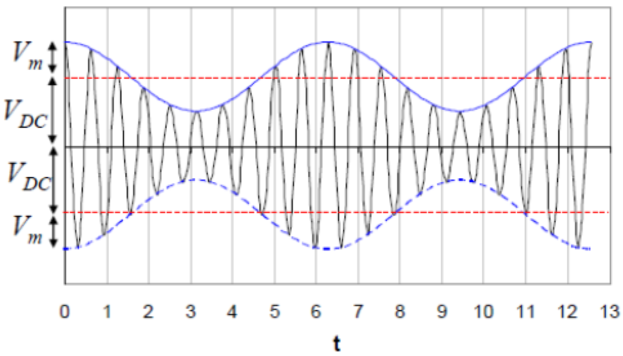
An AM signal can be produced; by using the instantaneous amplitude of the information signal (the baseband or modulating signal) to vary the peak of amplitude of higher frequency signal (Currier signal) as shown in Fig. 7-1.



(a) $V_{Dc}+V_mCos(2\pi f_mt)$



(b) $Cos(2\pi f_ct)$



(c) $V_{Dc}\,Cos(2\pi f_ct) + V_m/2\,Cos(2\pi(f_c - f_m)t) + V_m/2\,Cos(2\pi(f_c + f_m)t)$

Fig. 7-1 (a) shows a 1 kHz sine wave which can be combined with the 10 kHz signal shown in (b) to produce the AM signal of (c).

Amplitude modulation is essentially a nonlinear process. As with any non-linear interaction between signals, sum and difference frequencies are produced that, in the case of amplitude modulation, contain the information to be transmitted. Also we can see the linear addition representation for two signals in Fig. 7-2.

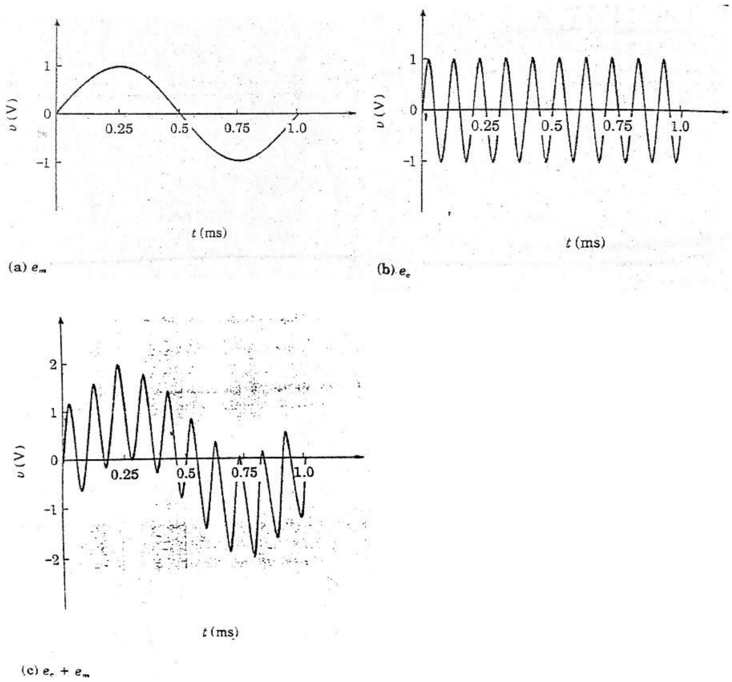


Fig. 7-2 (a) First signal (b) Second signal (c) Linear addition of two signals.

Fig. 7-3 shows the frequency-domain representation for AM modulation and the linear addition for two signals. **Note**, the AM signal has no components at the modulating frequency: all information is transmitted at frequencies near that of the carrier.

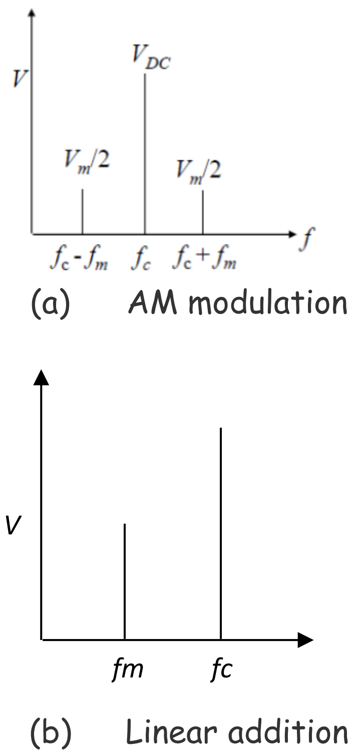


Fig. 7-3 AM modulation and linear addition in the frequency domain

7-2 A simple AM Modulator Circuit

There are many ways to achieve the result shown in Fig. 7-1. Perhaps the easiest circuit to understand is that shown in Fig. 7-4. This circuit be used in the transmitter side.

This is a simplified version of the type of circuit used in many AM transmitters:

1- The RF carrier is amplified in a Class C stage.
2- The modulating signal is applied to the primary of a transformer
3- The secondary of which is in series with the supply voltage to the amplifier.

From Class C lectures, the peak output voltage of a tuned Class C amplifier like that in Fig. 7-4 is approximately twice the collector supply voltage.

How this circuit work:

1- In this circuit, that voltage, and hence the output of the modulated stage (AM out), varies with the instantaneous value of the modulating signal (In em).
2- As the level of the audio signal increases, so does the effect on the RF signal.
3- When the modulating voltage, at the modulating transformer secondary, has a peak amplitude equal to Vcc, the collector supply voltage (measured at point A in the Fig.7-3) will vary from zero to 2Vcc and the amplitude of the output waveform will vary from zero to twice its value without modulation as in Fig. 7-1(c) this is called 100% modulation.

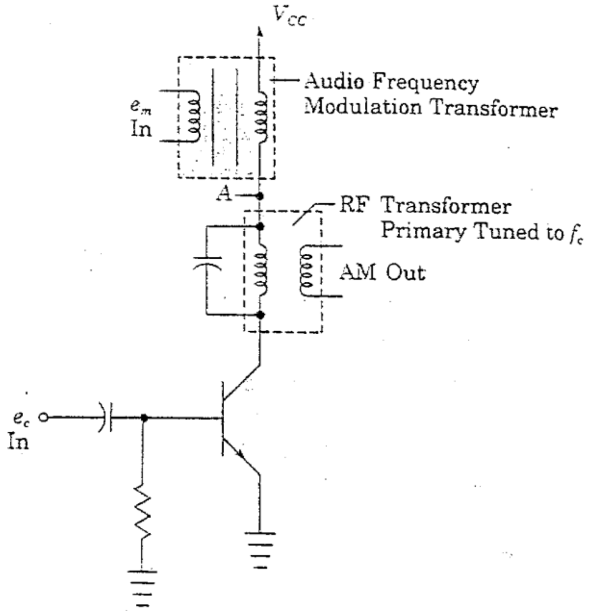


Fig. 7-4 Simplified AM modulator

7-3 A simple AM De-Modulator Circuit

The original information be recovered (demodulated) at the receiver. Demodulation can be done very simply with full-carrier AM.

As shown in Fig.7-5, the original signal can be recovered by first rectifying the modulated signal, then low-pass filtering the result, this is called an **envelope detector**.

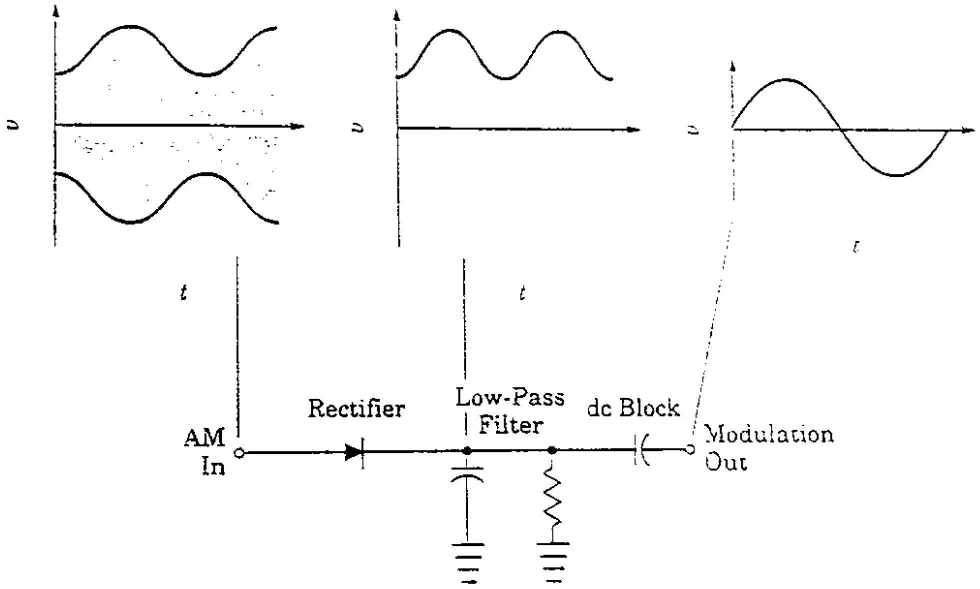


Fig. 7-5 Simplified AM demodulator

7-4 Time Domain Analysis for AM Modulation

An AM modulation is created by using the instantaneous modulating signal voltage to vary the amplitude of modulated signal.
The carrier signal is almost always a **Sine wave**. The modulating signal can be a sine wave, but is more often an arbitrary waveform, such as an audio signal.

However, an analysis of cos-wave modulation is very useful. We can express the AM modulation relationship by means of an equation:

$$V(t) = (V_{DC} + e_m(t))\, \mathrm{e}_c(\mathrm{t}) \quad \ldots\ldots\ldots\ldots (7.1)$$

Where;

$V(t)$ = Instantaneous amplitude of the modulated signal in volts.
V_{DC} = It represents the Vcc voltage in Fig. 7-4 can be from 0 to +V.
$e_m(t)$ = Instantaneous amplitude of the modulating signal in volts.
$e_c(t)$ = Instantaneous amplitude of the carrier signal in volts.

Note: You can see the Fig. 7-4 circuit to note the mean of Eq.(7.1).

- If the modulating (baseband or information) signal is a cosin wave and the carrier signal also a cosin wave, Eq.(7.1) has the following format;

$$V(t) = (V_{DC} + V_m \cos(2\pi f_m t)) \cos(2\pi f_c t) \quad \ldots\ldots\ldots\ldots (7.2)$$

Where;

V_m = **Peak amplitude** of the modulating signal in volts.
f_m = Frequency of the modulating signal
f_c = Frequency of the carrier signal.
and the other variables are as defined for Eq.7-1.

Using the trigonometric identity cos(A) cos(B) = 1/2 [cos(A+B) + cos(A- B)];

$$V(t) = \underbrace{V_{DC}\cos(2\pi f_c t)}_{Carrier} + \underbrace{\frac{V_m}{2}\cos(2\pi(f_c + f_m)t)}_{upper\ sideband} + \underbrace{\frac{V_m}{2}\cos(2\pi(f_c - f_m)t)}_{Lower\ sideband} \quad \ldots\ldots\ldots\ldots (7.3)$$

- If the modulating (baseband or information) signal is a sine wave and the carrier signal also a sine wave, Eq.(7.1) has the following format;

$$V(t) = (V_{DC} + V_m \sin(2\pi f_m t)) \sin(2\pi f_c t) \quad \ldots\ldots\ldots\ldots (7.4)$$

Using the trigonometric identity sin(A) sin(B) = 1/2 [cos(A-B) - cos(A+ B)];

$$V(t) = \underbrace{V_{DC}\sin(2\pi f_c t)}_{Carrier} + \underbrace{\frac{V_m}{2}\cos(2\pi(f_c - f_m)t)}_{Lower\ sideband} - \underbrace{\frac{V_m}{2}\cos(2\pi(f_c + f_m)t)}_{upper\ sideband} \quad \ldots\ldots\ldots\ldots (7.5)$$

The Eq.(7.3) & Eq.(7.5) represent the **Double Sideband Amplitude Modulation DSBAM**

Components	Carrier	Upper Sideband (USB)	Lower Sideband (LSB)
Amplitude	V_{DC}	$\frac{V_m}{2}$	$\frac{V_m}{2}$
Frequency	f_c	$f_c + f_m$	$f_c - f_m$

Table. 7-1 Simplified AM demodulator

Example 7.1: A cosine wave carrier which has an RMS voltage of 2 volt and a frequency of 1.5 MHz is modulated by as cosin wave with a frequency of 500 Hz and amplitude of 1 volt RMS. Write the equation for the resulting signal.

Sol:
The required voltage is peak voltage therefore we will cahnge the RMS to peak value

$$V_{Dc} = \sqrt{2}\, x\, 2 = 2.83\, v$$

$$V_m = \sqrt{2}\, x\, 1 = 1.41\, v$$

So the AM modulation equation is;

$$V(t) = 2.83\cos(2\pi 1.5 * 10^6 t) + \frac{1.41}{2}\cos(2\pi(1.5 * 10^6 - 500)t) + \frac{1.41}{2}\cos(2\pi(1.5 * 10^6 + 500)t)$$

$$V(t) = 2.83\cos(9.424 * 10^6 t) + 0.705\cos(9.421 * 10^6 t) + 0.705\cos(9.427 * 10^6 t)$$

7-4 AM Modulation Index

The amount by which the signal amplitude is changed in modulation depends on the ratio between the amplitudes of the modulating signal V_m and the carrier signal V_{DC}. This ratio is defined as the modulation index (or modulation depth) *m*. It can be expressed mathematically as;

$$m = \frac{V_m}{V_{DC}} \quad \text{.......................... (7.6)}$$

Substituting *m* into Eq.(7.4) gives;

$$V(t) = V_{DC}(1 + m\sin(2\pi f_m t))\sin(2\pi f_c t) \quad \text{.......................... (7.7)}$$

When;

- *m* = 0 this means V_m = 0 and we have unmodulated carrier signal as in Fig. 7-6(a)
- *m* varies between 0 and 1, the change due to modulation become more cleared as in Fig. 7-6(b,c,d)
- *m > 1* , over-modulation case.

Modulation can also be expressed as a percentage, with percent modulation found by multiplying *m* by 100.

