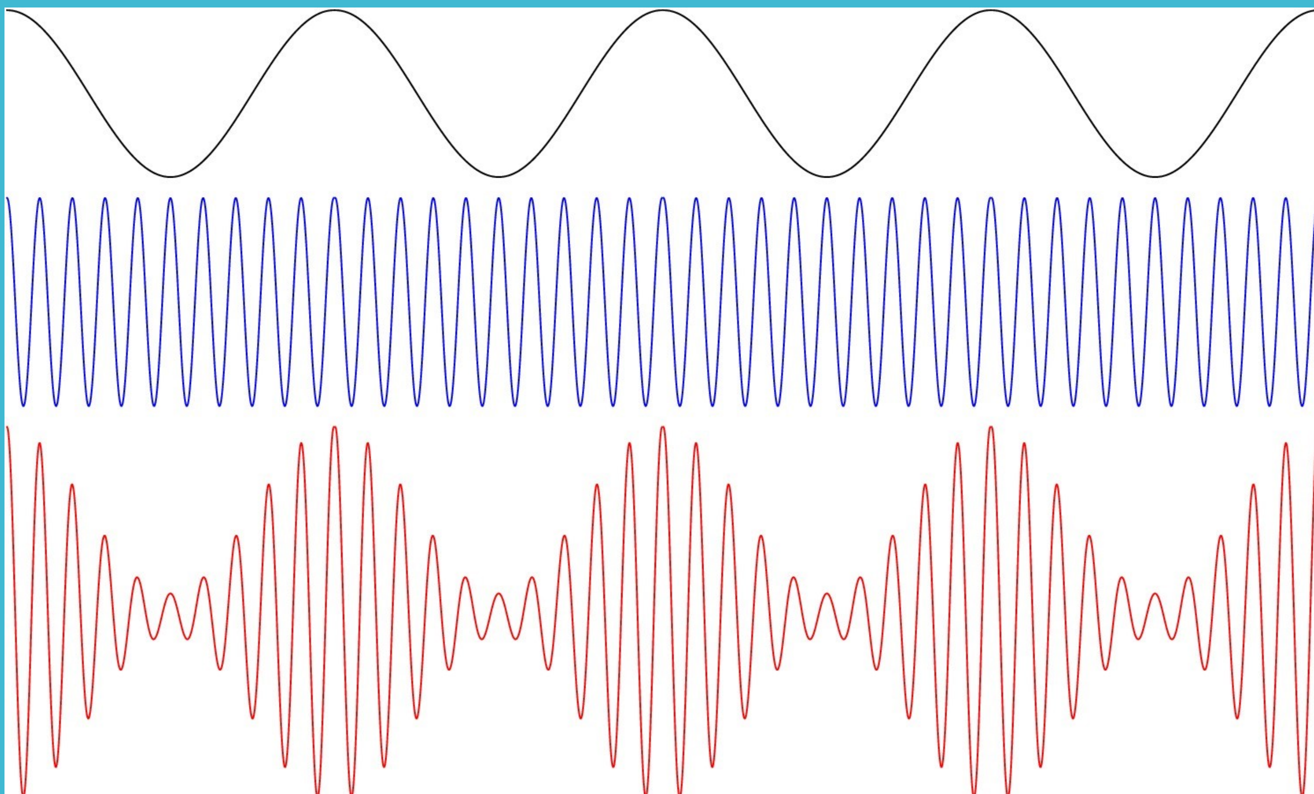


Marwadi University

Department of Information and Communication Technology

Unit no: 1
Unit title: Linear Modulation
Subject name: Analog and Digital Communication
Subject code: 01CT0404

8.5 Frequency Spectrum for Sinusoidal AM

Although the modulated waveform contains two frequencies f_c and f_m, the modulation process generates new frequencies that are the sum and difference of these. The spectrum is found by expanding the equation for the sinusoidally modulated AM as follows:

$$\begin{aligned} e(t) &= E_{c\,\max}(1 + m\cos 2\pi f_m t)\cos 2\pi f_c t \\ &= E_{c\,\max}\cos 2\pi f_c t + mE_{c\,\max}\cos 2\pi f_m t \times \cos 2\pi f_c t \\ &= E_{c\,\max}\cos 2\pi f_c t + \frac{m}{2}E_{c\,\max}\cos 2\pi(f_c - f_m)t + \frac{m}{2}E_{c\,\max}\cos 2\pi(f_c + f_m)t \end{aligned} \tag{8.5.1}$$

It is left as an exercise for the student to derive this result making use of the trigonometric identity

$$\cos(A \pm B) = \cos A\cos B \mp \sin A\sin B \tag{8.5.2}$$

Credit: Electronic Communication by Roddy and Coolen

Equation (8.5.1) shows that the sinusoidally modulated wave consists of three components: a carrier wave of amplitude $E_{c\,\max}$ and frequency f_c, a *lower side frequency* of amplitude $mE_{c\,\max}/2$ and frequency $f_c - f_m$, and an *upper side frequency* of amplitude $mE_{c\,\max}/2$ and frequency $f_c + f_m$. The amplitude spectrum is shown in Fig. 8.5.1.

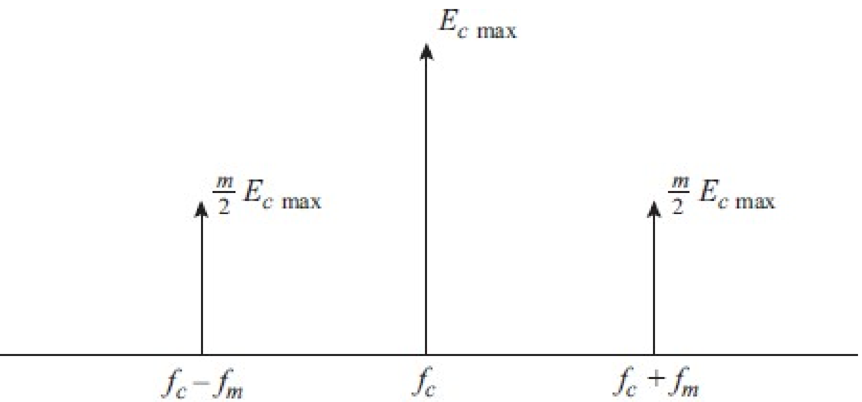
EXAMPLE 8.5.1

A carrier wave of frequency 10 MHz and peak value 10 V is amplitude modulated by a 5-kHz sine wave of amplitude 6 V. Determine the modulation index and draw the amplitude spectrum.

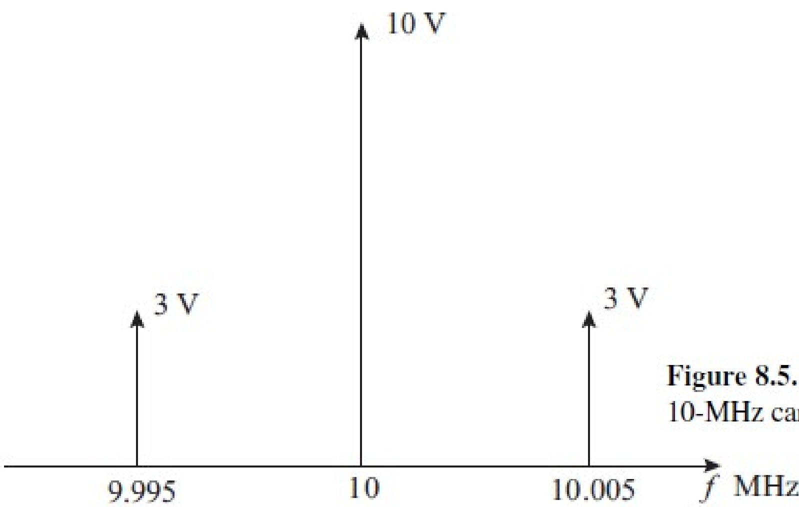
SOLUTION $m = \frac{6}{10} = 0.6$

The side frequencies are $10 \pm 0.005 = 10.005$ and 9.995 MHz. The amplitude of each side frequency is $0.6 \times 10/2 = 3$ V. The spectrum is shown in Fig. 8.5.1(b).

Credit: Electronic Communication by Roddy and Coolen



(a)



(b)

Figure 8.5.1 (a) Amplitude spectrum for a sinusoidally amplitude modulated wave. (b) The amplitude spectrum for a 10-MHz carrier of amplitude 10 V, sinusoidally modulated by a 5-kHz sine wave of amplitude 6 V (Example 8.5.1).

Credit: Electronic Communication by Roddy and Coolen

8.6 Average Power for Sinusoidal AM

Figure 8.5.1(a) shows that the sinusoidally modulated wave can be represented by three sinusoidal sources connected in series. A general result of ac circuit theory is that the average power delivered to a load R by series-connected sinusoidal sources of different frequencies is the sum of the average powers from each source. The average power in a sine (or cosine) voltage wave of peak value E_{max} developed across a resistor R is $P = E^2_{max}/2R$. Applying these results to the spectrum components of the sinusoidally modulated wave gives, for the average carrier power,

$$P_C = \frac{E^2_{c\,max}}{2R} \tag{8.6.1}$$

and for each side frequency

$$\begin{aligned} P_{SF} &= \frac{(mE_{c\,max}/2)^2}{2R} \\ &= \frac{m^2}{4}P_C \end{aligned} \tag{8.6.2}$$

Credit: Electronic Communication by Roddy and Coolen

Hence the total average power is

$$P_T = P_C + 2 \times P_{SF}$$
$$= P_C\left(1 + \frac{m^2}{2}\right) \tag{8.6.3}$$

At 100% modulation (m = 1), the power in any one side frequency component is $P_{SF} = P_C/4$ and the total power is $P_T = 1.5P_C$. The ratio of power in any one side frequency to the total power transmitted is therefore 1/6. The significance of this result is that all the original modulating information is contained in the one side frequency, and therefore a considerable savings in power can be achieved by transmitting just the side frequency rather than the total modulated wave. In practice, the modulating signal generally contains a band of frequencies that results in *sidebands* rather than single side frequencies, but, again, single-sideband (SSB) transmission results in more efficient use of available power and spectrum space. Single-sideband transmission is considered in Chapter 9.

Credit: Electronic Communication by Roddy and Coolen

8.7 Effective Voltage and Current for Sinusoidal AM

The effective or rms voltage E of the modulated wave is defined by the equation

$$\frac{E^2}{R} = P_T \tag{8.7.1}$$

Likewise, the effective or rms voltage E_c of the carrier component is defined by

$$\frac{E_C^2}{R} = P_C \tag{8.7.2}$$

Credit: Electronic Communication by Roddy and Coolen

It follows from Eq. (8.6.3) that

$$\frac{E^2}{R} = P_C\left(1 + \frac{m^2}{2}\right)$$

$$= \frac{E_C^2}{R}\left(1 + \frac{m^2}{2}\right) \tag{8.7.3}$$

from which

$$E = E_C\sqrt{1 + \frac{m^2}{2}} \tag{8.7.4}$$

A similar argument applied to currents yields

$$I = I_C\sqrt{1 + \frac{m^2}{2}} \tag{8.7.5}$$

Credit: Electronic Communication by Roddy and Coolen

where I is the rms current of the modulated wave and I_c the rms current of the unmodulated carrier. The current equation provides one method of monitoring modulation index, by measuring the antenna current with and without modulation applied. From Eq. (8.7.5),

$$m = \sqrt{2\left[\left(\frac{I}{I_c}\right)^2 - 1\right]} \tag{8.7.6}$$

The method is not as sensitive or useful as the trapezoidal method described earlier, but it provides a convenient way of monitoring modulation where an ammeter can be inserted in series with the antenna, for example. A true rms reading ammeter must be used, and care must be taken to avoid current overload because such instruments are easily damaged by overload.