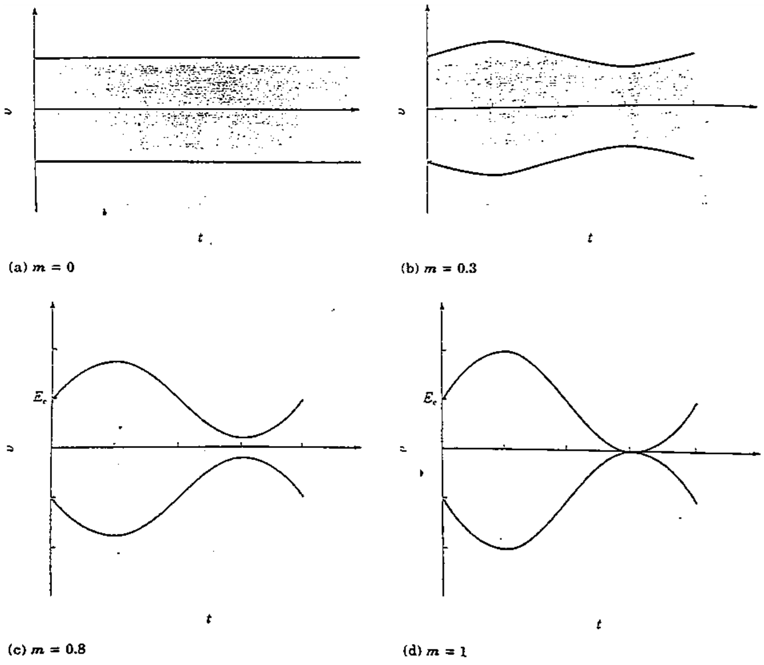


Fig. 7-6 AM envelopes for various values of *m*

Example 7.2: Calculate m for the signal of Example 7.1. and write the equation for this signal in the form of Eq.(7.7).

7-5 Over-modulation

When the modulation index is greater than 1 (m > 1), the modulation signal is said to be over- modulated.

Fig. 7-7(a) shows the result of simply substituting *m* = 2 into Eq. (7.7). As you can see, the envelope no longer resembles the modulating signal. Thus the type of demodulator described earlier no longer gives undistorted results, and the signal is no longer a full-carrier AM signal. Therefore, *m* must be kept less than or equal to 1.

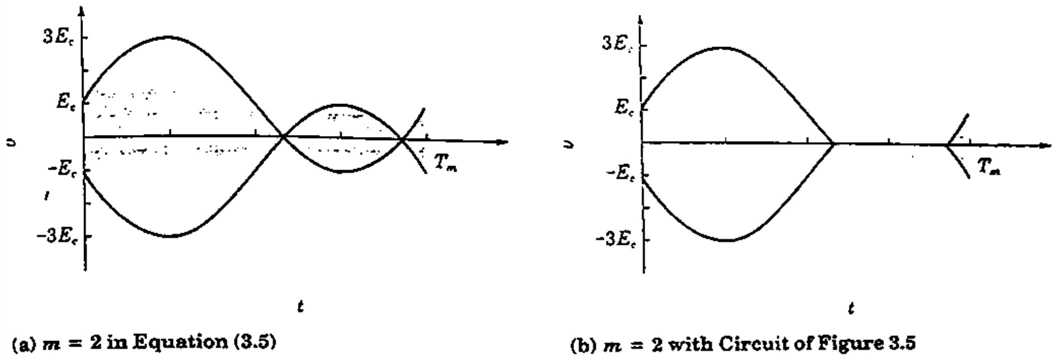


Fig. 7-7 Over-modulation

7-6 Modulation Index for Multiple Modulating Frequencies

Practical the information signal is more likely to be a voice signal, which contains many frequencies. We can use the idea of Fourier series to consider it as a series of sine waves of different frequencies. When there are two or more sine waves of different, uncorrelated fre- quencies (that is, frequencies that are not multiples of each other) modulating a single carrier, *m* is calculated by using the Eq. (7.8):

$$m_T = \sqrt{m_1^2 + m_2^2 + m_3^2 + m_4^2 + \cdots} \quad \text{.........................} (7.8)$$

Where:

m_T = total resultant modulation index

m_1, m_2 = modulation indices due to the individual modulating components.

Example 7.3: Find the modulation index if a 10 V carrier is amplitude modulated by three different frequencies, with amplitudes of 1 V, 2 V, and 3 V respectively.

Sol:

$$m_1 = \frac{1}{10} = 0.1 \qquad m_2 = \frac{2}{10} = 0.2 \qquad m_3 = \frac{3}{10} = 0.3$$

$$m_T = \sqrt{m_1^2 + m_2^2 + m_3^2}$$

$$m_T = \sqrt{0.1^2 + 0.2^2 + 0.3^2}$$

=0.374

7-7 Measurement of Modulation

If we let Vm and V_{DC}, be the peak modulation and carrier voltages respectively, then we can see, either by using Eq. (7.4) or by inspecting Fig. 7-8, that the maximum envelope voltage is simply;

$$V_{max} = V_{Dc} + V_m \quad \text{.......................... (7.9)}$$

and the minimum envelope voltage is;

$$V_{min} = V_{Dc} - V_m \quad \text{.......................... (7.10)}$$

Note, by the way, that this agrees with the conclusions expressed earlier: for m = 0, the peak voltage is V_{Dc}, and for m = 1, the envelope voltage ranges from $2V_{Dc}$ to zero.

Applying a little algebra to the above expressions, it is easy to show that:

$$m = \frac{V_{max} - V_{min}}{V_{max} + V_{min}} \quad \text{.......................... (7.11)}$$

Note: So it is quite easy to find *m* by displaying the envelope on an oscilloscope and measuring the maximum and minimum peak-to-peak values for the envelope voltage.

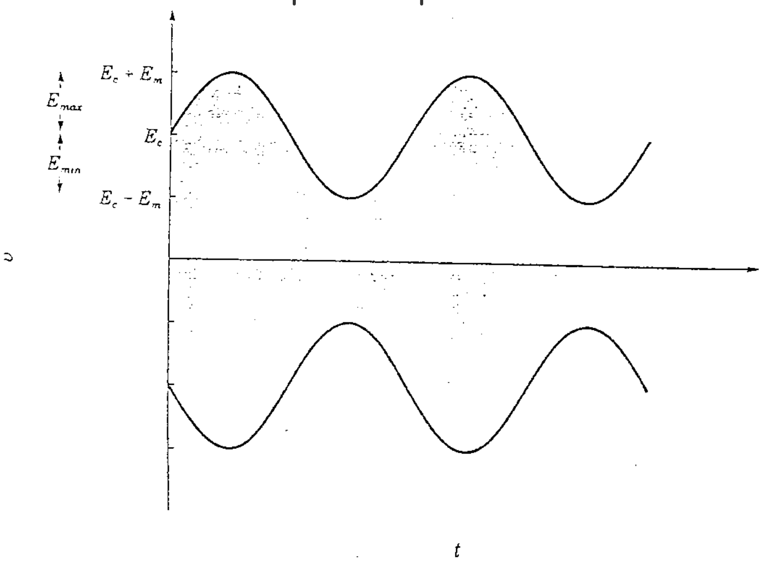
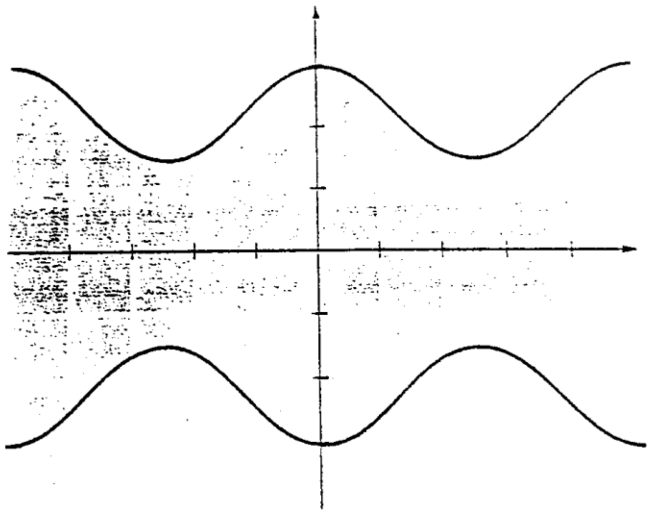


Fig. 7-8 Voltage relationship in an AM signal

Example 7.4: Calculate the modulation index for the waveform shown in Fig.7-9.



Vertical: 25 mV/division

Horizontal: 200 μs/division

Fig. 7-9 AM modulation

7-8 Frequency Domain Analysis for AM Modulation

In order to find out more about this signal, however, it is necessary to consider its spectral makeup. We could use Fourier methods to do this, but for a simple AM waveform it is easier, and just as valid, to use trigonometry.

To start, we should observe carefully that although both the carrier and the modulating signal may be sine waves, the modulated AM waveform is not a sine wave as shown in Eq.(7.5);

$$V(t) = \underbrace{V_{DC}\sin(2\pi f_c t)}_{Carrier} + \underbrace{\frac{V_m}{2}\cos(2\pi(f_c - f_m)t)}_{Lower\ sideband} - \underbrace{\frac{V_m}{2}\cos(2\pi(f_c + f_m)t)}_{upper\ sideband}$$

When the complete signal is sketched in the frequency domain, as in Fig. 7-10, we see the carrier and two additional frequencies, one to each side called side frequencies.

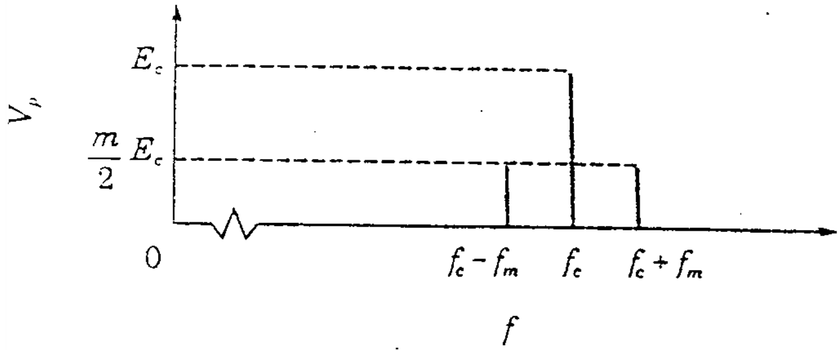


Fig. 7-9 AM in the frequency domain

The separation of each side frequency from the carrier is equal to the modulating frequency, and the relative amplitude of the side frequency, compared with the carrier, is proportional to m, becoming half the carrier voltage for m = 1.

In a realistic situation there will generally be more than one set of side frequencies, because there will be more than one modulating frequency. Each modulating frequency will produce two side frequencies. Those above the carrier can be grouped into a band of frequencies called the upper sideband. Similarly, there will be a lower sideband, which looks like a mirror image of the upper, reflected in the carrier.

From now on, we will generally use the term sideband, rather than side frequency;

$$f_{usb} = f_c + f_m \quad \text{.......................... (7.11)}$$

$$f_{lsb} = f_c - f_m \quad \text{.......................... (7.12)}$$

And

$$V_{lsb} = V_{usb} = \frac{m\,V_{Dc}}{2} \quad \text{.......................... (7.13)}$$

Where;

f_{usb} = frequency of the upper sideband.
F_{lsb} = frequency of the lower sideband.
V_{lsb} = peak voltage of the lower sideband component.
V_{usb} = peak voltage of the upper sideband component.

Example 7.5:

(a) A 1 MHz carrier with an amplitude of 1 V peak is modulated by a 1 kHz signal with m = 0.5. Sketch the voltage spectrum.

(b) An additional 2 kHz signal modulates the carrier with m = 0.2. Sketch the voltage spectrum.

7-9 Bandwidth of an AM signal

Signal bandwidth is one of the most important characteristics of any modulation scheme. In general, a narrow bandwidth is desirable. In any situation where spectrum space is limited, a narrow bandwidth will allow signals to be transmitted simultaneously than will a wider bandwidth.

However, the receiver must have a wide enough bandwidth to pass the complete signal including all the sidebands, or distortion will result. Consequently, we will have to calculate the signal bandwidth for each of the modulation schemes we consider. A glance at Equation (3.10) and Figure 3.10 will show that this calculation is very easy for AM. The signal extends from the lower side frequency, which is at the carrier frequency minus modulation frequency, to the upper side frequency, at the carrier frequency plus the modulation frequency. The difference between these is simply twice the modulation frequency.

If we have a complex modulating signal, with more than one modulating frequency, as in Figure 3.11(b), the bandwidth will be twice the highest modulating frequency.

For telephone-quality voice, for instance, a bandwidth of about 6 kHz would suffice, while a video signal with a 4 MHz maximum baseband frequency would need 8 MHz of bandwidth, if transmitted in this way. (Since a television channel is only 6 MHz wide, we can surmise, correctly, that television must actually be transmitted by a more complex modulation scheme less bandwidth.) Mathematically, the relationship is:

$$B = 2F_m \quad \text{.........................} \quad (7.13)$$

Where;

B = bandwidth in hertz.

F_m = highest modulating frequency in hertz.

Example 7.6: CB radio channels are 10 kHz apart. What is the maximum modulation frequency that can be used if a signal is to remain entirely within its assigned channel?

7-10 Power in an AM Signal

Power is important in any communication system, because the signal to noise ratio at the receiver depends as much on the signal power being large as the noise power being small.

The easiest way to look at the power in an AM signal is to use the frequency domain. We can find the power in each frequency component, then add to get total power.

We will assume that;

1- The signal appears across a resistance R, so that reactive volt-amperes can be ignored.
2- The power required is average power.

Suppose that the modulating signal is a sine wave. Then the AM signal consists of three sinusoids, the carrier and two side frequencies (usually called sidebands), as shown in Eg.7.3, Eq.7.5 and Fig.7-9.