Credit: Electronic Communication by Roddy and Coolen

EXAMPLE 8.7.1

The rms antenna current of an AM radio transmitter is 10 A when unmodulated and 12 A when sinusoidally modulated. Calculate the modulation index.

SOLUTION $m = \sqrt{2\left[\left(\frac{12}{10}\right)^2 - 1\right]} = \mathbf{0.94}$

Credit: Electronic Communication by Roddy and Coolen

8.8 Nonsinusoidal Modulation

Nonsinusoidal modulation has already been illustrated in Fig. 8.2.1 and the modulation index determined as shown in Section 8.3. Sometimes the *modulation depth*, rather than modulation index, is used as a measure of modulation. The modulation depth is the ratio of the downward modulation peak to the peak carrier level, usually expressed as a percentage. As shown in Fig. 8.3.1, overmodulation occurs if the modulation depth exceeds 100%, irrespective of the modulating waveshape. (For sinusoidal modulation, modulation depth is equal to the modulation index. Signal generators generally employ sinusoidal modulation but have meters calibrated in modulation depth.)

Nonsinusoidal modulation produces upper and lower *sidebands,* corresponding to the upper and lower side frequencies produced with sinusoidal modulation. Suppose, for example, that the modulating signal has a line spectrum as shown in Chapter 2 so that it can be represented by

$$e_m(t) = E_{1\max} \cos 2\pi f_1 t + E_{2\max} \cos 2\pi f_2 t + E_{3\max} \cos 2\pi f_3 t + \cdots \tag{8.8.1}$$

Credit: Electronic Communication by Roddy and Coolen

As before, the AM wave is

$$e(t) = [E_{c\,\max} + e_m(t)] \cos 2\pi f_c t \tag{8.8.2}$$

If in general the ith component is denoted by subscript i, then individual modulation indexes may be defined as $m_i = E_{i\,\max}/E_{c\,\max}$ and the trigonometric expansion for Eq. (8.8.2) yields a spectrum with side frequencies at $f_c \pm f_i$ and amplitudes $m_i E_{c\,\max}/2$. This is sketched in Fig. 8.8.1(a). Thus, taken together, the side frequencies form sidebands either side of the carrier component. Again, the practicalities of AM demand that the carrier frequency be much greater than the highest frequency in the modulating wave, so the sidebands are bandlimited about the carrier frequency as shown.

Credit: Electronic Communication by Roddy and Coolen

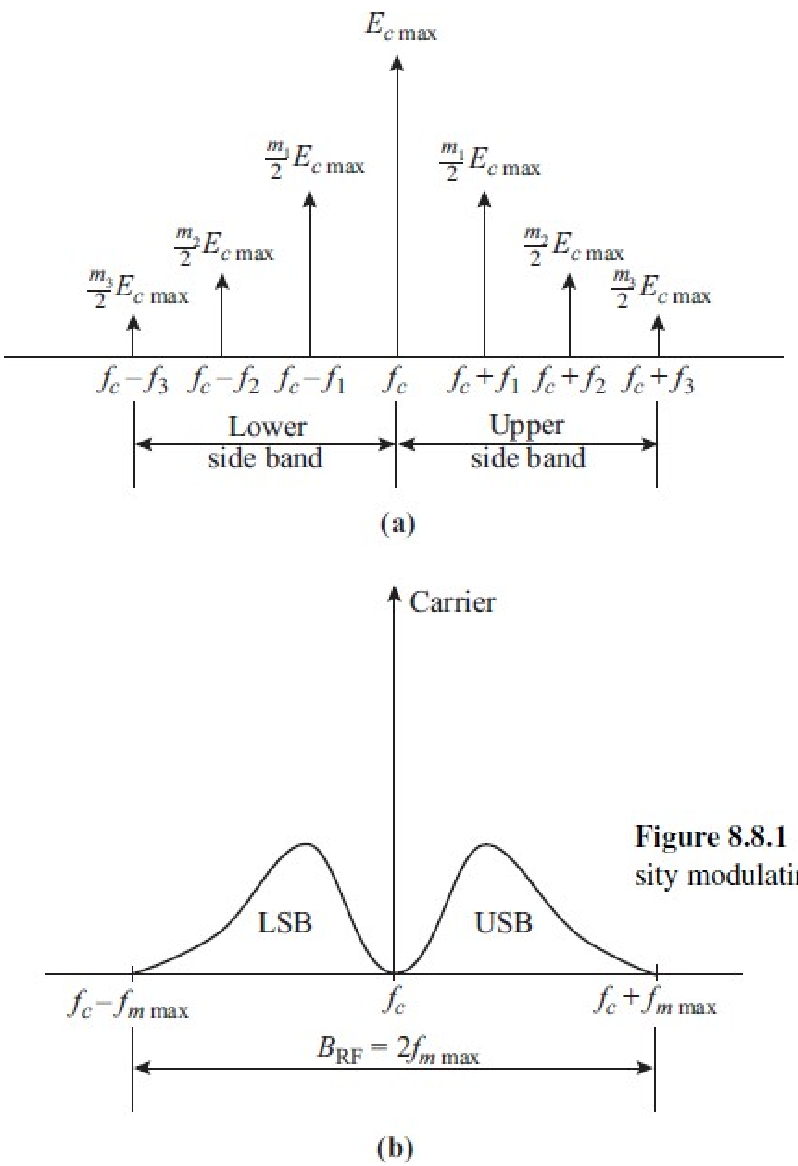


Figure 8.8.1 (a) Amplitude spectrum resulting from line spectra modulation. (b) Amplitude spectrum for a power density modulating spectra.

Credit: Electronic Communication by Roddy and Coolen

The total average power can be obtained by adding the average power for each component (just as was done for single-tone modulation), which results in

$$P_T = P_c\left(1 + \frac{m_1^2}{2} + \frac{m_2^2}{2} + \frac{m_3^2}{2} + \cdots\right) \tag{8.8.3}$$

Hence an effective modulation index can be defined in this case as

$$m_{\text{eff}} = \sqrt{m_1^2 + m_2^2 + m_3^2 + \cdots} \tag{8.8.4}$$

It follows that the effective voltage and current in this case are

$$E = E_c\sqrt{1 + \frac{m_{\text{eff}}^2}{2}} \tag{8.8.5}$$

$$I = I_c\sqrt{1 + \frac{m_{\text{eff}}^2}{2}} \tag{8.8.6}$$

Credit: Electronic Communication by Roddy and Coolen

When the modulating signal is a random power signal such as speech or music, then the concept of power spectral density must be used, as shown in Chapter 2. Thus, if the power spectral density curve is as sketched in Fig. 2.17.1, when used to amplitude modulate the carrier, double sidebands are generated as shown in Fig. 8.8.1(b). Again it is assumed that the modulating signal is bandlimited such that the highest frequency in its spectrum is much less than the carrier frequency.

It will be seen therefore that standard AM produces upper and lower sidebands about the carrier, and hence the RF bandwidth required is double that for the modulating waveform. From Fig. 8.8.1,

$$
\begin{aligned}
B_{\mathrm{RF}} &= (f_c + f_{m\,\max}) - (f_c - f_{m\,\max}) \\
&= 2 f_{m\,\max}
\end{aligned}
\tag{8.8.7}
$$

Credit: Electronic Communication by Roddy and Coolen

8.9 Double-sideband Suppressed Carrier (DSBSC) Modulation

Certain types of amplitude modulators make use of a multiplying action in which the modulating signal multiplies the carrier wave. The balanced mixer described in Section 5.10 is one such circuit, and in fact these are generally classified as *balanced modulators*. As shown by Eq. (5.10.17), the output current contains a product term of the two input voltages. When used as a modulator, the oscillator input becomes the carrier input, and the signal input becomes the modulating signal input. The output voltage can then be written as

$$e(t) = ke_m(t)\cos 2\pi f_c t \tag{8.9.1}$$

where k is a constant of the multiplier circuit. The expression for standard AM with carrier is

$$\begin{aligned} e(t) &= (E_{c\,\max} + e_m(t))\cos 2\pi f_c t \\ &= E_{c\,\max}\cos 2\pi f_c t + e_m(t)\cos 2\pi f_c t \end{aligned} \tag{8.9.2}$$

Credit: Electronic Communication by Roddy and Coolen

Double-sideband Suppressed Carrier (DSBSC) Modulation

The major difference between the multiplier result and this is that carrier term $E_{c\,\max}\cos 2\pi f_c t$ is absent from the multiplier result. This means that the carrier component will be absent from the spectra, which otherwise will be the same as for AM with carrier. The constant multiplier k can be regarded simply as a scaling factor, and it will not materially affect the results. This type of amplitude modulation is therefore known as double-sideband suppressed carrier (DSBSC). The spectra are sketched in Fig. 8.9.1 for sinusoidal modulation and for the general case.

The absence of a carrier component means that DSBSC utilizes the transmitted power more efficiently than standard AM; however, it still requires twice the bandwidth compared to single sideband (SSB). It should be noted that, although the bandwidth is double that required for SSB, the received power is also double that obtained with SSB, and therefore the signal-to-noise ratio is the same. However, conserving bandwidth is an important aim in communications systems, and usually DSBSC represents one step in generating SSB, as described in Chapter 9.

Credit: Electronic Communication by Roddy and Coolen

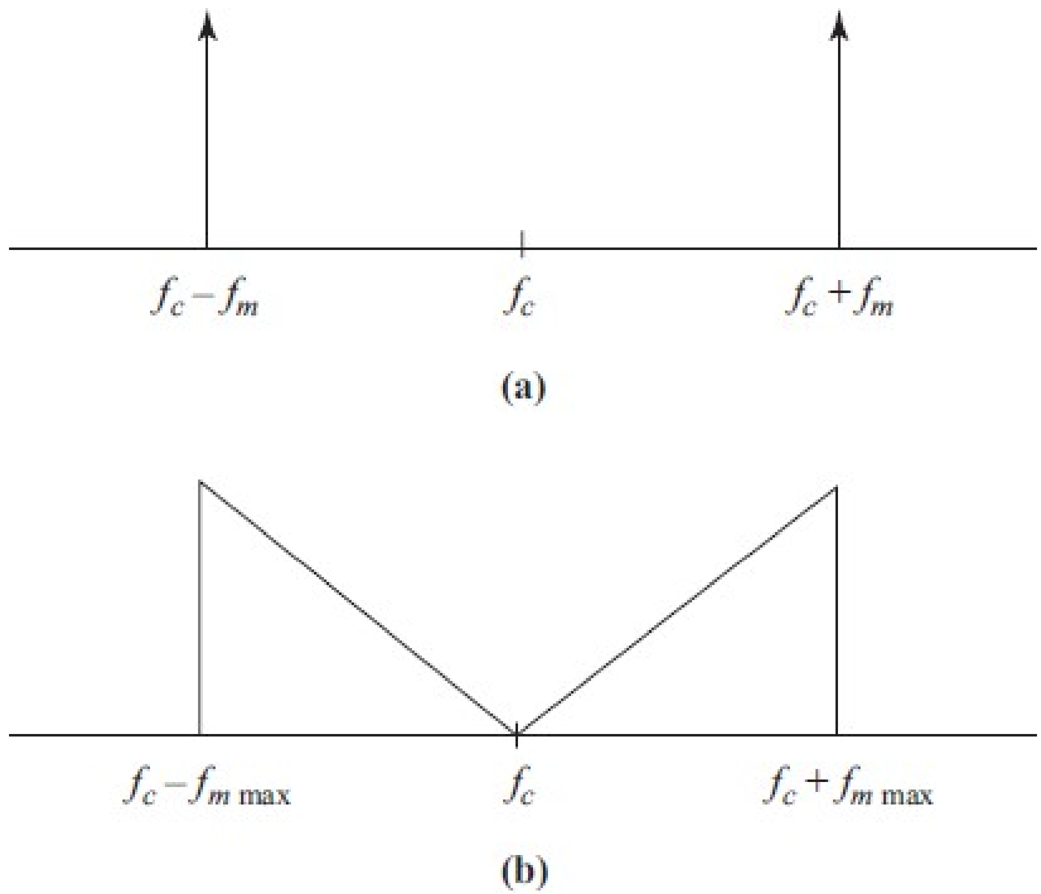
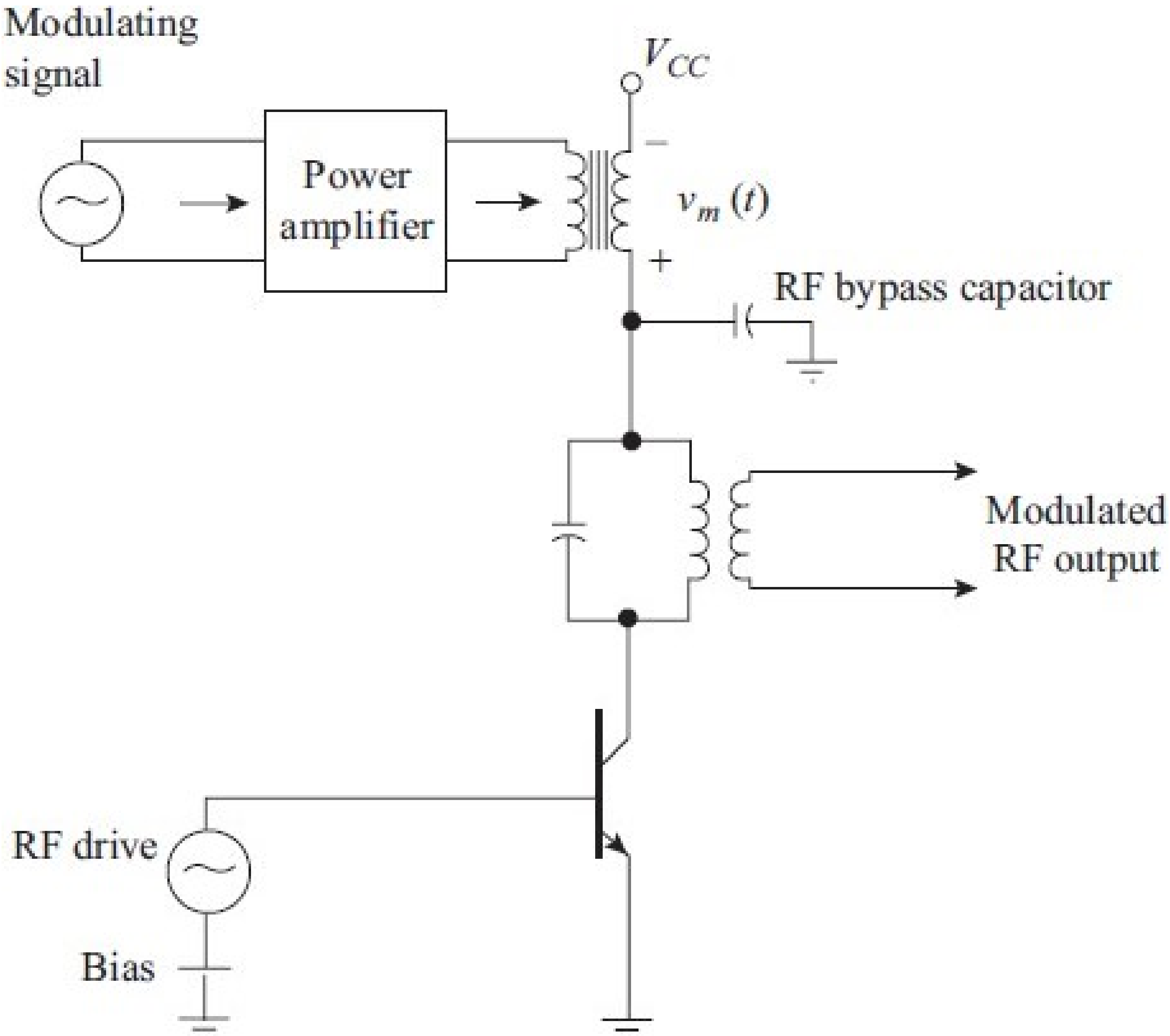


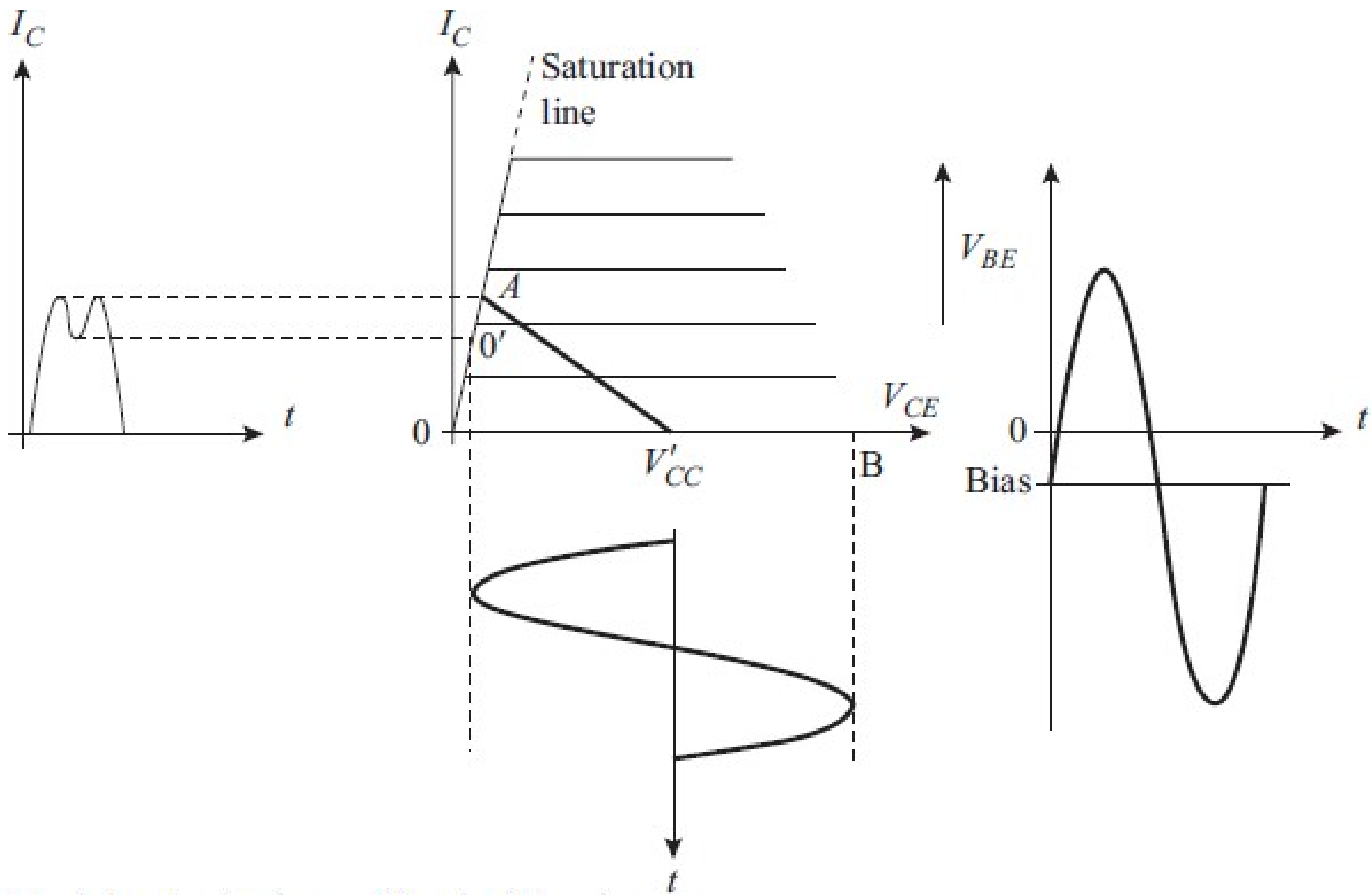
Figure 8.9.1 DSBSC spectrum for (a) sinusoidal modulation and (b) the general case.

Credit: Electronic Communication by Roddy and Coolen



Credit: Electronic Communication by Roddy and Coolen

Basic circuit for a BJT collector modulator.



The BJT output characteristics, also showing one RF cycle of operation.

Credit: Electronic Communication by Roddy and Coolen

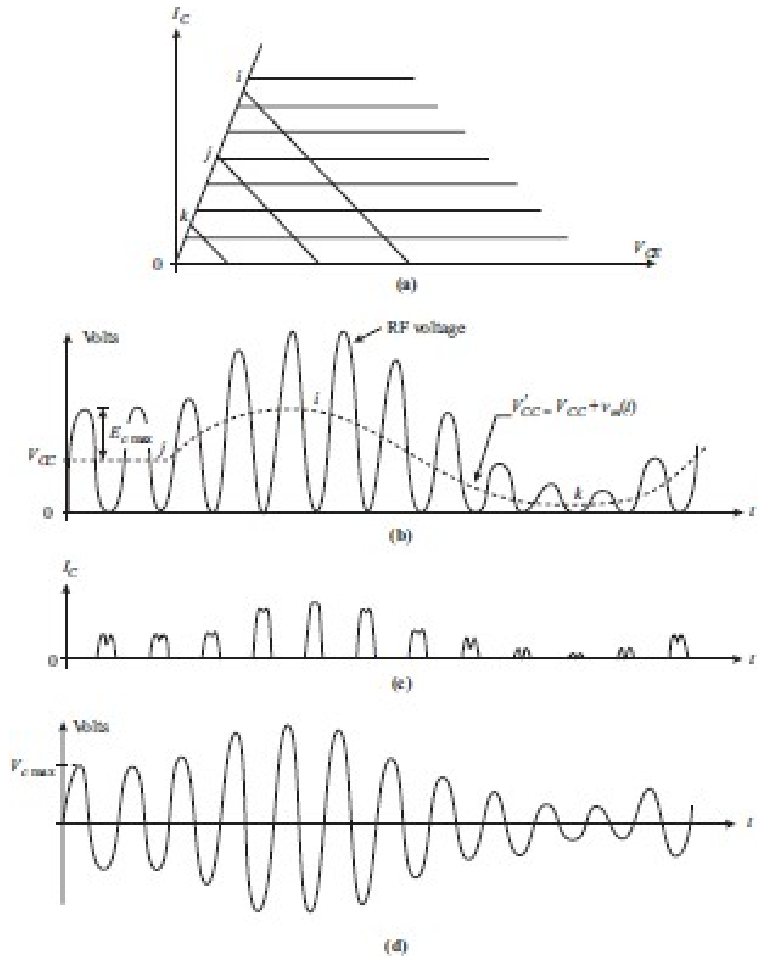


Figure 8.10.3 (a) Output characteristics showing three load lines. (b) Collector voltage. (c) Current pulses. (d) Modulated output voltage.