The power in carrier is;

$$e_c(t) = V_{DC}\sin(2\pi f_c t) \quad \text{.........................} \quad (7.14)$$

Since V_{Dc} is the peak carrier voltage, the power Pc appears when the RMS value of this signal appears across a resistance R is simply;

$$P_c = \frac{\left(V_{DC}/\sqrt{2}\right)^2}{R} \quad \text{.........................} \quad (7.15)$$

$$P_c = \frac{{V_{DC}}^2}{2R} \quad \text{.........................} \quad (7.16)$$

The tow side have the same power and appears when the RMS value of the sideband amplitudes appear across a resistance R is simply;

$$P_{lsb} = \frac{{V_{lsb}}^2}{2R} \quad \text{.........................} \quad (7.17)$$

The two sidebands have the same amplitude from Eq. 7.13;

$$V_{lsb} = V_{usb} = \frac{m\,V_{Dc}}{2}$$

By substitute Eq.7.13 in Eq.7.17;

$$P_{lsb} = \frac{\left(\frac{m\,V_{Dc}}{2}\right)^2}{2R} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (7.18)$$

By using some mathematics steps;

$$P_{lsb} = \frac{m^2}{4}\ x\ \frac{\mathrm{V_{DC}}^2}{2R} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (7.19)$$

Then

$$P_{lsb} = \frac{m^2}{4}\ x\ P_c \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (7.20)$$

The same reason will show that;

$$P_{usb} = \frac{m^2}{4}\ x\ P_c \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (7.21)$$

Since the two sidebands have the same power, the total sideband power given by;

$$P_{ts} = \frac{m^2}{2}\ x\ P_c \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (7.22)$$

The total power in the whole modulation signal is just the some of the carrier power and the sidebands power, so it is;

$$P_t = (\frac{m^2}{2}\ x\ P_c) +\ P_c \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (7.23)$$

Or

$$P_t = P_c(1 + \frac{m^2}{2}) \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (7.24)$$

These latest equations tell us several useful things:

1- The total power in an AM signal increases with modulation, reaching a value 50% greater than that of the unmodulated carrier for 100% modulation. As we shall see, this has implications for transmitter design.
2- The extra power with modulation goes into the sidebands: the carrier power does not change with modulation.
3- The useful power, that is, the power that carries information, is rather small, being a maximum of one third of the total sigmal power for 100% modulation, and much less at lower modulation indices. For this reason, AM transmission is more efficient when the modulation index is as close to 1 as practicable.

Example 7.7: An AM broadcast transmitter has a carrier power output of 50 kW. What would be the total power produced with 80% modulation?

7-11 Measuring the Modulation Index in the Frequency Domain

Since the ratio between sideband and carrier power is a simple function of m it is quite possible to measure the modulation index by observing the spectrum of an AM signal. The only slight complication is that spectrum analyzers generally display power ratios in decibels. The power ratio between sideband and carrier power can easily be found from the relation:

$$\frac{P_{lsb}}{P_c} = antilog(\ dB\ ) \text{ .... } (7.25)$$

Where

P_c = carrier power

P_{lsb} = power in one sideband (The lower has been chosen for the example, but of course the upper sideband has the same %3D power.)

dB = difference between sideband and carrier signals, measured in decibels (This number will be negative.)

Once the ratio between carrier and sideband power has been found, it is easy to find the modulation index from Eq.7.20:

$$P_{lsb} = \frac{m^2}{4}\ x\ P_c$$

$$m = 2\sqrt{\frac{P_{lsb}}{P_c}} \text{ .... } (7.26)$$

Fig.7-10 is a handy monograph for finding m directly from the difference in decibels between sideband and carrier powers.

Although the time-domain measurement described earlier is simpler an uses less-expensive equipment, frequency-domain measurement enables much smaller values of m to be found. A modulation level of 5%, for instance would be almost invisible on an oscilloscope, but it is quite obvious, and to measure, on a spectrum analyzer. The spectrum analyzer also allows the contribution from different modulating frequencies to be observed and calculated separately.

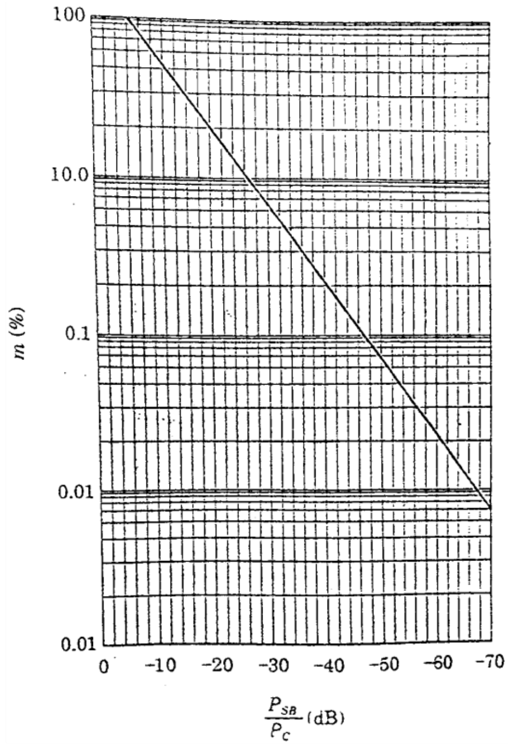


Fig.7-10 modulation index to modulation power

Example 7.8 Calculate the modulation frequency f_m and modulation index m for the spectrum analyzer display shown in Fig. 7-11

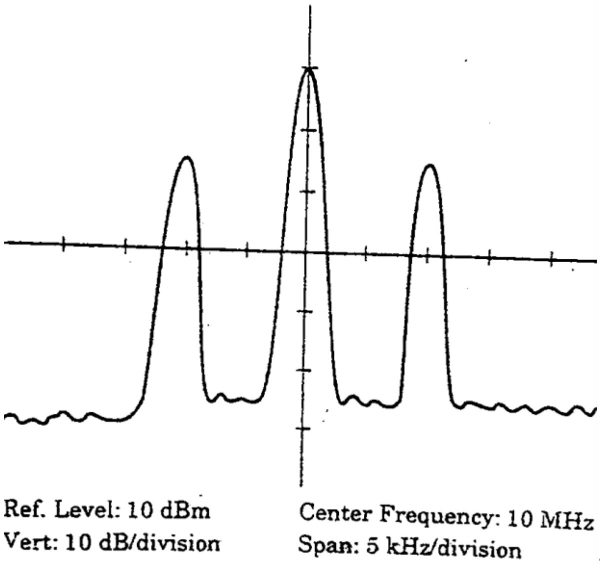


Fig. 7-11 Spectrum analyzer display for AM modulation

7-12 RMS and Peak Voltages and Currents

The well-known power equation

$$P = \frac{V^2}{R}$$

is valid for any kind of waveform, provided that

P = represents average power

V = represents RMS voltage.

To find the RMS voltage when the power is known, rearrange this equation to get;

$$V = \sqrt{PR} \text{ .... (7.27)}$$

The total power in an AM waveform was found earlier in Eq.7.24 substituting this into Eq.7.27 gives;

$$V = \sqrt{P_c(1 + \frac{m^2}{2})\, R} \text{ .... (7.28)}$$

Similarly, from the fundamental equation

$$P = I^2 R$$

we can easily find the RMS current:

$$I = \sqrt{\frac{P}{R}} \text{ .... (7.29)}$$

For a modulated signal, the power required is the total power, giving the equation;

$$I = \sqrt{\frac{P_c(1+\frac{m^2}{2})}{R}} \text{ .... (7.30)}$$

Example 7.9 Find the RMS voltage and current at the output of a transmitter that has a carrier power of 500W into a load impedance of 50ohm:

(a) Without modulation

(b) With 70% modulation

Sometimes an RF ammeter in the transmission line from transmitter to antenna is used to measure modulation index for an AM station. From Equation (7.30) we know that:

$$I = \sqrt{\frac{P_c(1+\frac{m^2}{2})}{R}}$$

We can rearrange this equation to express m in terms of the current:

$$1 + \frac{m^2}{2} = \frac{I^2 R}{P_c}$$

$$\frac{m^2}{2} = \frac{I^2 R}{P_c} - 1$$

$$m = \sqrt{2\left(\frac{I^2 R}{P_c} - 1\right)} \quad \text{.......................... (7.31)}$$

In this equation, of course, I is the current with modulation, as measured by the RF ammeter. The ammeter will not measure P_C directly, but we do know that

$$P_C = I_0^2 R \quad \text{.......................... (7.32)}$$

I_0 = current without modulation

If we substitute this into Eq.7.31 for m, we get

$$m = \sqrt{2\left(\frac{I^2 R}{I_0^2 R} - 1\right)}$$

R cancels out, so we find

$$m = \sqrt{2\left(\frac{I^2}{I_0^2} - 1\right)} \quad \text{.......................... (7.33)}$$

Example 7.10 An RF ammeter in the transmission line from transmitter to antenna measures 5A without modulation and 5.5A with modulation. What is the modulation index?

Now, how about peak values for voltage? We may be tempted to use the formula

$$V_p = \sqrt{2}\, V_{RMS} \text{ ……………………… (7.33)}$$

Let us go back to Eq.7.8 and Eq.7.6;

$$V_{max} = V_{Dc} + V_m$$

$$m = \frac{V_m}{V_{DC}} \Rightarrow V_m = m\ V_{DC}$$

So;

$$V_{max} = V_{Dc}(1 + m) \text{ ……………………… (7.34)}$$

Thus, we can find the peak voltage of an AM signal by first finding the peak carrier voltage and the modulation index.

Example 7.11 Find the peak voltage at the output of a CB transmitter with 4W carrier power output, operating into a 50ohm resistive load, with m = 0.8.

Example 7.12 An AM transmitter has a carrier power of 1 kW at a carrier frequency of 10 MHz. It operates into an antenna with an impedance of 50 ohm resistive. It is modulated with m = 0.5, by a sine wave with a frequency of 1 kHz.

(a) Find:

(i) the signal bandwidth

(ii) the total signal power

(iii) the RMS voltage of the signal

(iv) the peak signal voltage

(b) sketch the signal in the time domain

(c) sketch the signal in the frequency domain

7-13 QUADRATURE AM AND AM STEREO

It is possible to send two separate information signals using amplitude modulation at one carrier frequency. This can be accomplished by generating two carriers, both at the same frequency but separated in phase by 90°. Then, each is modulated by a separate information signal and the two resulting signals are summed.

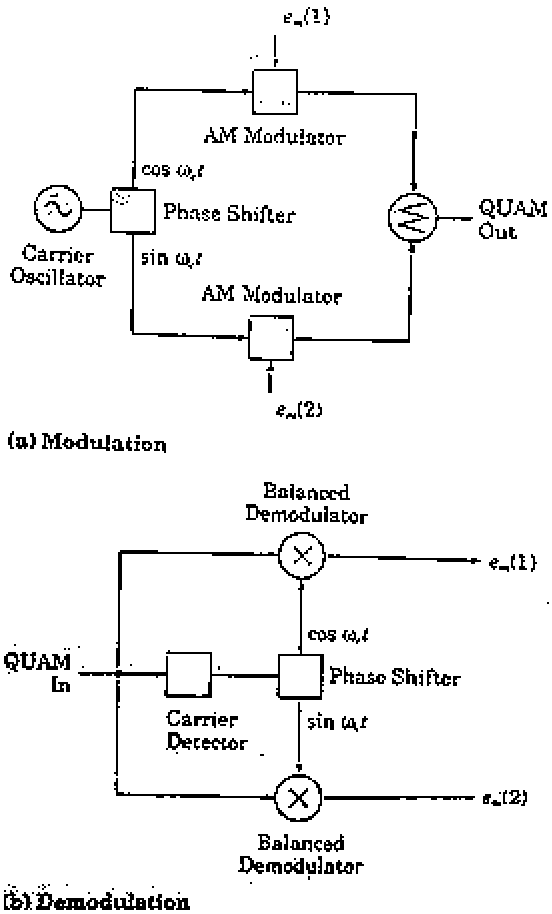


Fig. 7-12 Quadrature AM

Because of the 90° phase shift involved, the scheme is called quadrature AM (QUAM or QAM). Fig. 7.12(a) shows how it can be implemented.

Recovery of the two information signals requires synchronous detection using two balanced demodulators. These must be supplied with reference carriers having exactly the same frequency and phase as the originel carriers. Otherwise the output from each detector will be some combination of the two baseband signals.

Figure 7.12(b) gives the general idea Balanced demodulators are essentially the same as the balanced mixers, their operation as detectors.

8- AM Transmitters

8-1 Transmitter Requirements

The signal must be generated with sufficient power, at the right frequency, with reasonable efficiency and be coupled into an antenna.

1- **Frequency Accuracy and Stability**
The accuracy and stability of the transmitter frequency are essentially fixed by the oscillator. Depending on the application, frequency accuracy and stability may be specified in hertz or as a percentage of the operating frequency.

Example 8.1 A crystal oscillator is accurate within 0.005%. How far off frequency could its output be at 27 MHz?

2- **Frequency Agility**
Frequency agility refers to the ability to change operating frequency rapidly, without extensive retuning. In a broadcast transmitter, this is not a requirement, since such stations rarely change frequency.

3- **Spectral Purity**
All transmitters produce spurious signals. That is, they emit signals at frequencies other than those of the carrier and the sidebands required for the modulation scheme in use. Spurious signals are often harmonics of the operating frequency, or of the carrier oscillator if it operates at a different frequency. The filtering of harmonics in modern, well-designed transmitters it is very effective.

4- **Power Output**
AM transmitters are generally rated in terms of carrier power output. Sometimes dc power input to the RF power amplifier stage has been used, because it is easier to measure. However, different amplifier efficiencies result in a considerable difference in output power for the same power input, so output power is a more useful rating.

5- **Efficiency**

Transmitter efficiency is important for two reasons.

A- The most obvious one is energy conservation. This is especially important where very large power levels are involved, as in broadcasting, or, at the other extreme of the power-level range, where hand-held operation from batteries is required.

B- Another reason is the power that enters the transmitter from the power supply but does not exist via the antenna: it is converted into heat in the transmitter, and this heat must be dissipated.

6- **Modulation Fidelity**

An ideal communications system allows the original information signal to be recovered exactly, except for a time delay. Any distortion introduced at the transmitter is likely to remain; in most cases it will not be possible to remove it at the receiver.

8-2 Typical Transmitter Topologies

AM transmitters can take several different forms, as shown in block diagrams bellow.

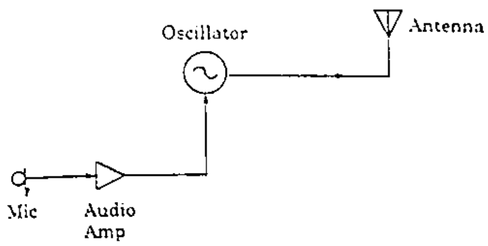


Fig. 8-1 Modulated oscillator (AM transmitter)

In Fig.8-1 an oscillator is modulated and connected directly to an antenna.

This is not a practical circuit for several reasons;

1- Oscillators are susceptible to frequency changes due to variation of any of the operating conditions of the active device used.

2- Changes in load impedance (such as would be created by ice on the antenna, for example) will also result frequency changes.

3- Oscillators should run at very low power levels to reduce heating effects that can also result in frequency changes.

In spite of all these disadvantages, "transmitters" with this topology are built but only as toys.

A more practical transmitter is shown in Fig.8-2. A Class C amplifier stage has been added to provide isolation and power gain.