Credit: Electronic Communication by Roddy and Coolen

$$m = \frac{E_{\max} - E_{\min}}{E_{\max} + E_{\min}}$$

$$= \frac{E_{\max\,pp} - E_{\min\,pp}}{E_{\max\,pp} + E_{\min\,pp}}$$

$$= \frac{V_{\max\,pp} - V_{\min\,pp}}{V_{\max\,pp} + V_{\min\,pp}}$$

It is left as an exercise for the student to show that for sinusoidal modulation in particular

$$m = \frac{V_{m\,\max}}{V_{CC}} \tag{8.10.2}$$

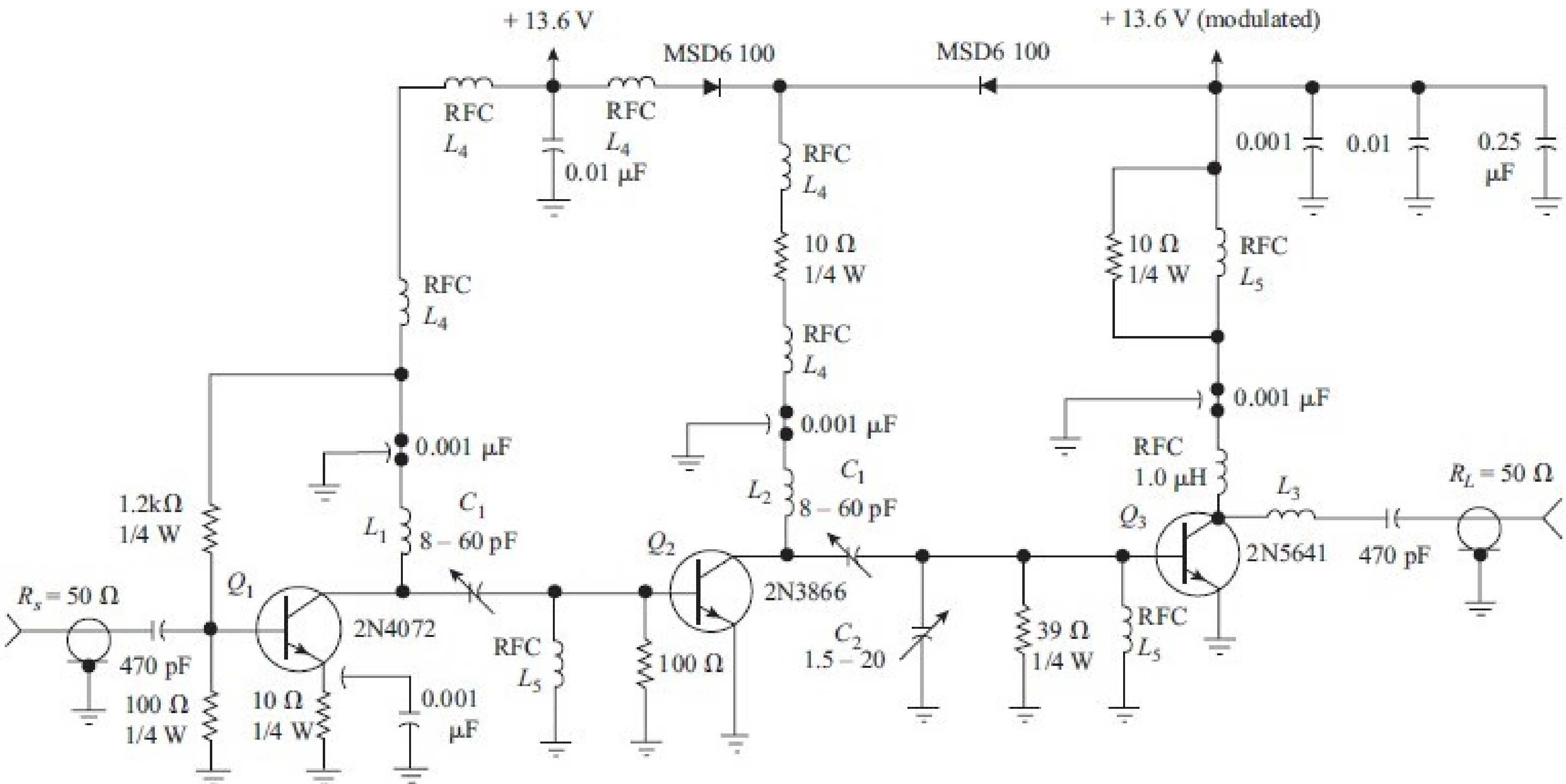
where $V_{m\,\max}$ is the peak value of $\upsilon_{\mathrm{m}}(t)$.

Credit: Electronic Communication by Roddy and Coolen

The class C amplifier can be considered as a power converter. When no modulation is applied, it converts the dc input power to the collector to the unmodulated RF output power, or $P_C = \eta P_{CC}$, where P_C is the unmodulated carrier power, P_{CC} the dc input power to the collector, and η is the conversion efficiency (conversion efficiencies can be quite high, typically 70% to 75%). When modulation is applied, the "dc input" becomes a slowly varying input and assuming that the conversion efficiency remains the same, the additional power supplied by the modulator goes into the creation of the sidebands or $P_{SB} = \eta P_{\text{mod}}$, where P_{SB} is the total average sideband power and P_{mod} the average power supplied by the modulator. Looking at this in another way, the modulator has to supply power equal to

$$P_{\text{mod}} = \frac{P_{\text{SF}}}{\eta} \tag{8.10.3}$$

Credit: Electronic Communication by Roddy and Coolen



L_1 = 6T #26 wire wound on toroid (micro-metals T30-13) with $\frac{3}{32}''$ spacing

L_2 = 2T #26 wire wound on toroid (see L_1) with $\frac{1''}{8}$ spacing

L_3 = 2T #26 wire wound on toroid (see L_1) with $\frac{5''}{16}$ spacing

L_4 = RF bead (one hole), $\frac{1''}{8}$

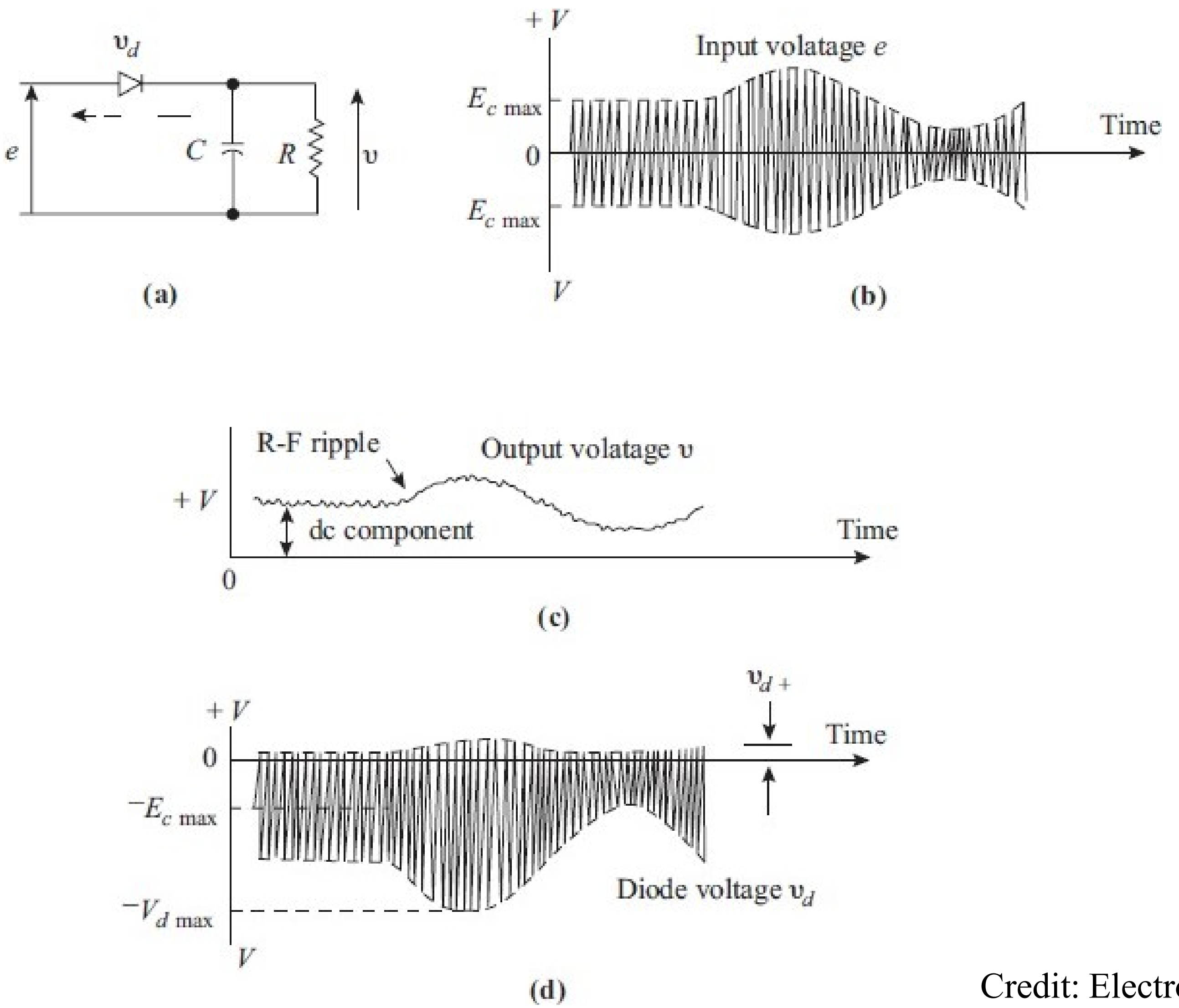
L_5 = Ferrite Choke (Ferroxcube VK 200)

C_1 = 8 – 60 pF (Arco 404)

C_2 = 1.5-20 pF (Arco 402)

Transistorized collector modulator circuit. (Courtesy of Motorola Inc, Applicable Note AN507.)

Credit: Electronic Communication by Roddy and Coolen



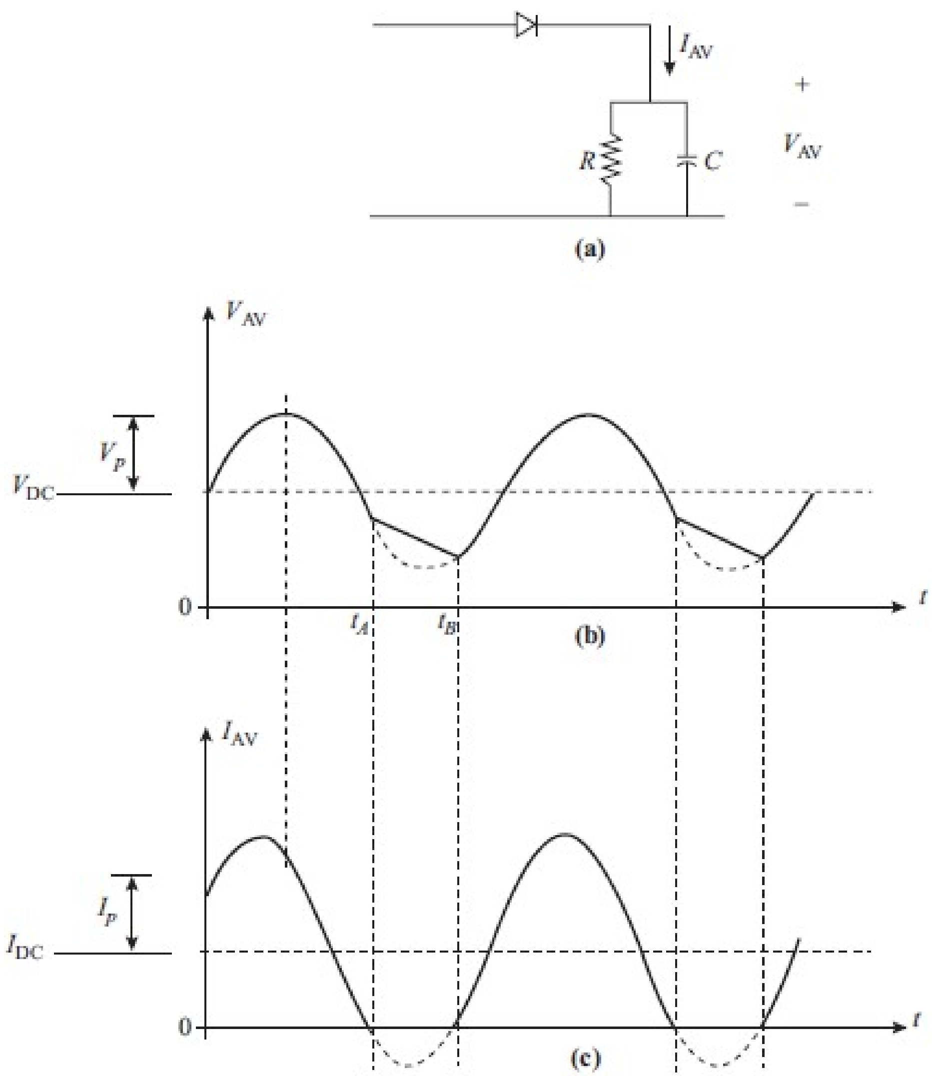
Credit: Electronic Communication by Roddy and Coolen

Figure 8.11.1 (a) Basic diode envelope detector. (b) Voltage input waveform. (c) Voltage output waveform. (d) Voltage across the diode.

Diagonal Peak Clipping. This is a form of distortion that occurs when the time constant of the *RC* load is too long, thus preventing the output voltage from following the modulation envelope. The output voltage is labeled V_{AV} in Fig. 8.11.2(a), to show that it is the average voltage that follows the modulation envelope (that is, the RF ripple is averaged out). The curve of V_{AV} for sinusoidal modulation is shown in Fig. 8.11.2(b). At some time t_A the modulation envelope starts to decrease more rapidly than the capacitor discharges. The output voltage then follows the discharge curve of the *RC* network until time t_B, when it meets up with the modulation envelope as it once again increases.

For sinusoidal modulation the condition necessary for the avoidance of diagonal peak clipping is found as follows. Because of the capacitive nature of the *RC* load, the current leads the voltage as shown in

Credit: Electronic Communication by Roddy and Coolen



Credit: Electronic Communication by Roddy and Coolen

Figure 8.11.2 (a) Diode circuit supplying an average current I_{AV} to the RC load. (b) Voltage waveform, illustrating diagonal peak clipping. (c) Current waveform.

Fig. 8.11.2(c). The average current consists of two components, a dc component I_{DC} and an ac component that has a peak value I_P, as shown in Fig. 8.11.2(c). The dc component of voltage is approximately equal to the maximum unmodulated carrier voltage or $V_{DC} \cong E_{c\ \max}$ and the direct current is $I_{DC} = V_{DC}/R$. The peak value of the average output voltage is $Vp \cong m_{ec\ \max}$, and the corresponding value of the peak current is $I_P = V_P/Z_P$, where Z_p is the impedance of the RC load at the modulating frequency.

If the envelope falls faster than the capacitor discharges, the diode ceases to conduct (since the capacitor voltage biases it off), and the current I_{AV} supplied by the diode goes to zero. This is shown in Fig. 8.11.2(c). During the period the current is zero, the load voltage follows the discharge law of the RC network, resulting in the diagonally clipped peak shown in Fig. 8.11.2(b). From Fig. 8.11.2(c), it is seen that for the avoidance of diagonal peak clipping the direct current has to be greater than the peak current, or $I_{DC} \geq |I_P|$. Hence

$$\frac{V_{DC}}{R} \geq \frac{mV_{DC}}{|Z_p|}$$

$$\therefore m \leq \frac{|Z_p|}{R} \qquad (8.11.2)$$

Credit: Electronic Communication by Roddy and Coolen

EXAMPLE 8.11.1

The *RC* load for a diode detector consists of a 1000-pF capacitor in parallel with a 10-kΩ resistor. Calculate the maximum modulation depth that can be handled for sinusoidal modulation at a frequency of 10 kHz if diagonal peak clipping is to be avoided.

SOLUTION The admittance of the *RC* load is

$$Y_p = \frac{1}{R} + j2\pi f_m C$$

$$= 10^{-4} + j6.2810^{-5}\ \mathrm{S}$$

Hence

$$|Z_p| = \frac{1}{|Y_p|} = 8467\ \Omega$$

The maximum modulation index that can be handled without distortion is therefore

$$m = \frac{|Z_p|}{R} \cong \mathbf{0.85}$$

Credit: Electronic Communication by Roddy and Coolen

Negative Peak Clipping. This is similar in appearance to diagonal peak clipping, but results from the loading effect of the network R_1C_1 following the RC load [Fig. 8.11.3(a)]. Capacitor C_1 is a dc blocking capacitor, and resistor R_1 represents the input resistance of the following stage.

Considering the normal situation where the reactance of C_1 is very small, and that of C is very large at the modulating frequency (assumed sinusoidal), the ac impedance is simply R in parallel with R_1 or $|Z_p| = R_p = RR_1/(R = R_1)$. The modulation index must now meet the condition

$$m \le \frac{R_p}{R} \tag{8.11.3}$$

In this situation, the current I_{AV} is in phase with V_{AV}, and over the period when I_{AV} is zero [Fig. 8.11.3(b)], the C_1 capacitor voltage remains approximately constant at V_{DC}, which in turn develops a voltage equal to $V_{\text{MIN}} = V_{DC}\,R\,/\,(R + R_1)$ across R. It is this voltage that keeps the diode biased off. The voltages across R and R_1 are shown in Fig. 8.11.3(c). The shape of the output voltage curve shows why the term *negative peak clipping* is used.

Credit: Electronic Communication by Roddy and Coolen

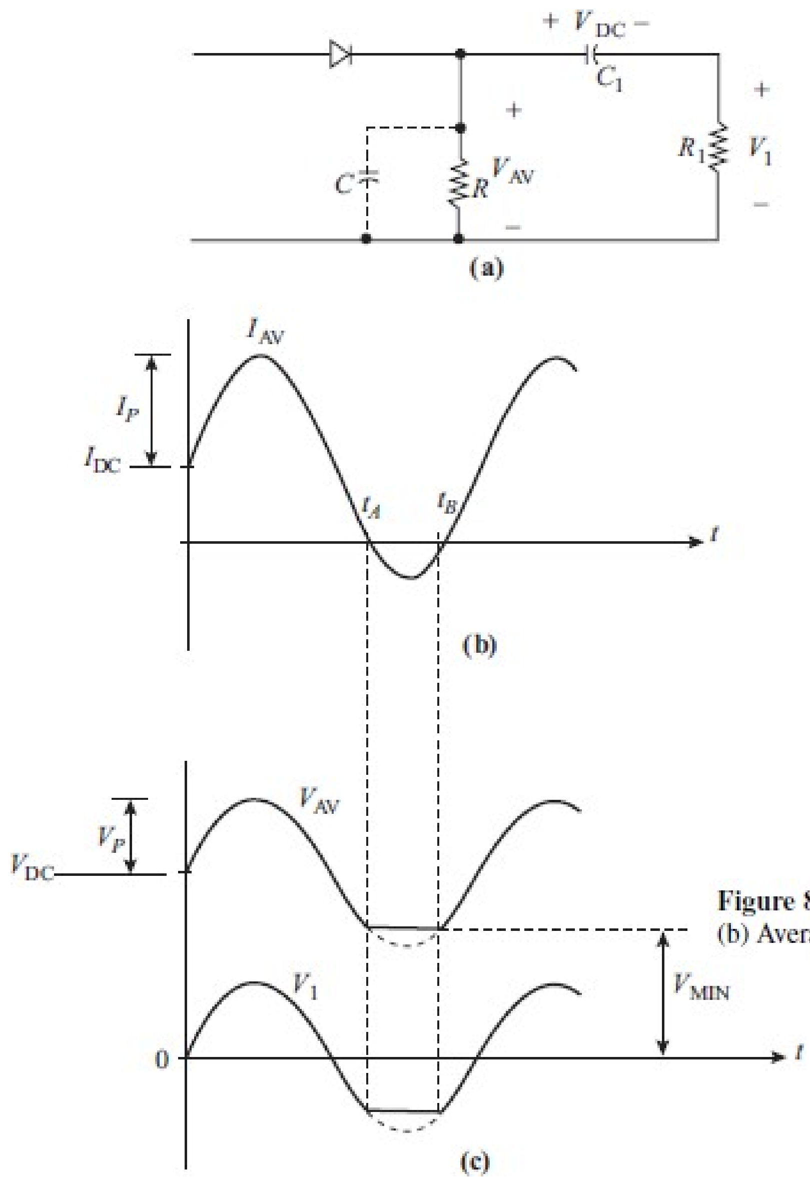


Figure 8.11.3 (a) Diode circuit including a dc blocking capacitor C_1 and the input resistor R_1 of the following stage. (b) Average diode current. (c) The voltages across R and R_1.