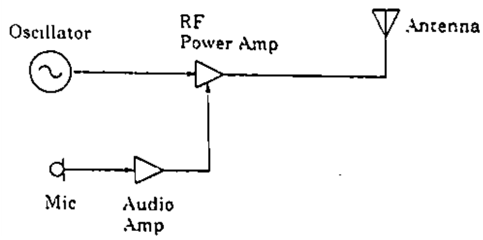


Fig. 8-2 Oscillator with power amplifier (AM transmitter)

Most practical transmitters use more than one stage between oscillator and antenna, as shown in Fig.8-3. The buffer isolates the oscillator from any load changes caused by modulation of the power amplifier, and the driver supplies the power needed at the input to the power amplifier.

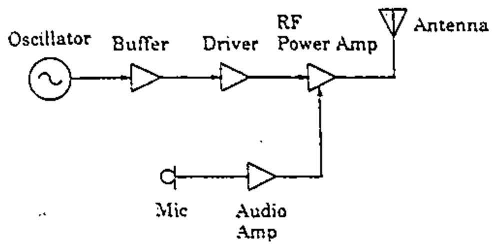


Fig. 8-3 AM transmitter with buffer and driver stage

If an AM transmitter is to operate in the HF or VHF range, or higher in frequency, one or more frequency multipliers will probably be used between the oscillator and the driver, as shown in Fig.8-4. Operation at a relatively low frequency allows for a more stable oscillator design. All of the designs shown so far have the RF power stage modulated.

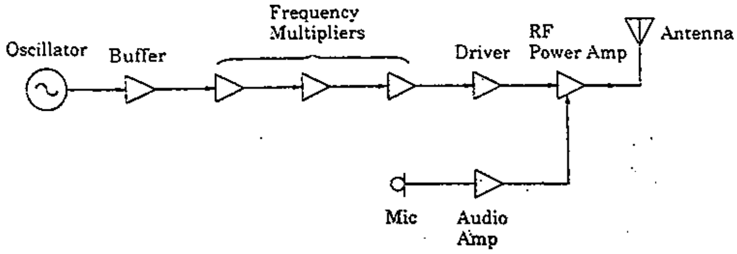


Fig. 8-4 VHF Transmitter with frequency multiplication

8-3 Transmitter Stages

1- The Oscillator Stage

Good stability in modern transmitter design generally requires a crystar controlled oscillator. Where variable-frequency operation is required the usual practice is to use a frequency synthesizer locked to a crystal-controlled master oscillator. It possible to use a variable-frequency oscillator (VFO) in applications (as in the military and amateur services).

2- **The Buffer and Multiplier Stages**
Whatever type of oscillator is used to generate the operating frequency, it must be isolated from any changes in load impedance in order to maintain good stability. The buffer stage accomplishes this. The buffer stage likely to be a wideband amplifier.

3- **The Driver Stage**
Depending on the output power of the transmitter, its power-amplifier stage may require considerable power at its input.
If the power amplifier of a 10 kW transmitter has a power gain of 20 dB, it will need an input power of 100 W. This calls for a power amplifier, probably operating Class C, to drive the final stage. This stage may be referred to as the intermediate power amplifier (IPA).

4- **The Power Amplifier/Modulator**
In small transmitters, a single transistor operating in Class C is likely to be used in the final amplifier, Fig. 8-5 shows a circuit for the modulated power amplifier of a low-powered AM transmitter. This circuit is similar to the AM modulator introduced in Chapter 7, which was in turn based on the Class C amplifier.

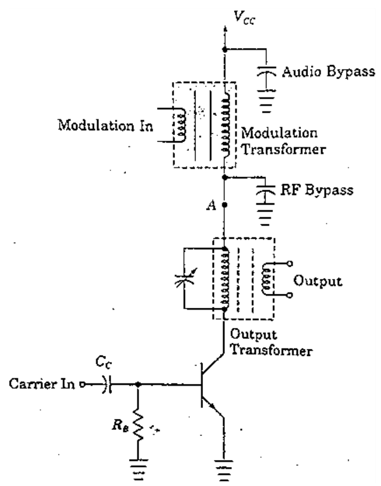


Fig. 8-5 Modulated Class C amplifier

Ideally, for 100% modulation, the voltage at point A, the transistor side of the modulation transformer secondary, will vary between zero and twice the supply voltage, Vcc.

The power provided directly by the supply to the final amplifier does not change with modulation.

Basic AM theory tells us that modulation will increase the power output of the amplifier by 50% for 100% modulation. Therefore, the power input will have to increase by the same percentage.
Where does the extra power come from, if not directly from the supply, Vcc? There is only one other possible source, and that is the modulating amplifier, in this case assumed to be an audio amplifier.

It is apparent that an audio amplifier will have to provide 50% of the dc carrier power input to the final amplifier.

$$P_a = 0.5P_i \text{ ……………………… (8.1)}$$

Where

P_a = power required from the audio amplifier

P_i = dc power input to the final amplifier

Example 8.2 A transmitter has a carrier power output of 10W at an efficiency of 70%. How much power must be supplied by the modulating amplifier for 100% modulation?

The impedance looking from the transformer secondary into the modulated stage is easy to calculate. P_i is the dc power from the V_{CC} supply, so;

$$P_i = V_{cc}\, I_c \text{ ……………………… (8.2)}$$

Where

Vcc = power supply voltage

Ic = average collector current

For 100% modulation, we know from Eq.8.1 so

$$P_a = 0.5V_{cc}\, I_c \text{ ……………………… (8.3)}$$

For 100% modulation, we also know that the voltage from the modulating amplifier, when added to the supply voltage Vcc, must be capable of reducing it to zero on one peak and doubling it on the other. In other words, at the transformer secondary;

$$V_{a(pk)} = V_{cc} \quad \text{.........................} \quad (8.4)$$

Where

V_a = audio voltage measured across the modulation transformer secondary

Assuming the modulating signal is a sine wave,

$$V_{a(RMS)} = \frac{V_{a(pk)}}{\sqrt{2}} \quad \text{.........................} \quad (8.5)$$

$$V_{a(RMS)} = \frac{V_{cc}}{\sqrt{2}} \quad \text{.........................} \quad (8.6)$$

Now, letting Z_a be the impedance seen from the modulation transformer secondary, and using the power equation for RMS voltage and impedance:

$$P = \frac{V^2}{Z} \quad \text{.........................} \quad (8.7)$$

Which can be rearranged to get

$$Z = \frac{V^2}{P} \quad \text{.........................} \quad (8.8)$$

$$Z_a = \frac{V_{a(RMS)}^2}{P_a} \quad \text{.........................} \quad (8.9)$$

$$= \frac{\left(\frac{V_{cc}}{\sqrt{2}}\right)^2}{0.5 V_{cc}\, I_c} \quad \text{.........................} \quad (8.10)$$

$$Z_a = \frac{V_{cc}}{I_c} \quad \text{.........................} \quad (8.11)$$

Example 8.3 A transmitter operates from a 12 V supply, with a collector current of 2 A. The modulation transformer has a turns ratio of 4:1. What is the load impedance seen by the audio amplifier?

6- **Audio circuit**

Audio circuitry is required to amplify the very small signal from a micro-phone, on the order of 1 mV, to a sufficient level to modulate the transmitter. For a large transmitter of 50 kW or more, that can be a considerable amount of power.

8-4 Output Impedance Matching

Most practical transmitters are designed to operate into a 50 ohm resistive load, to match the characteristic impedance of the coaxial cable that is generally used to carry the transmitter power to the antenna.

The output impedance requirements vary dramatically, depending on the type of amplifier. Transistors usually require a load impedance less than 50 ohm.

The simplest way to find the appropriate required load impedance for a Class C amplifier is to note that, for an unmodulated carrier, the collector voltage varies between approximately zero and 2Vcc.

The peak value of the output voltage is;

$$V_{o(pk)} = V_{cc} \quad \text{......................... (8.12)}$$

$$V_{o(RMS)} = \frac{V_{cc}}{\sqrt{2}} \quad \text{......................... (8.13)}$$

According the general power equation;

$$P = \frac{V^2}{R} \quad \text{......................... (8.14)}$$

Since we know P and V and require R, we rearrange this to

$$R = \frac{V^2}{P} \quad \text{......................... (8.15)}$$

$$R_L = \frac{V_o^2}{P_c} \quad \text{......................... (8.15)}$$

Now, substitute the output voltage and power into this equation to get

$$R_L = \frac{\left(\frac{V_{CC}}{\sqrt{2}}\right)^2}{P_C} \quad \text{.........................} (8.16)$$

$$R_L = \frac{{V_{CC}}^2}{2P_C} \quad \text{.........................} (8.17)$$

Where

R_L = load resistance seen at the collector
P_c = carrier power output, without modulation
V_{cc} = collector supply voltage.

Example 8.4 An AM transmitter is required to produce 10 W of carrier power when operating from a 15 V supply. What is the required load impedance as seen from the collector?

For many years, a popular output circuit has been the Pi network, shown in Fig.8-6(a). The pi network can be used to transform impedances either up or down, but is best suited to active devices that require a fairly high load impedance. One common network is shown in Fig.8-6(b) it is called a T network.

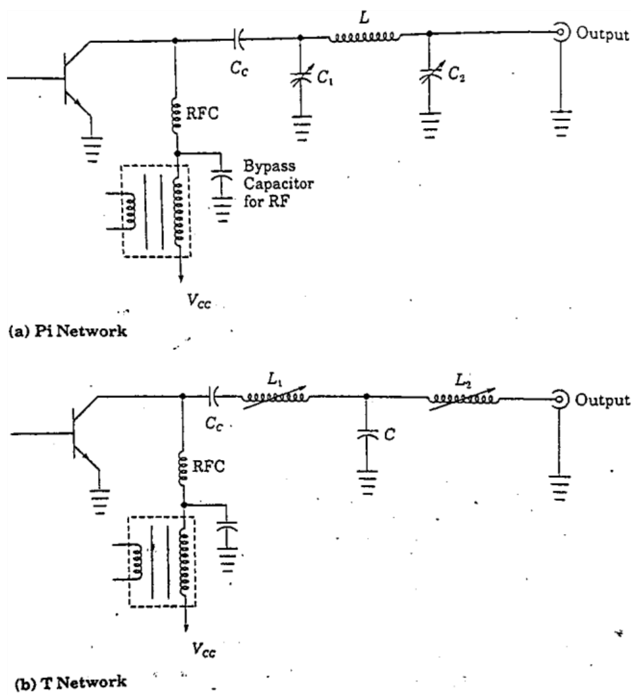


Fig. 8-6 Narrowband AM transmitter output circuits

8-5 An AM Citizens' Band CB transmitter

A CB radio transmitter is always found as part of a transceiver, that is, a combination receiver and transmitter as shown in Fig.8-7. Compared with a separate transmitter and receiver, transceivers are compact, convenient to install and operate, and economical due to the use of some of the same components for both transmit and receive functions. Fig.8-8 shows only the transmitter section of CB transceiver.

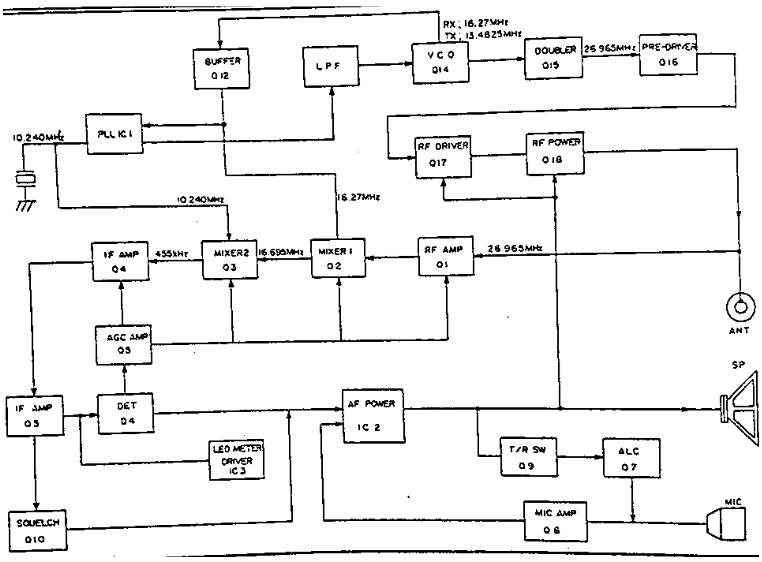


Fig. 8-7 CB transceiver

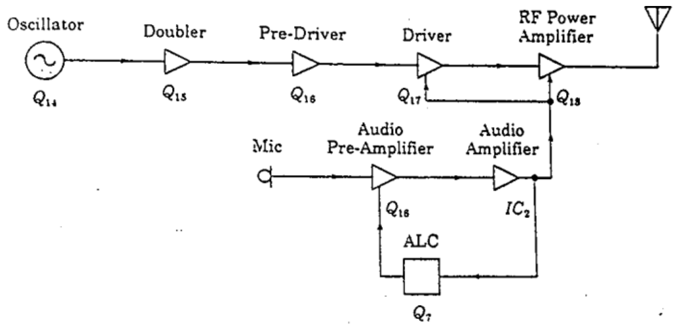


Fig. 8-8 Transmitter section of CB transceiver

8-5 AM Transmitter Measurement of Modulation Percentage

The use of an envelope display on an oscilloscope to measure AM modulation depth technique works quite well in a laboratory situation where the modulating signal is a **sine wave** with constant amplitude and frequency, but it is inconvenient in actual operation where the baseband signal is more likely to be **voice or music**.

The setup shown in Fig.8-9(a) solves some of these problems. This time the oscilloscope is set to x-y mode. The vertical input samples the modulated waveform as before. The horizontal input, however, is connected to the output of the modulating amplifier; that is, it receives the baseband signal just before modulation. This creates a trapezoidal pattern on the screen, as shown in Fig.8-9(b).