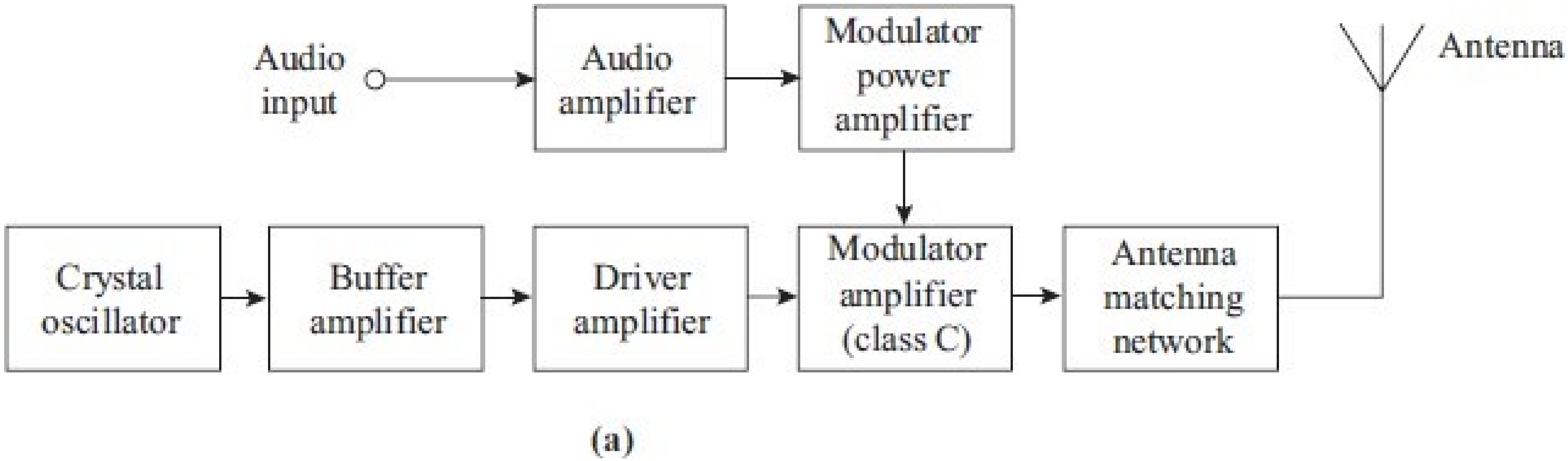
Credit: Electronic Communication by Roddy and Coolen

Figure 8.12.1(a) shows the block diagram of a typical AM transmitter. The carrier source is a crystal-controlled oscillator at the carrier frequency or a submultiple of it. This is followed by a tuned buffer amplifier and a tuned driver, and if necessary frequency multiplication is provided in one or more of these stages.

The modulator circuit used is generally a class C power amplifier that is collector modulated as described in Section 8.10. The audio signal is amplified by a chain of low-level audio amplifiers and a power amplifier. Since this amplifier is controlling the power being delivered to the final RF amplifier, it must have a power driving capability that is one-half the maximum power the collector supply must deliver to the RF amplifier under 100% modulation conditions. A transformer-coupled class B push–pull amplifier is usually used for this purpose.

Low-power transmitters with output powers up to 1 kW or so may be transistorized, but as a rule the higher-power transmitters use vacuum tubes in the final amplifier stage, even though the low-level stages may be transistorized. In some cases where the reliability and high overall efficiency of the transistor are mandatory, higher powers can be obtained by using several lower-power transistorized amplifiers in parallel. The system is complicated, and usually the vacuum-tube version will do the same job at lower capital cost.

Credit: Electronic Communication by Roddy and Coolen



(a) transmitter with a modulated class C final power amplifier

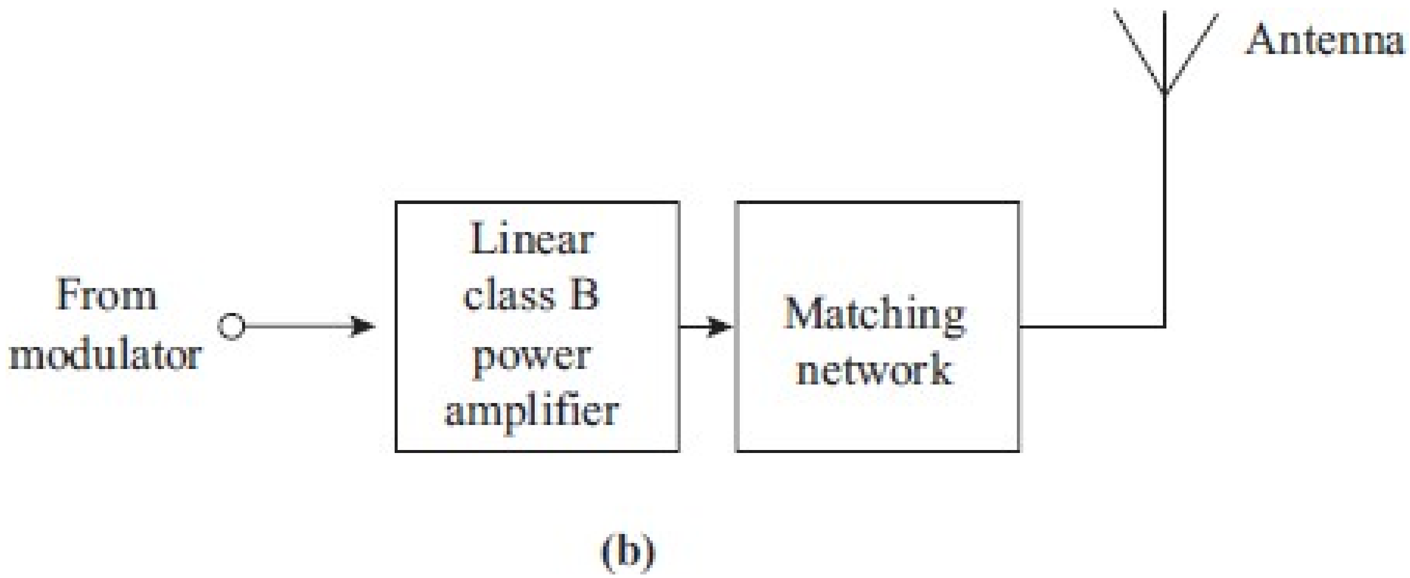
Credit: Electronic Communication by Roddy and Coolen

Sometimes the modulation function is done in one of the low-level stages. This allows low-power modulation and audio amplifiers, but it complicates the RF final amplifier. Class C amplifiers cannot be used to amplify an already modulated (AM) carrier, because the transfer function of the class C amplifier is not linear. The result of using a class C amplifier would be an unacceptable distortion of the modulation envelope. A linear power amplifier, such as the push–pull class B amplifier, must be used to overcome this problem [Fig. 8.12.1(b)].

Unfortunately, the efficiency of this type of amplifier is lower than that of the comparable class C amplifier, resulting in more costly equipment. Larger tubes or transistors must be used that are capable of dissipating the additional heat generated.

The output of the final amplifier is passed through an impedance-matching network that includes the tank circuit of the final amplifier. The Q of this circuit must be low enough so that all the sidebands of the signal are passed without amplitude/frequency distortion, but at the same time must present an appreciable attenuation at the second harmonic of the carrier frequency. The bandwidth required in most cases is a standard 3 dB at ± 5 kHz around the carrier. For amplitude-modulation broadcast transmitters, this response may be broadened so that the sidebands will be down less than 1 dB at 5 kHz where music programs are being broadcast and very low distortion levels are desired, or special sharp-cutoff filters may be used. Because of the high power levels present in the output, this is not usually an attractive solution.

Credit: Electronic Communication by Roddy and Coolen

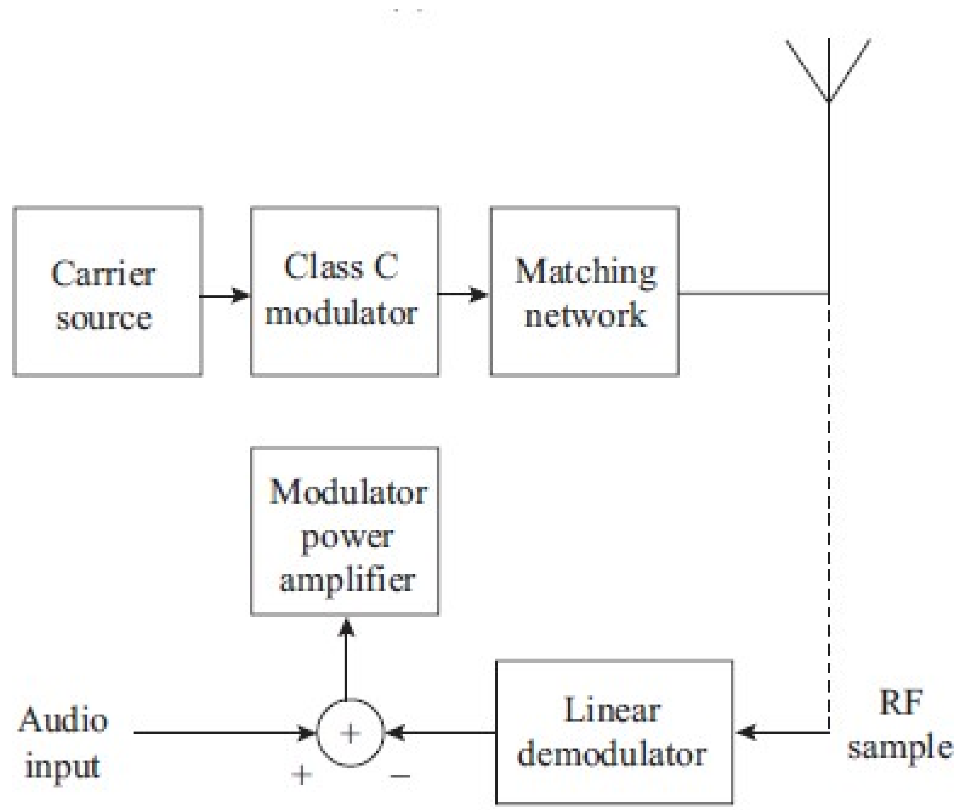


(b) Linear class B push–pull power amplifier used when modulation takes place in a low-level stage

Credit: Electronic Communication by Roddy and Coolen

Negative feedback is quite often used to reduce distortion in a class C modulator system. The feedback is accomplished in the manner shown in Fig. 8.12.1(c), where a sample of the RF signal sent to the antenna is extracted and demodulated to produce the feedback signal. The demodulator is designed to be as linear in its response as possible and to feed back an audio signal that is proportional to the modulation envelope. The negative feedback loop functions to reduce the distortion in the modulation.

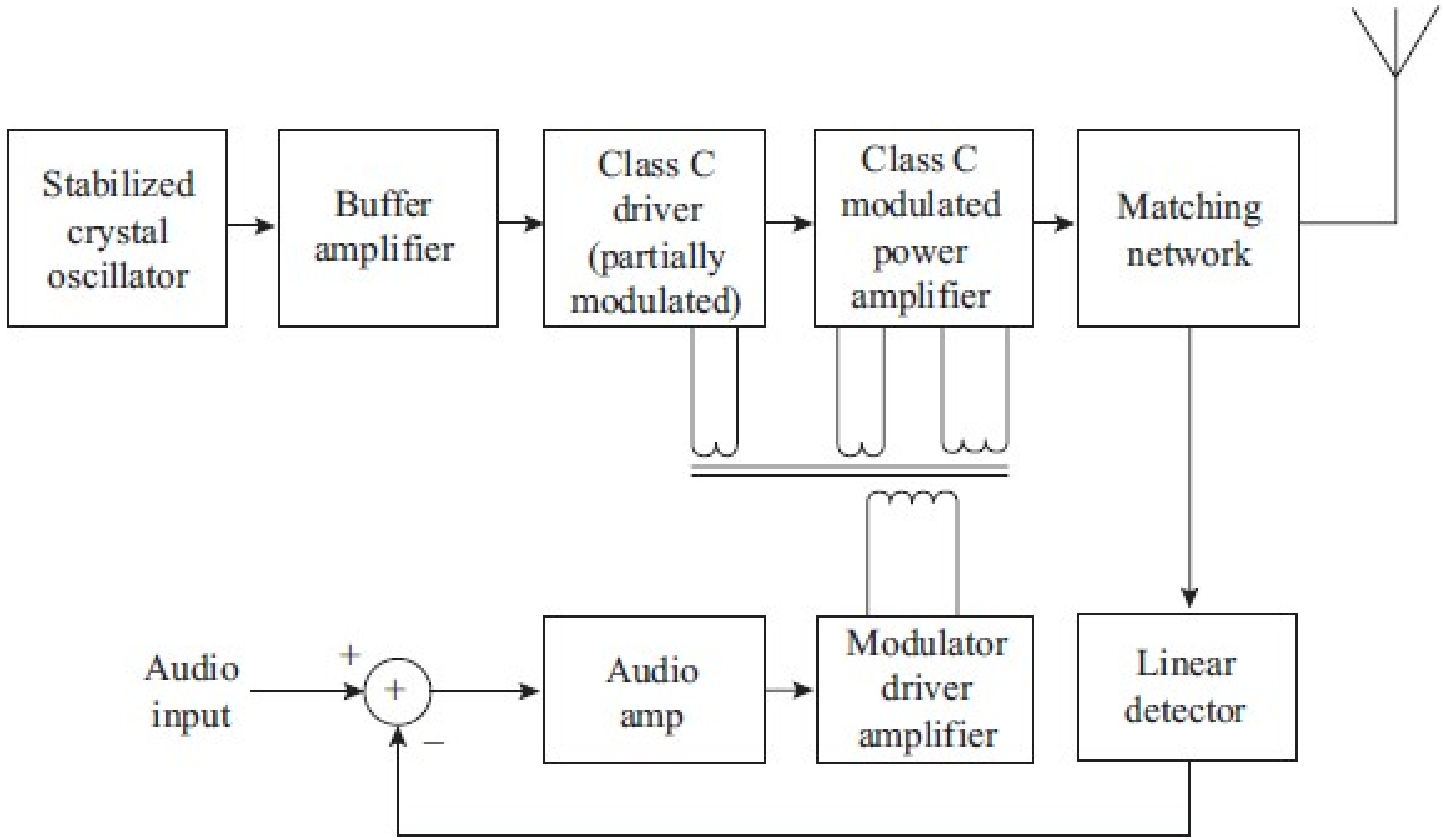
Credit: Electronic Communication by Roddy and Coolen



(c)

(c) negative feedback applied to linearize a class C modulator.

Credit: Electronic Communication by Roddy and Coolen



AM broadcast transmitter

Credit: Electronic Communication by Roddy and Coolen

Most domestic AM broadcast services use the medium-wave band from 550 to 1600 kHz. International AM broadcasts take place in several of the HF bands scattered from 1600 kHz up to about 15 MHz. The mode of transmission in all cases is double-sideband full carrier, with an audio baseband range of 5 kHz. Station frequency assignments are spaced at 10 kHz intervals, and power outputs range from a few hundred watts for small local stations to as much as 100 kW in the MW band and even higher for international HF transmitters.

A main requirement of an AM broadcast transmitter is to produce, within the limits of the 5-kHz audio bandwidth available, the highest possible fidelity. The modulator circuits in the transmitter must produce a linear modulation function, and every trick available is used to accomplish this. A typical AM broadcast transmitter is shown in Fig. 8.12.2. The crystal oscillator is temperature-controlled to provide frequency stability. It is followed by a buffer amplifier and then by tuned class C amplifiers that provide the necessary power gain to drive the final power amplifier. For high power output, vacuum tubes would be used as described next. The modulator system is the *triple equilibrium* system, in which the main part of the modulation is performed by plate-modulating the final class C power amplifier. Secondary modulation of both the final grid and the plate of the driver stage is also included to compensate for bias shift in the final amplifier that results from the nonlinear characteristic of the amplifier.

Credit: Electronic Communication by Roddy and Coolen

The final power amplifier is a push–pull parallel stage in which each side of the push–pull stage is composed of several vacuum tubes operating in parallel, to obtain the power required. A further advantage of this system is that, if one or more of the tubes in the system should fail, the remaining tubes will provide partial output until repairs can be made, thus making a more secure system. Power dissipation in these final tubes can be as high as 50 kW, in addition to several kilowatts of heater power. Water cooling systems are used to dissipate the large quantities of heat produced.

The modulator amplifier is an audio-frequency push–pull parallel amplifier, which is transformer-coupled to the modulator. The audio preamplifier stage includes a difference amplifier and an envelope detector that demodulates a sample of the transmitter output and uses the signal to provide negative feedback. This feedback further linearizes the modulation characteristic of the system.

Credit: Electronic Communication by Roddy and Coolen

Single-sideband Principles

In Section 8.9 it is shown that the output from a balanced modulator contains the term

$$e(t) = ke_m(t)\cos\omega_c t \tag{9.2.1}$$

where k is the multiplier constant. With cosinusoidal input $e_m(t) = E_{m\ \max}\cos\omega_c t$, the modulator output becomes a DSBSC signal containing two side frequencies,

$$\begin{aligned} e(t) &= kE_{m\ \max}\cos\omega_m t\cos\omega_c t \\ &= E_{\max}[\cos(\omega_c - \omega_m)t + \cos(\omega_c + \omega_m)t] \end{aligned} \tag{9.2.2}$$

where $E_{\max} = k(E_{m\ \max}/2)$.

Credit: Electronic Communication by Roddy and Coolen

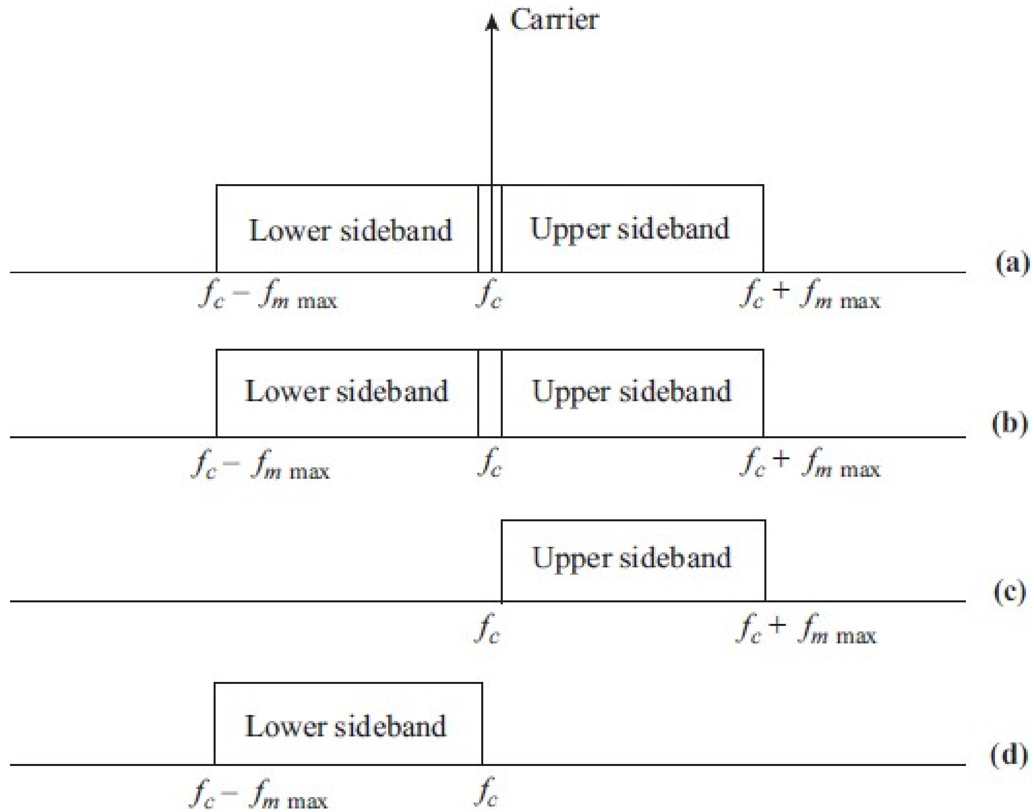


Figure 9.1.1 Amplitude-modulated signal spectra: (a) normal amplitude modulation, or double-sideband full carrier; (b) double-sideband suppressed carrier (DSBSC); (c) single-sideband suppressed carrier (SSBSC) using the upper sideband (USB); (d) single-sideband suppressed carrier (SSBSC) using the lower sideband (LSB).

Credit: Electronic Communication by Roddy and Coolen

Now if one of the side frequencies in the DSBSC signal is removed, either by filtering or by cancellation, the other side frequency will remain. For cosinusoidal modulation the *upper side frequency* (USF) signal is described by

$$e_{\mathrm{USF}} = E_{\max} \cos(\omega_c + \omega_m)t \tag{9.2.3}$$

and the *lower side frequency* (LSF) signal is described by

$$e_{\mathrm{LSF}} = E_{\max} \cos(\omega_c - \omega_m)t \tag{9.2.4}$$

Since all the transmitted power goes into the side frequency, then

$$P_T = \frac{E_{\max}^2}{2R} \tag{9.2.5}$$

Credit: Electronic Communication by Roddy and Coolen

Where the modulating signal contains a band of frequencies (usually the case in practice), the terms *upper sideband* (USB) and *lower sideband* (LSB) are used.

Demodulation of a single-sideband signal is achieved by multiplying it with a locally generated *synchronous* carrier signal at the receiver. Detectors using this principle are called *product detectors*, and balanced modulator circuits are used for this purpose. It is important that the carrier be as closely synchronized in frequency and phase with the original carrier as possible to avoid distortion of the modulated output.

To demonstrate that the multiplying process does demodulate an SSB signal, consider an LSF signal $E_{\max}\cos(\omega_c - \omega_m)t$ multiplied by a local oscillator signal $E_{c\,\max}\cos\omega_c t$ using a balanced modulator with a gain k. Equation (5.10.11) shows that the mixer operating with a large oscillator input signal contains a term

$$\begin{aligned} e_{\text{out}} &= kE_{\max}\cos(\omega_c - \omega_m)t\cos\omega_c t \\ &= \frac{kE_{\max}}{2}[\cos\omega_m t + \cos(2\omega_c - \omega_m)t] \end{aligned} \tag{9.2.6}$$

The first term on the right of the equation is the required information signal, while the second term is the lower side frequency at the second harmonic of the local carrier frequency. Low-pass filtering easily removes this, leaving only the demodulated information (or baseband) signal as

$$e_{bb}(t) = \frac{kE_{\max}}{2}\cos\omega_m t \tag{9.2.7}$$

Credit: Electronic Communication by Roddy and Coolen

Balanced Modulators

Balanced modulators are the building blocks from which a wide variety of frequency mixers, modulators, and demodulators are built. Any circuit that multiplies two input signals while canceling the feedthrough of one of these is a singly balanced modulator, and one that cancels both is a doubly balanced modulator. The output contains a double-sideband suppressed carrier signal.