The calculation of the modulation index from this pattern is the same for the envelope pattern.

$$m = \frac{E_{max} - E_{min}}{E_{max} + E_{min}} \quad \text{..........................} \quad (8.18)$$

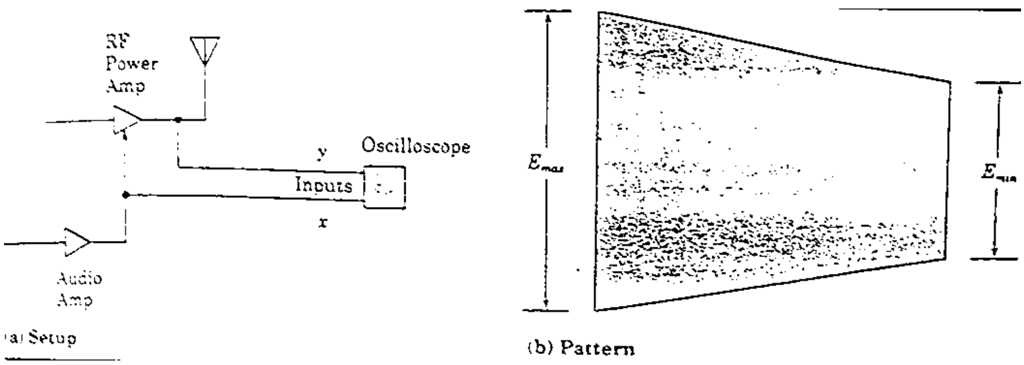


Fig.8-9 Trapezoidal Method of AM Modulation Monitoring

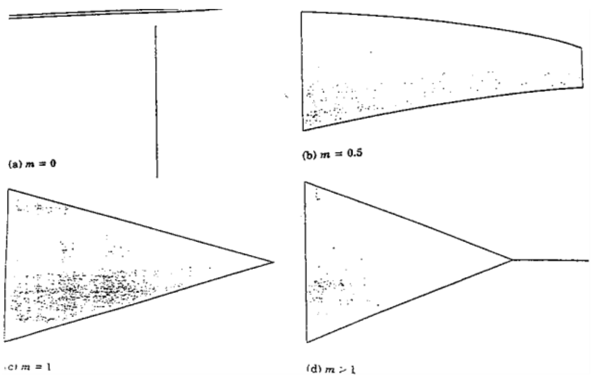


Fig.8-10 Trapezoidal patterns

Example 8.5 Calculate the modulation index for the trapezoidal pattern shown in Fig.8-11.

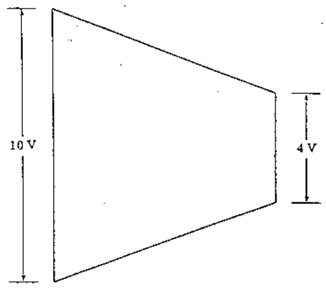


Fig.8-11

9- AM Receivers

The practical receiver would also need means of amplifying the signal and of limiting the bandwidth to reduce noise and interference. The time has now come to see how this can be done.

There are two important specifications that are fundamental to all receivers;

1- **Sensitivity**: is a measure of the signal strength required to achieve a given signal-to-noise ratio
2- **Selectivity**: is the ability to reject unwanted signals at frequencies different from that of the desired signal.

9-1 Receiver Types

Almost all modern receiver designs use the **Superheterodyne** principle. However, we should look first at some more simple methods.

1- **Untuned detector** the simplest receiver would be an envelope detector connected directly to an antenna, as in Fig.9-1(a). Any AM signal arriving at the antenna would be demodulated, and the detector output would be connected to sensitive headphones. There are no sensitivity and selectivity.
2- **Crystal Radios** this receiver could be improved by adding a tuned circuit at the input, as shown in Fig.9-1(b). This would provide some selectivity; that is, the receiver could be tuned to a particular station. Signals at the resonant frequency of the tuned circuit would be passed to the detector, and those at other frequencies would be attenuated and it is still no gain.
3- **Addition of an audio amplifier**, as shown in Fig.9-1(c), could provide enough output power to operate a speaker. However, the selectivity will remain poor because of the single tuned circuit, and the receiver will not he sensitive enough to receive weak signals. This is because a diode detector needs a reiatively large input voltage to operate efficiently, with low noise and distortion.

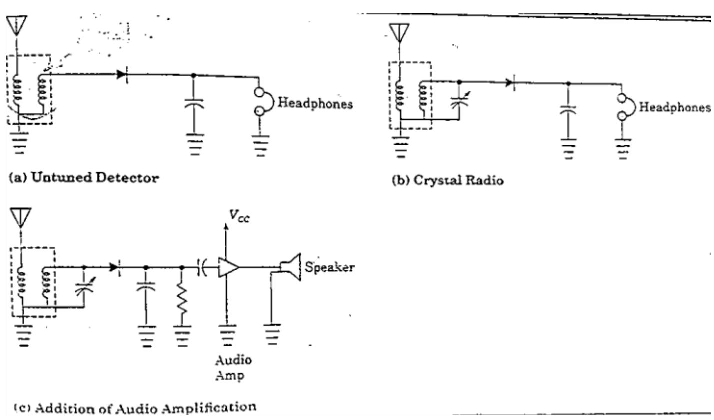


Fig. 9-1 Simple AM Receivers

9-1-1 Tuned-Radio-Frequency Receiver (TRF)

Figure 9-2 shows a block diagram for a tuned-radio-frequency (TRF) receiver. Several RF amplifiers, each tuned to the signal frequency, provide **Gain** and **Selectivity** before the detector. An audio amplifier after the detector supplies the necessary power amplification to drive the speaker.

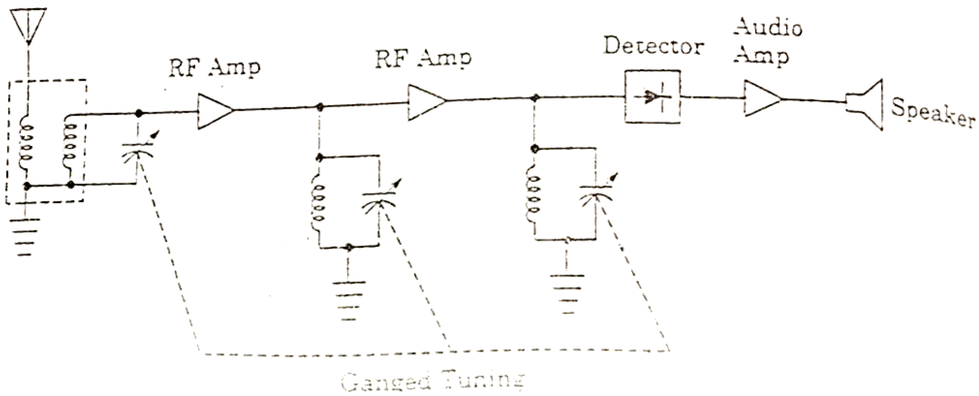


Fig. 9-2 TRF Receivers

The problems with this receiver are with the RF stages. To achieve the suitable gain and selectivity all of the tuned circuits must tune together to the same frequency which tends to cause problems both electrical and mechanical.
Another problem arises from the fact that the bandwidth of a tuned circuit does not remain constant as its resonant frequency is changed, since for an inductor,

$$Q = \frac{X_L}{R} \quad \text{.........................} \quad (9.1)$$

X_L, increases directly with frequency the required Q to achieve a given bandwidth also varies directly with frequency, as can be seen from the equation

$$B = \frac{f_0}{Q} \quad \text{.........................} \quad (9.2)$$

However, in practice the resistance of a coil will also increase with frequency. This is caused by the **skin effect**.
At higher frequencies, internal magnetic fields in the wire cause the current to flow mainly in the region near the surface of the conductor. This decreases the effective cross-sectional area of the conductor, increasing its resistance.
The resistance varies with the square root of frequency. Therefore, the bandwidth of a tuned circuit increases approximately with the square root of frequency.

Example 9.1 A tuned circuit tunes the AM radio broadcast band (from 540 to 1700 kHz). If its bandwidth is 10 kHz at 540 kHz, what is it at 1700 kHz?

Solution

The bandwidth varies with the square root of frequency. Therefore, at the high end the bandwidth will increase to:

$$B = 10kHz * \sqrt{\frac{1700}{540}} \quad \Rightarrow = 17.7kHz$$

Assuming that the bandwidth is correct at the lower end, there may well be interference from adjacent stations at the top end.

9-1-2 The Superheterodyne Receiver

The **superheterodyne receiver or *superhet*** is still almost used in many variations. Fig.9-3 shows its basic layout.

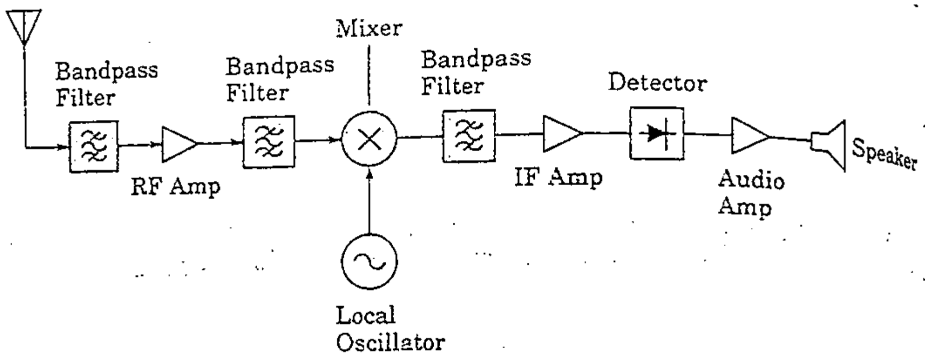


Fig. 9-3 Basic Superheterodyne receiver

There may be one or more stages of RF amplification. The RF stage may be tuned, as in the TRF receiver, or may be broadbanded. This stage shoula have a good noise figure as, being the first stage in the receiver.

The next stage is a mixer. The signal frequency is mixed with a sine-wave signal generated by an associated stage called the local oscillator. A difference frequency is created, which is called the intermediate frequency (IF). The combination of mixer and local oscillator is known as a converter.

Fig. 9-4 shows the signal and local oscillator frequencies for a typical AM broadcast receiver with an IF of 455 kHz. In Figure 5.4(a), the frequency of the desired signal is 740 kHz and the local oscillator is set to 1195 kHz. The difference frequency is 455 kHz, and this is passed to the next stage.

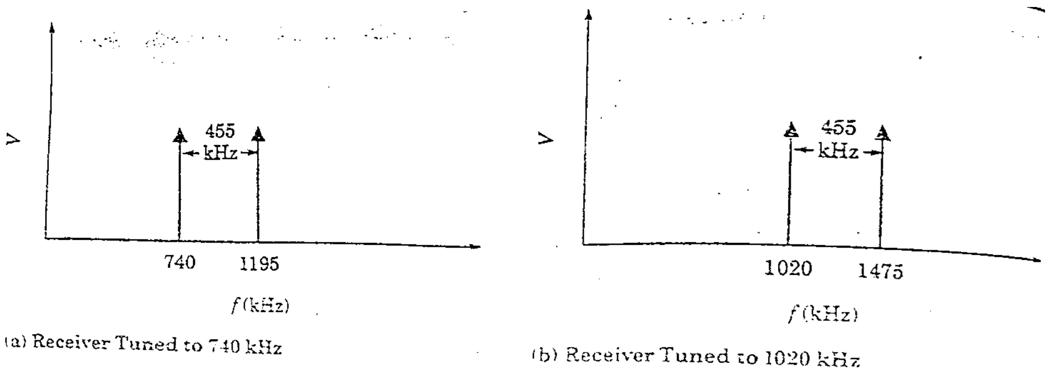


Fig. 9-4 Signal and local oscillator frequencies

The mixer will also produce a sum frequency of 1650 kHz, but this is easily removed by filtering.

9-2 RECEIVER CHARACTERISTICS

Let's look more closely at these characteristics, and to introduce some others.

9-2-1 Sensitivity

The ability to receive weak signals with an acceptable signal-to-noise ratio (S/N) is called sensitivity. One common specification for AM receivers is the signal strength required for a 10 dB signal-plus-noise to noise [(S+N)/N] ratio, at a specified output power level.

9-2-2 Selectivity

The ability to discriminate against interfering signals is known as selectivity.

Selectivity can be expressed as;

1- The bandwidth of the receiver at two different levels of attenuation can be specified. The bandwidth at the points where the signal is 3 or 6 dB down is helpful in determining whether all the sidebands of the desired signal will be passed without attenuation.
2- To indicate the receiver's effectiveness in rejecting interference, a bandwidth for much greater attenuation, for example 60 dB, should also be given.

The frequency-response curve for an ideal IF filter would have a square shape, with no difference between its bandwidths at 6 dB and 60 dB down as shown in Fig.9-5(a). The closer the two bandwidths are, the better the design. The ratio between these bandwidths is called the **Shape Factor (SF)**.

$$SF = \frac{B_{-60\ dB}}{B_{-6\ dB}} \quad \text{.........................} \quad (9.2)$$

Where;

SF = shape factor

$B_{-60\ dB}$ = bandwidth at 60 dB down from maximum

$B_{-6\ dB}$ = bandwidth at 6 dB down from maximum

The shape factor should be as close to one (SF ≈ 1) as possible.

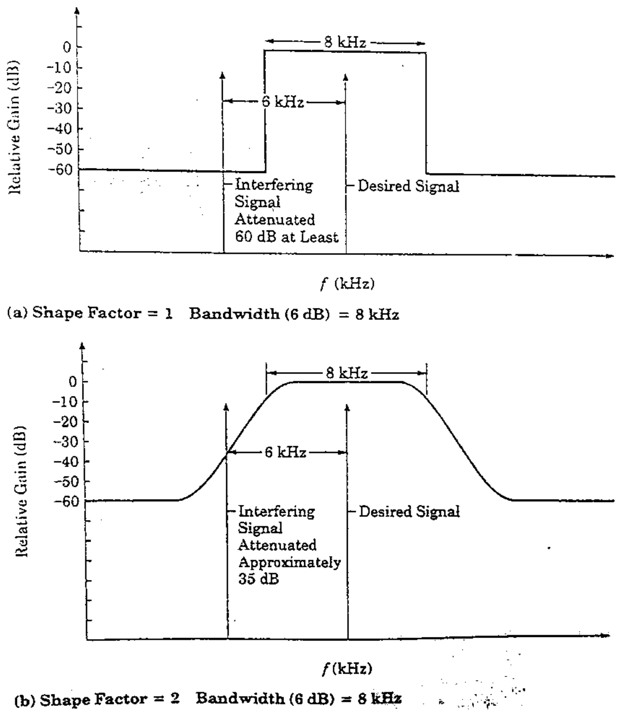


Fig. 9-5 An IF frequency response curve

Example 9.2 Calculate the shape factors for the two IF response curves shown in Fig.9-5, and calculate the amount by which the interfering signal shown would be attenuated in each case.

Solution;

In Fig.9-5(a) is an IF, since the -6dB and -60dB bandwidths are equal and, SF= 1. The interfering signal attenuated by 60db.

In Fig.9-5(b), the –d dB bandwidth is 8kHz and the -60dB bandwidth is 16kHz, so SF=2. The interfering signal is attenuated approximately 35dB compared to the desired signal.

9-2-3 Distortion

An ideal receiver would reproduce the original modulation exactly. A real receiver will subject the signal to several types of distortion as harmonic and intermodulation distortion, uneven frequency response, and phase distortion.

Harmonic occurs when the frequencies generated are multiples of those in the original modulating signal.

Intermodulation takes place when frequency components in the original signal mix in a nonlinear device, creating sum and difference frequencies.

Phase distortion is consist of irregular shifts in phase, and is quite a common occurrence when signals pass through filters is slightly more difficult to understand than frequency response. Of course, the signal at the receiver output will not be in phase with the input to the transmitter: There will be some time delay, which can be translated into a phase shift that increases linearly with frequency.

9-2-4 Dynamic Range

A receiver must operate over a considerable range of signal strengths. The response to weak signals is usually limited by noise generated within the receiver. On the other hand, signals that are too strong will overload one or more stages, causing unacceptable levels of distortion.

So **the dynamic range** of a receiver is the ratio between these two received signal levels (expressed in dB), the too strong and weak signals (P_1/P_2).

The dynamic range in the receiver can be cause the signal **blocking** which is a reduction in sensitivity to the desired signal.

Example 9.3 A receiver has a sensitivity of 0.5 μV and a blocking dynamic range of 70dB. What is the strongest signal that can be present along with a 0.5 μV signal without blocking taking place?

Solution:

$$\frac{P_1}{P_2}(dB) = 20\,log\frac{V_1}{V_2} \quad \Rightarrow \quad \frac{V_1}{V_2} = antilog\frac{\frac{P_1}{P_2}(dB)}{20}$$

We can let V_2 = 0.5 μV

$$V_1 = 0.5 * 10^{-6}\,antilog\frac{70}{20} \quad \Rightarrow \quad V_1 = 1.58\,mV$$

9-2-4 Spurious Responses

In particular, the receiver has a tendency to receive signals at frequencies to which it is not tuned and sometimes to generate signals internally, which can interfere with reception.

Image Frequencies

In any receiver there will be two frequencies one below and one above the local oscillator frequency f_{LO}. These frequencies will mix with f_{LO} to produce the intermediate frequency f_{IF} (below the f_{LO} when high-side injection or above the f_{LO} when low-side injection) and the image frequency f_{image} (above the f_{LO} when high-side injection or below the f_{LO} when low-side injection) as shown in Fig.9-6

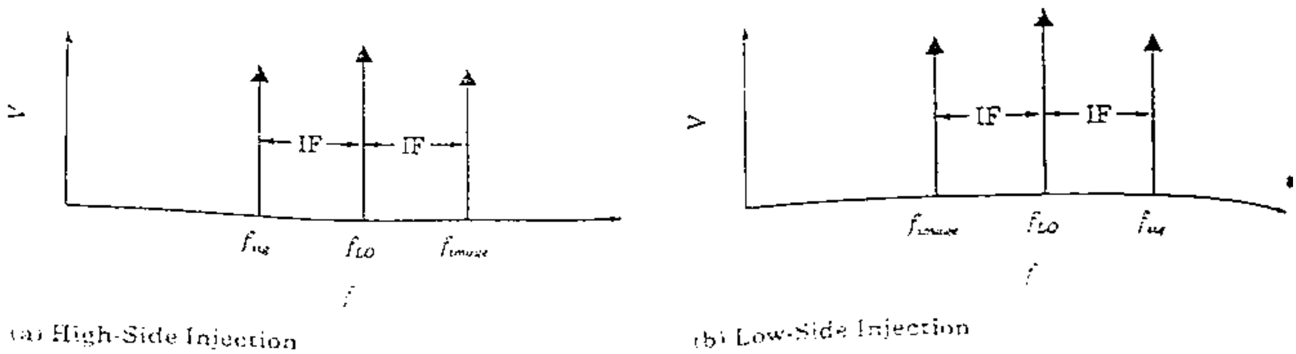


Fig. 9-6 Image response of a receiver

9-3 Receiver Circuits

Now that we have looked at the general structure of a typical superheterodyne receiver, it is time to consider the design of the various stages in a little more detail.

9-3-1 The Radio-Frequency Amplifier

The RF stage is a Class A amplifier. It should have a good noise figura and a wide dynamic range.

To prevent overloading, a switch is sometimes provided to;

1- Remove the RF stage from the signal path for strong signals; such signals are applied directly to the mixer after going through the input filter.
2- To prevent very strong signals from overloading the mixer by adding a few decibels of attenuation.

Fig. 9-7 shows a narrowband bipolar amplifier in a common-base (C.B) configuration. This is quite common in RF amplifiers as;

1- It gives better stability
2- Higher cutoff frequency than a common-emitter circuit.

The variable capacitors must track the receiver's local oscillator tuning.

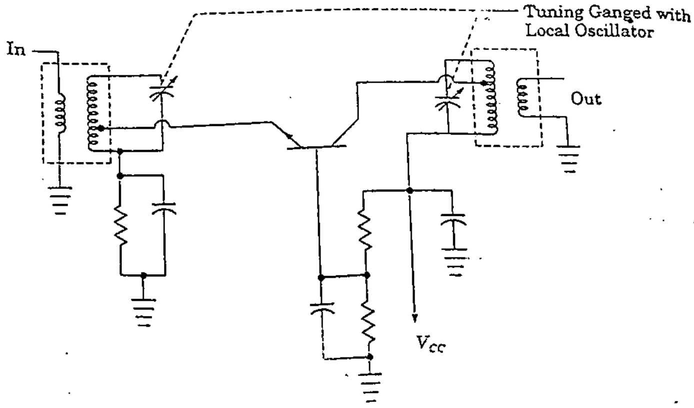


Fig. 9-7 Narrowband C.B bipolar RF amplifier

9-3-2 The Mixer/Converter

The mixer and local oscillator can be combined for economy. The combination is called an autodyne converter or a self-excited mixer. However, better designs use a separate local oscillator. It is extremely important that the local oscillator be stable, as any frequency change will result in the receiver drifting away from the station to which it is tuned.

9-3-3 The Intermediate-Frequency Amplifier

This is basically a linear, fixed-tuned amplifier. The IF amplifier of an AM receiver must be Class A to avoid distorting the signal envelope. The IF amplifier accounts for most of the receiver's gain and selectivity.

The classical method to provide this is to several stages, to give sufficient gain, coupled by tuned transformers that provide the selectivity, Fig.9-10 shows a Bipolar NPN IF amplifier circuit.

9-3-4 The Detector

An envelope detector is essentially a rectifier followed by an RC network as in Fig.9-8. The capacitor C_1, charges to the peak value of the RF waveform, and is slowly discharged by the resistor. If the RC, time constant of C_1 charge and discharge is;

1- Much longer than the period of the RF waveform (AM signal)
2- Much shorter than that of the modulating signal (Information signal

The output will be a reasonably faithful reproduction of the modulating signal. There will be an additional dc component which can easily be removed by a blocking capacitor C_2.

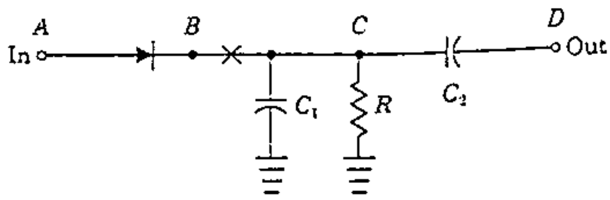


Fig. 9-8 Envelope detector circuit

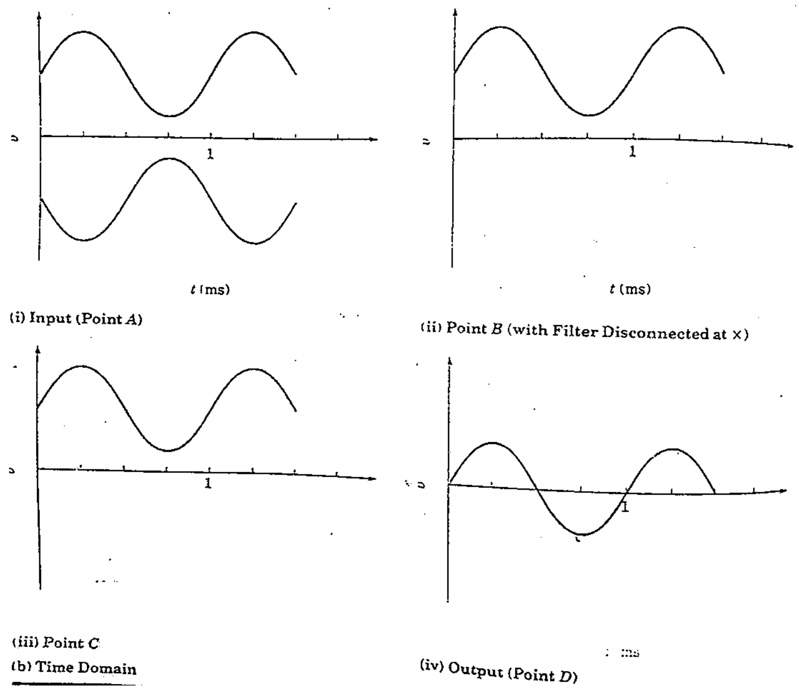


Fig. 9-9 Time domain of signal detection

9-3-5 Automatic Gain Control AGC

Some form of gain control is necessary before the detector to reduce gain with strong signals and prevent overloading. This is usually done automatically using a feedback circuit.

A representative IF amplifier stage, with *AGC* applied, is shown in Fig.9-10. The amplifier circuit uses conventional voltage-divider bias, except that the negative end of the voltage divider is connected to the *AGC* bus instead of to ground. As the signal becomes stronger at the detector. The *AGC* bus becomes more negative, and the voltage at the transistor base becomes lower. This reduces the transistor emitter current, lowering the gain of the stage.

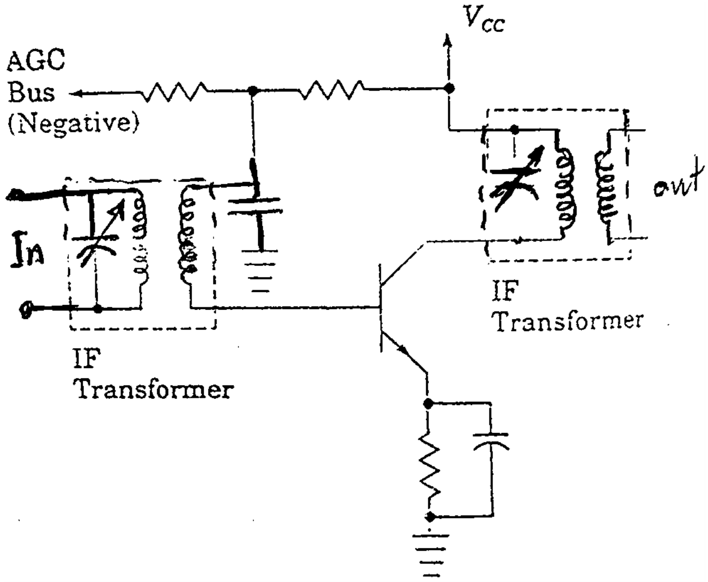


Fig. 9-10 *AGC* to NPN Bipolar IF Amplifier

10- Frequency Modulation FM

Frequency modulation (FM) is modulating of information signal in a carrier signal by varying the instantaneous frequency of the carrier signal according to the amplitude of the information signal.
For example an unmodulated carrier frequency in Fig. 10.1(a) modulate by a square wave modulating signal in Fig. 10.1(a) and results a FM signal as shown in Fig. 10.1(b). The FM signal shows how the amplitude of the carrier signal remains as before, and the frequency varies with time in accordance with the amplitude of the modulating signal.

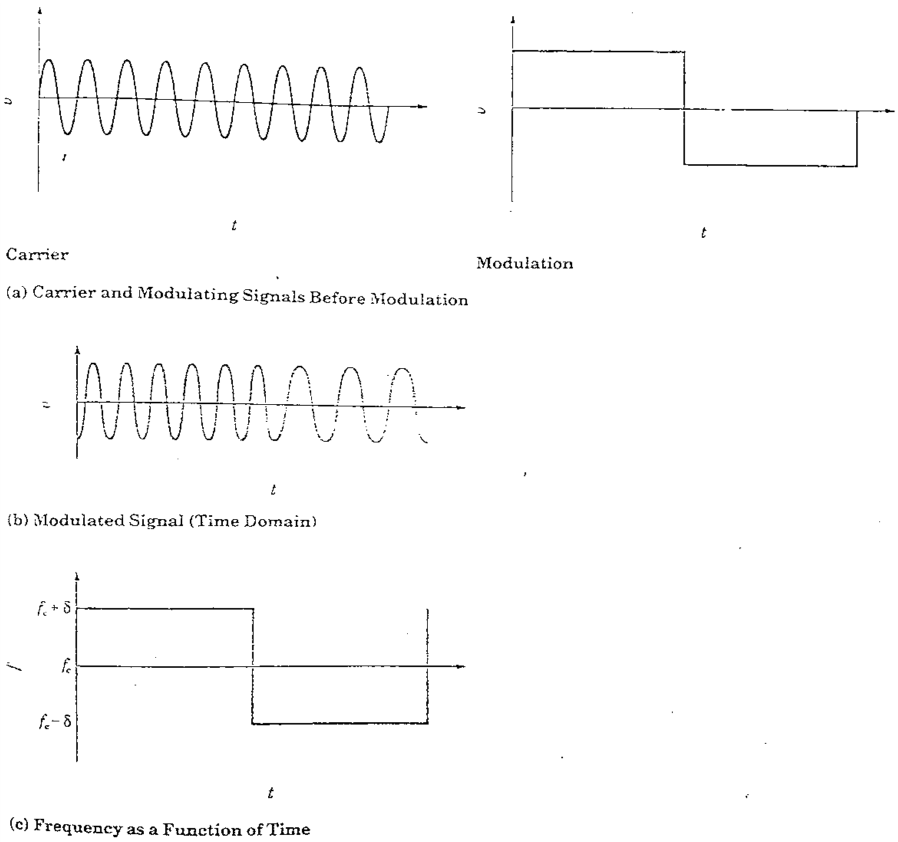


Fig. 10-1 FM of a Sine-wave carrier by a square wave

The simplest method to generate FM is to use a voltage-controlled oscillator (VCO) to generate the carrier frequency, and to apply the modulating signal to the oscillator's control signal input, as indicated in Fig. 10-2.

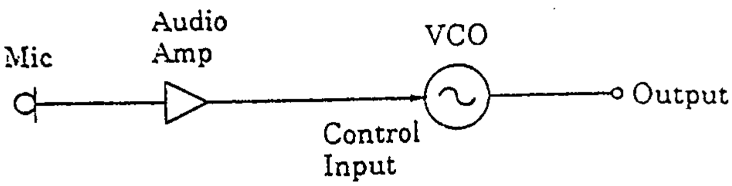


Fig. 10-2 Simple FM generator

10-1 Frequency Deviation

The carrier frequency modulated will cause the signal frequency to vary, or deviate, from its original value. If the modulation system is properly designed, this deviation **Δf** will be proportional to the amplitude of the modulating signal $\mathbf{e_m}$ Fig.10-3 demonstrates this term.

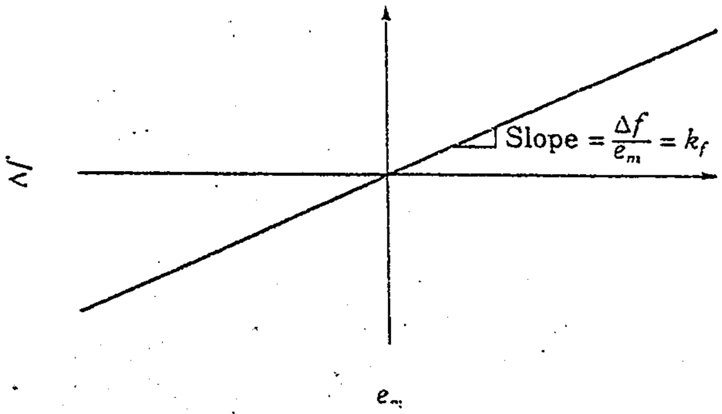


Fig. 10-3 Deviation sensitivity of FM modulator

The deviation sensitivity of the modulator, with units of Hz per volt for line slop of the Fig.10-3 can be given as;

$$k_f = \frac{\Delta f}{e_m} \quad \text{.......................... (10.1)}$$

The frequency deviation is proportional to the amplitude, not the frequency of the modulating signal. The times per second that the frequency varies from its lowest to its high is equal to the modulating-signal frequency.

It is possible to write an equation for the signal frequency as a function of time:

$$f_{sig}(t) = f_c + k_f\, e_m(t) \quad \text{.......................... (10.2)}$$

Where;

$f_{sig}(t)$ = signal frequency as a function of time
f_c = unmodulated carrier frequency
k_f = modulator deviation constant or modulation sensitivity
$e_m(t)$ = modulating voltage as a function of time

If the modulating signal is a sine wave with the equation

$$e_m(t) = E_m \sin w_m\, t \quad \text{.......................... (10.3)}$$

Then Eq.(10.2) becomes;

$$f_{sig}(t) = f_c + k_f\ E_m \sin w_m\ t \quad \text{………………………. (10.4)}$$

and the peak frequency deviation, each side of the carrier frequency, will be ($k_f E_m$) Hz then;

$$\delta = k_f\ E_m \quad \text{………………………. (10.5)}$$

Where;

δ = peak frequency deviation in hertz
k_f = modulator sensitivity in hertz per volt
E_m = peak value of the modulating signal in volts

So the Eq.(10.4) can be written in δ as;

$$f_{sig}(t) = f_c + \delta \sin w_m\ t \quad \text{………………………. (10.6)}$$

Example 10.1 An FM modulator has k_f = 30 kHz/V, and operates at a carrier frequency of 175 MHz. Find the output frequency for an instantaneous value modulating signal equal to:

a) 150 mV b) -2 V

Solution:

Eq.(10.2) can be used for both parts of the question.

$$f_{sig}(t) = f_c + k_f\ e_m(t)$$

a)

f_{sig} = $(175 * 10^6) + (30 * 10^3)(150 * 10^{-3})$

= 175.0045 MHz

b)

f_{sig} = $(175 * 10^6) + (30 * 10^3)(-2)$

= 174.94 MHz

Example 10.2 The same FM modulator as in the previous example 10.1 is modulated by a 3V sine wave. Calculate the deviation.

Solution:

Unless otherwise stated, ac voltages are assumed to be RMS. On the other hand, δ is a peak value. Therefore, the modulating voltage must be converted to a peak value before Equation (8.5) can be used.

$E_m = 3\sqrt{2} = 4.24$ V

$\delta = k_f E_m$

$= (30 * 10^3) * 4.24$

$= 127.2$ kHz

10-2 Frequency Modulation Index

Frequency modulation index m_f to be confused with f_m, which is the modulating frequency. By definition, for sine-wave modulation;

$$m_f = \frac{\delta}{f_m} \quad \text{.......................... (10.7)}$$

Note frequency modulation index m_f, unlike the amplitude modulation index which cannot exceed one, there are no theoretical limit on m_f. It can exceed one, and often does.

By substituting Eq.(10.7) in Eq.(10.6) gives;

$$f_{sig}(t) = f_c + m_f\, f_m \sin w_m\; t \quad \text{.......................... (10.8)}$$

As an equation for the frequency of an FM signal with sine-wave modulation.

Example 10.3 An FM broadcast transmitter operates at its maximum deviation of 75 kHz. Find the modulation index for a sinusoidal modulating signal with a frequency of:

a) 15 kHz b) 50 Hz

Solution:

a) $m_f = \frac{\delta}{f_m} = \frac{75KHz}{15KHz} = 5$

b) $m_f = \frac{\delta}{f_m} = \frac{75*10^3 Hz}{50Hz} = 1500$

10-3 FM Transmitters

The processes of modulation and demodulation are very different for FM than for AM. In addition, there is more variety in the circuits used for the modulation and demodulation of FM than there is for AM.

Figure 10-4 is a simplified block diagram of a typical FM transmitter and it is divided into two sections, the exciter and the power amplifier.

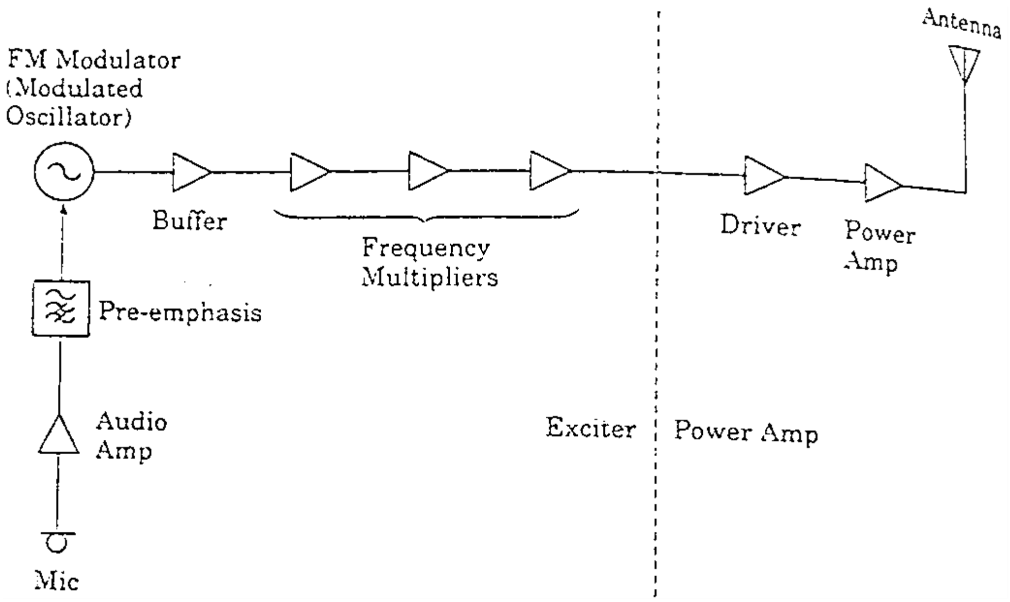


Fig. 10-4 Deviation sensitivity of FM modulator

10-3-1 Direct FM modulator

FM signals can be generated directly, by varying the frequency of the carrier oscillator.

Direct FM requires that the frequency of the carrier oscillator be varied in accordance with the instantaneous amplitude of the modulating signal (after any required pre-emphasis has been added). The ***reactance modulator*** is the simplest way to do this.

One common way to build a reactance modulator is to put a Varactor into the frequency determining circuit of the carrier oscillator.

The circuit in Fig. 10-5 is a Clapp oscillator;

- The frequency is determined mainly by L and the varactor.
- Capacitor C_3 has a large value to isolate the varactor from Vcc
- C_1 and C_2 determine the feedback fraction
- The tuning potentiometer adjusts the dc bias on the varactor.

As for the crystal-controlled version in Fig.10-6, its frequency changed only very slightly by modulation: usually only a few tens of Hz.

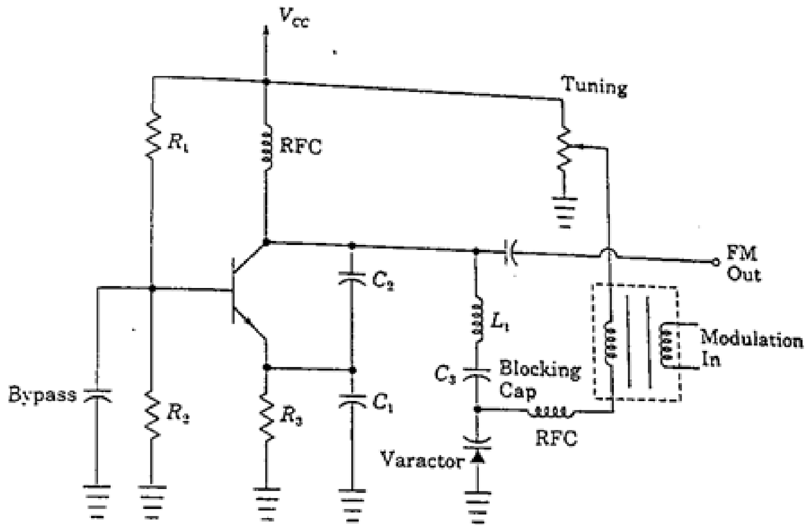


Fig. 10-5 Direct FM Varactor modulator using an LC oscillator

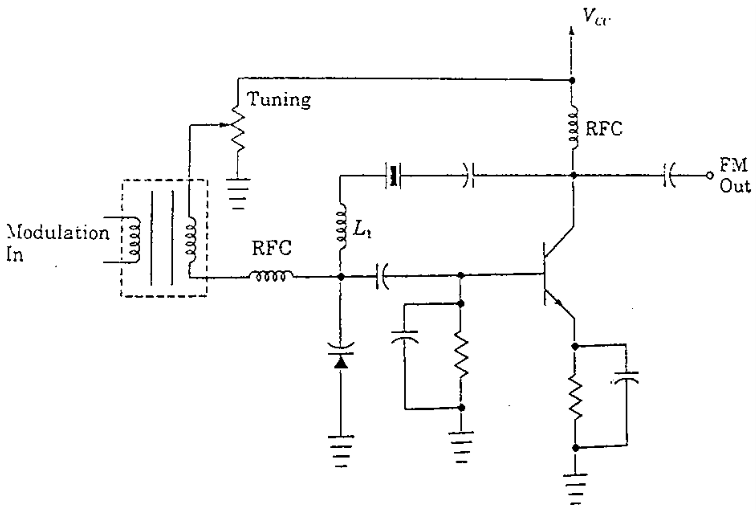


Fig. 10-6 Direct FM Varactor modulator using a Crystal-control oscillator

10-3-2 Frequency Multipliers

Frequency amplifier is a Class C amplifier tuned to the input signal harmonics to increase the frequency in the output.

Doubles and triplets stages are the most common types of frequency multiplier.

If the frequency of modulation input signal changes by an amount *N*, the output signal will have frequency;

$$f_o = N(f_c + \delta) \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (10.9)$$

$$f_o = Nf_c + N\delta \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (10.10)$$

Thus a frequency multiplier can increase the carries frequency f_c and the deviation δ obtained at the modulator by any required amount.

Example 10.4 A direct-FM transmitter has a varactor modulator with k_f = 2 kHz/V and a maximum deviation of 300 Hz. This modulator is followed by a buffer and three stages of frequency multiplication: a tripler, a doubler, and another tripler, followed by a driver and power amplifier.

1) Draw a block diagram of this transmitter.
2) Will this transmitter be capable of 5 kHz deviation at the output?
3) What should the oscillator frequency be if the transmitter is to operate at a carrier frequency of 150 MHz?
4) What audio voltage will be required at the modulator input to obtain full deviation?

Solution:

10-3-3 Automatic Frequency Control Systems (AFC)

A stability is a serious problem with such oscillators, and some form of AFC is always required. The way to do this is known as the **Crosby system**, as in Fig.10-7 essentially, the transmitter carrier frequency is mixed with a crystal-controlled reference signal. The difference frequency is sent to a frequency discriminator

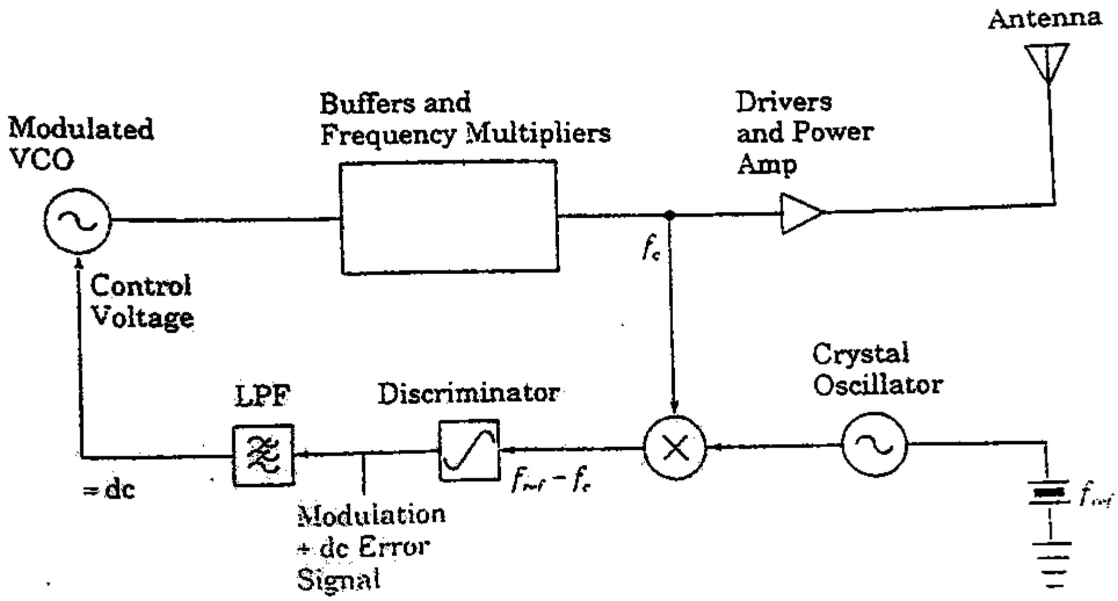


Fig. 10-7 Crosby AFC system

The discriminator circuit produces an output voltage proportional to the difference between the frequency at its input and the center frequency to which the circuit has been tuned.

Example 10.5 A Crosby-type FM transmitter has the block diagram of Fig.10-7. The transmitter is to have a carrier frequency of 99.9 MHz. The crystal oscillator has a frequency of 105 MHz. What should be the center frequency of the discriminator?

Solution:

The difference frequency output from the mixer will be

$f_{ref} - f_c$ = 105.0 – 99.9

= 5.1 MHz

This is the frequency to which the discriminator should be tuned.

10-3-4 Phased Locked Loop (PLL) FM generator

More modern way to solve the stability problem is to make the modulator VCO part of a phase-locked loop (PLL) as in Fig.10-8.

The VCO is locked to some multiple of a crystal-controlled reference frequency. The loop filter will allow the system to ignore the rapid variations of frequency associated with modulation, while preventing drift of the center frequency of the VCO away from its nominal value.

Unlike the Crosby AFC system, the PLL FM transmitter is capable of locking the transmitting frequency exactly to the crystal-controlled reference frequency. With a well-designed VCO, the PLL system is also capable of producing wideband FM without frequency multiplication. These advantages make the use of PLLs very popular in new designs for both communications and broadcast transmitters.

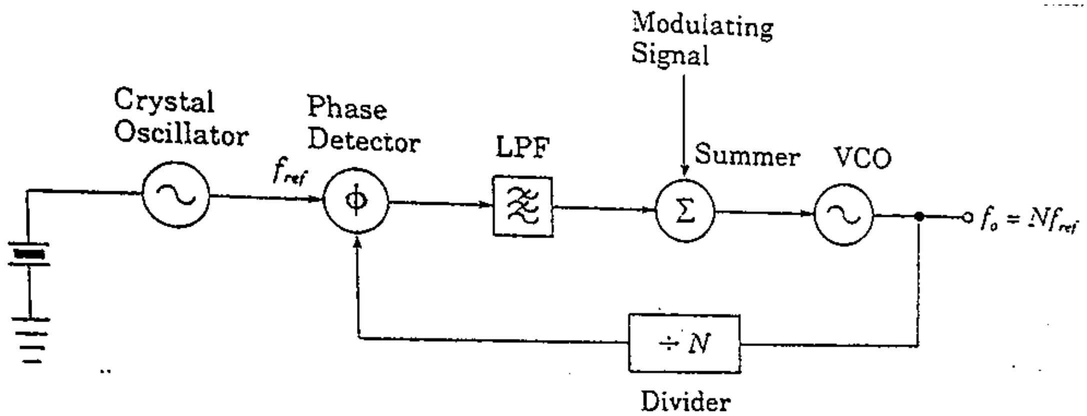


Fig. 10-8 PLL FM transmitter

The carrier frequency of the output signal and the modulator sensitivity with this system is determined by the VCO.

$$f_O = N\, f_{ref} \quad \text{.........................} \ (10.11)$$

$$k_f = \frac{\Delta f}{\Delta v} \quad \text{.........................} \ (10.12)$$

Where:

f_O = Carrier frequency of output signal
N = Divider constant
k_f = VCO sensitivity in Hz per volt
Δf = change in frequency
Δv = change in control voltage required for the change in frequency

Example 10.6 A PLL FM generator has the block diagram of Fig.10-8, with f_{ref} = 100 kHz, N = 200, and k_f= 50 kHz/V.
(a) Calculate the carrier frequency of the output signal.
(b) What RMS modulating voltage will be required for a deviation of 10 kHz at the carrier frequency?

Solution:

a) $f_O = N\, f_{ref}$

= 200 x 100 kHz = 20MHz

b) $k_f = \frac{\Delta f}{\Delta v}$

$\Delta v = \frac{\Delta f}{k_f}$

$= \frac{10kHz}{50\ KHz/V}$ = 0.2v

10-4 FM RECEIVERS

10-4-1 Demodulators

The FM demodulators, to demodulate FM properly, the amplitude of the output must be proportional to the frequency deviation of the input.

This results in a characteristic **S-curve** for many FM detectors (see Fig.10-9). The output voltage is proportional to frequency deviation over a range at least equal to 2δ, since the deviation is the distance the signal frequency moves above and below the carrier frequency.
The sensitivity of an FM detector can be given as;

$$k_d = \frac{V_o}{\delta} \quad \text{..........................} \ (10.13)$$

Where;

k_d = detector sensitivity in volts per hertz
V_o = output voltage
δ = frequency deviation required for the output voltage

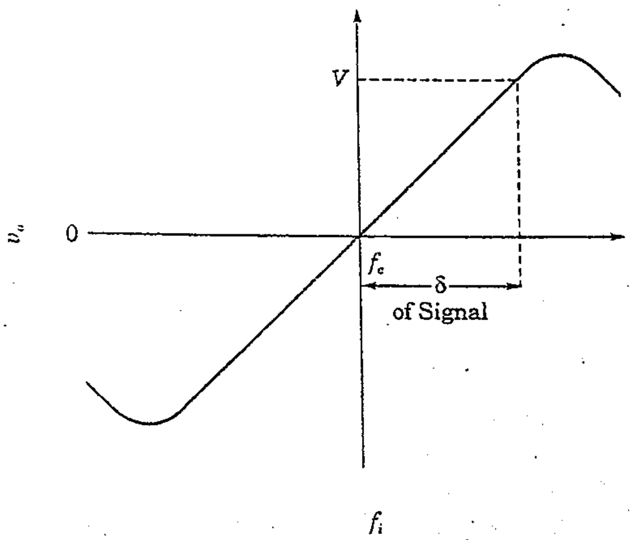


Fig. 10-9 S-curve characteristic of FM detector

Note the sensitivity of a detector is the slope of the straight-line portion of the S-curve of Figure 9.12.

Example 10.7 An FM detector produces a peak-to-peak output voltage of 1.2 V from an FM signal that is modulated to 10 kHz deviation by a sine wave. What is the detector sensitivity?

Solution:

First, the peak-to-peak voltage must be changed to peak, since that is the way deviation is specified.

$$V_o\ peak = \frac{V_o\ peak\ to\ peak}{2}$$
$$= \frac{1.2}{2} = 0.6\ Volt$$

Now the detector sensitivity is

$$k_d = \frac{V_o}{\delta}$$
$$= \frac{0.6 volt}{10 Khz} \quad = 60\ \mu\ V/Hz$$

10-4-2 FM Demodulator Types

There are three major type of FM detector.

1- **Foster-Seeley Discriminator and Ratio Detector**

These circuits convert frequency changes first to phase shifts and then to amplitude variations. The resulting AM signal is then demodulated by a combination of two diode detectors.

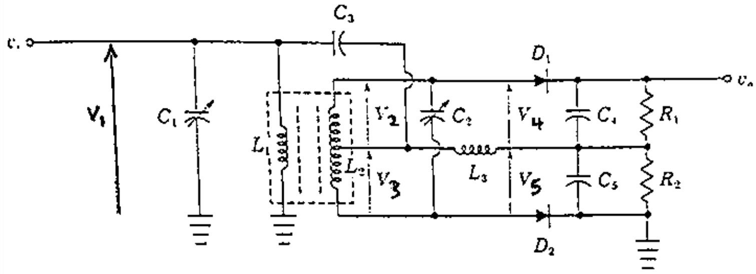


Fig. 10-10 Foster-Seeley Discriminator circuit

The conversion from frequency to phase modulation is achieved by using the fact that the phase angle between voltage and current in a tuned circuit will change as the applied frequency goes through resonance.

As shown in Fig.10-10,

1- The transformer is double-tuned, with both the primary (L1-C1) and secondary (L2-C2) circuits resonant at the carrier frequency. The secondary voltage at resonance is 90° out of phase with the primary voltage.

2- The primary voltage is also applied, in phase, to the center of the secondary winding through capacitor C3 to have low reactance at the carrier frequency.

3- Capacitors C4 and C5 should have low reactance at the carrier frequency, but high reactance at the modulating frequency, so that D_1-C_4 and D_2-C_5 are detectors.

4- Inductor L_3 is an RF choke with high reactance at the carrier frequency and low reactance at the modulating frequency. It can be seen the low-reactance connection made by C_3 causes virtually the entire primary voltage to appear across L_3.

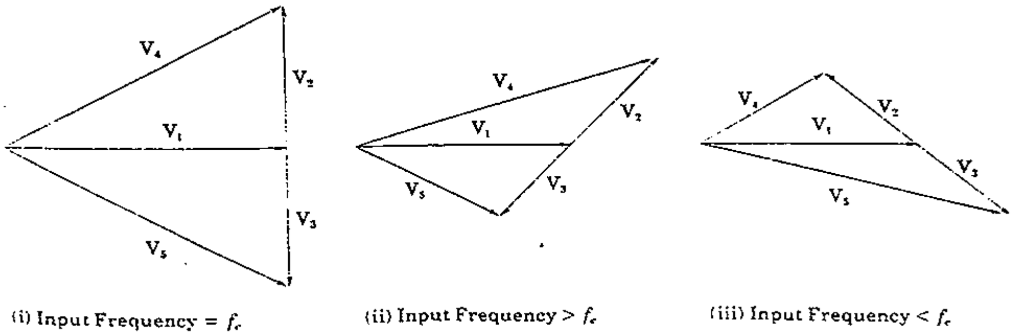


Fig. 10-11 Foster-Seeley Discriminator phasor diagram

Fig.10-11 shows how the primary and secondary voltages add vector. When the incoming signal frequency is equal to the resonant frequency of the tuned circuits, the voltages applied to the two detectors are equal in magnitude. This can be seen by observing that the vectors V_4 and V_5 in Fig.10-11(i) are the same length. Therefore, the net output voltage is zero.

Increases the frequency above resonance frequency, the secondary voltage has a leading phase angle, and the relative length of these vectors changes. Vector V_4 is now greater than V_5 as shown in Fig.10-11(ii). The output voltage becomes positive.

Reducing the frequency below resonance causes a similar but opposite phase change, causing V_4 to be less than V_5 and producing a negative output voltage. The result is an output voltage that follows the modulation.

2- **Phase Locked Loop PLL Method Demodulator**

The incoming FM signal is used to control the frequency of the VCO. When the incoming frequency varies, the PLL will generate a control voltage to change the frequency of VCO, which will follow the incoming signal frequency as shown in Fig.10-12.

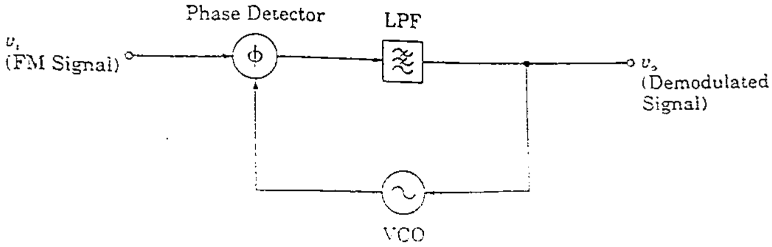


Fig. 10-12 PLL FM Detector block diagram

This control voltage varies at the same rate as the frequency of the incoming signal, and so it can be used directly as the output of the circuit.

It is easy to calculate the output voltage for a PLL detector, provided k_f for the VCO is known. Since the loop stays locked as it follows the modulation, the VCO frequency follows the signal frequency, varying over the range from (f_c- δ) to (f_c+ δ). Therefore, the peak output voltage is the voltage necessary to move the local oscillator by the amount of the deviation δ, that is;

$$V_o = \frac{\delta}{k_f} \quad \text{.........................} \quad (10.14)$$

Where;

V_O = output voltage from the detector

δ = deviation of the signal in hertz

k_f = VCO proportionality constant in hertz per volt.

Example 10.8 A PLL FM detector uses a VCO with k_f = 100 kHz/V. If it is receiving an FM signal with a deviation of 75 kHz and sine-wave modulation, what is the RMS output voltage from the detector?

Solution:

From Equation (10.14),

$$V_o = \frac{\delta}{k_f} \Rightarrow \quad = \frac{75\ kHz}{100\ kHz/v}$$

= 0.75 v

For sine-wave modulation

$$V_{rms} = \frac{V_O\ peak}{\sqrt{2}} \quad \Rightarrow \quad = \frac{0.75}{\sqrt{2}}$$

= 0.53 volt

3- **Quadrature FM Detector**

In the quadrature detector as in Fig. 10-13, the incoming signal is applied to one input of a phase detector. The signal is also applied to a phase-shift network. This consists of a capacitor C_1 with high reactance at the carrier frequency, which will cause a 90° phase shift.

The tuned circuit consisting of L_1 and C_2 is resonant at the carrier frequency. Therefore, it will cause no phase shift at the carrier frequency, but will cause a phase shift at other frequencies that will add (+) to or subtract (-) from the basic 90° shift caused by C_1.

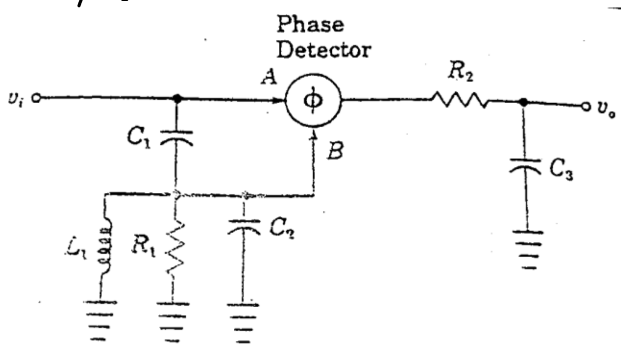


Fig. 10-13 Quadrature FM Detector

The output of the phase-shift network is applied to the second input **B** of the phase detector.

When the input frequency **A** changes, the angle of phase shift in the quadrature circuit will vary, as the resonant circuit becomes inductive or capacitive. The output from the phase detector will vary at the signal frequency but will have an average value proportional to the amount the phase angle differs from 90°. Low-pass (R_3 – C_3) filtering the output will recover the modulation.