Credit: Electronic Communication by Roddy and Coolen

An FET Singly Balanced Modulator Circuit

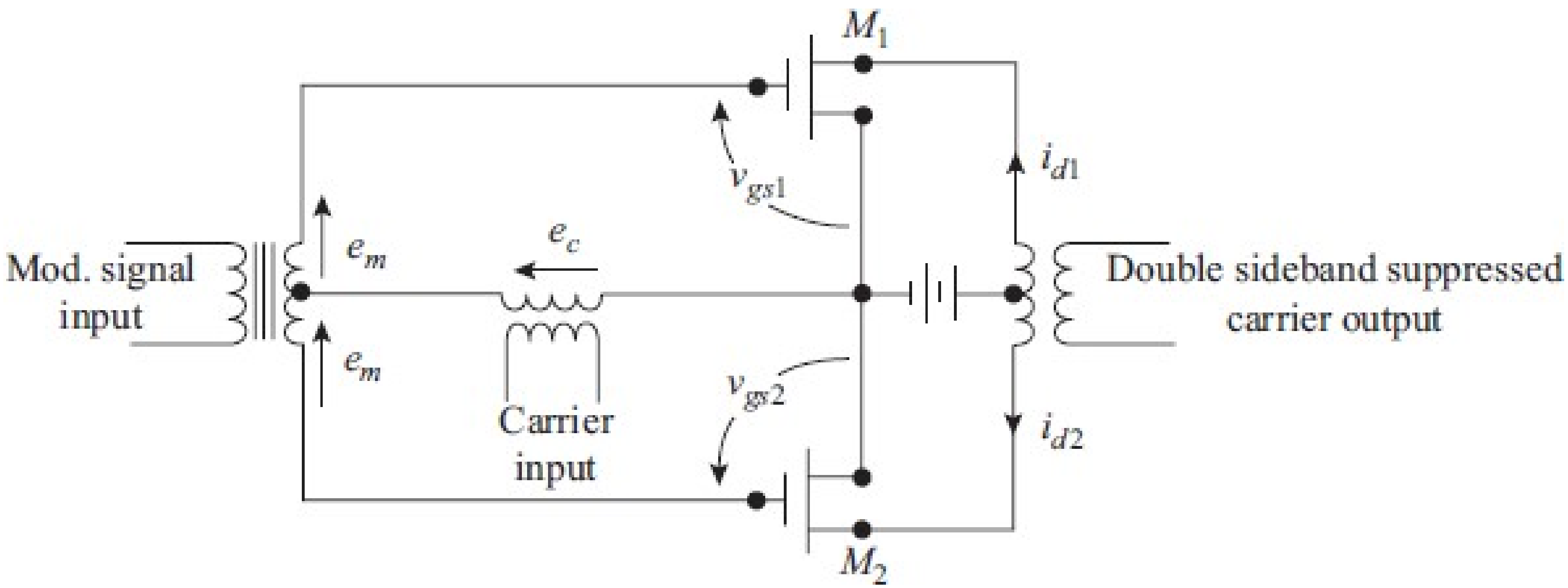


Figure 9.3.1 FET singly balanced modulator circuit.

Credit: Electronic Communication by Roddy and Coolen

An FET Singly Balanced Modulator Circuit

Figure 9.3.1 shows two matched FETs connected in a differential amplifier, which acts as a singly balanced modulator in which the carrier oscillator signal is canceled from the output, but the modulating signal appears in the output. The input (modulating) signal is applied in the differential input mode, and the carrier signal is applied as a common-mode signal. The signal applied to the gate of M_1 is the sum of the two input voltages ($e_c + e_m$), while the signal applied to M_2 is the difference ($e_c - \mathrm{e}_m$). These two components are squared by the second-order terms of the transistor transfer functions. The common-mode carrier signal remaining is canceled as the two drain currents are subtracted in the output transformer primary.

Credit: Electronic Communication by Roddy and Coolen

An FET Singly Balanced Modulator Circuit

$$V_{gs1} = e_c + e_m \quad (9.3.1)$$

$$V_{gs2} = e_c - e_m \quad (9.3.2)$$

$$i_{d1} = I_o + aV_{gs1} + bV_{gs1}^2 \quad (9.3.3)$$

$$i_{d2} = I_o + aV_{gs2} + bV_{gs2}^2 \quad (9.3.4)$$

$$i_p = i_{d1} - i_{d2} = a(V_{gs1} - V_{gs2}) + b(V_{gs1} + V_{gs2})(V_{gs1} - V_{gs2}) \quad (9.3.5)$$

Credit: Electronic Communication by Roddy and Coolen

Substituting Eqs. (9.3.1) and (9.3.2) into (9.3.5) gives

$$i_p = 2a(e_m) + 4b(e_m)(e_c) \tag{9.3.6}$$

Substituting cosinusoidal signals in Eq. (9.3.6) yields the output

$$i_p = 2aE_{m\,\max} \cos \omega_m t + 2bE_{c\,\max}E_{m\,\max}[\cos(\omega_c - \omega_m)t + \cos(\omega_c + \omega_m)t] \tag{9.3.7}$$

This output contains the original modulating signal and the two sidebands about the carrier frequency position. The carrier is absent. It should be noted that any imbalance in the circuit so that either the a's or b's for the two FETs differ from each other will allow some of the carrier signal to feed through to the output. In practice, the FETs would be a very closely matched pair on a single chip, and the bias currents to the two FETs would be adjusted for minimum carrier feedthrough. Since the output would be fed through a band-pass filter, the low-frequency modulating signal component would be removed at that point.

Credit: Electronic Communication by Roddy and Coolen

Integrated-circuit Doubly Balanced Modulators

The disadvantages of the FET circuit described are that the modulating input signal feeds through to the output, the circuit is difficult to balance, and the input and output require specially balanced transformers. Integrated-circuit doubly balanced modulators like the LM1596 described in Section 5.10 operate as multiplier circuits that produce only sideband pairs at the output. Application is simple, requiring only bias and an appropriate band-pass filter to eliminate sideband pairs at harmonics of the carrier. Very little adjustment is required to obtain good balance.

An important advantage of the integrated-circuit balanced modulator is that, when it is operated with a large carrier signal, the output signal amplitude is independent of the carrier amplitude. The result is that the output amplitude depends only on the amplitude of the input signal (which is the modulating signal when it is used as a modulator or the sideband signal when it is used as a demodulator).

Credit: Electronic Communication by Roddy and Coolen

Doubly Balanced Diode Ring Modulator

A circuit known as the *double-balanced ring modulator,* which is widely used in carrier telephony, is shown in Fig. 9.3.2(a). The name comes from the fact that the circuit is balanced to reject both the carrier and modulating signals using a ring of diodes. The output contains only sideband pairs about the carrier frequency position and several of its harmonics.

Operation of the circuit is similar to that described for the integrated-circuit balanced modulator in Section 5.10. A large signal carrier acts as a switching signal to alternate the polarity of the modulating signal at the carrier frequency. With a negative carrier voltage V_c applied, diodes *AB* and *CD* conduct and diodes *AD* and *BC* block to give the effective connection shown in Fig. 9.3.2(b). With a positive carrier voltage, diodes *AD* and *BC* conduct and diodes *AB* and *CD* block to give the connection of Fig. 9.3.2(c). The effect is to multiply the modulating signal by a fixed-amplitude square wave at the carrier frequency, producing the required DSBSC signal, with harmonics. Band-pass filters remove the unwanted harmonics from the output. Fig. 9.3.3 shows the time response waveform for a single sinusoid of modulation and its spectrum.

Credit: Electronic Communication by Roddy and Coolen

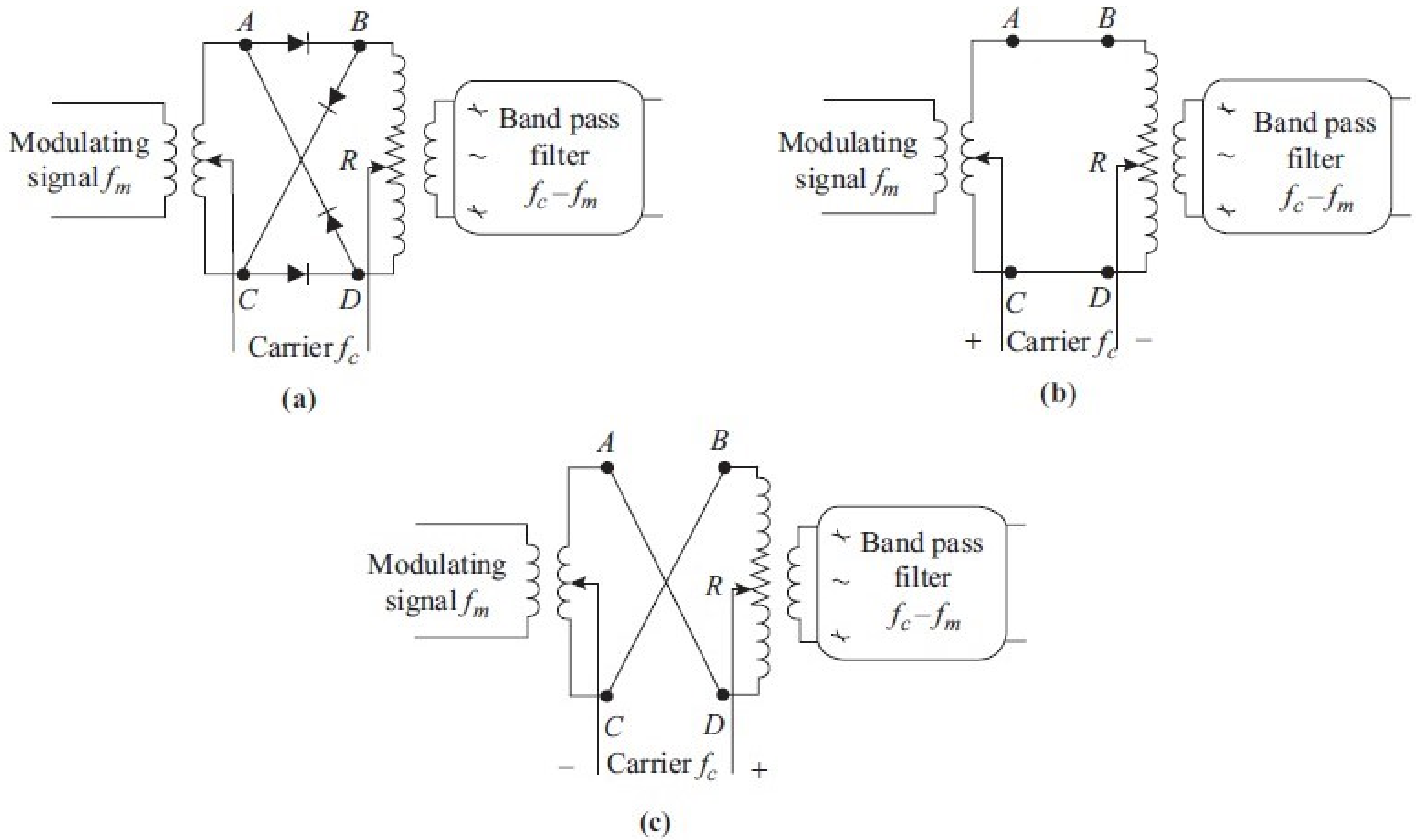


Figure 9.3.2 (a) Double-balanced ring modulator; (b) the conducting paths when diodes *AB* and *CD* are forward biased; (c) the conducting paths when diodes *BC* and *DA* are forward biased.

Credit: Electronic Communication by Roddy and Coolen

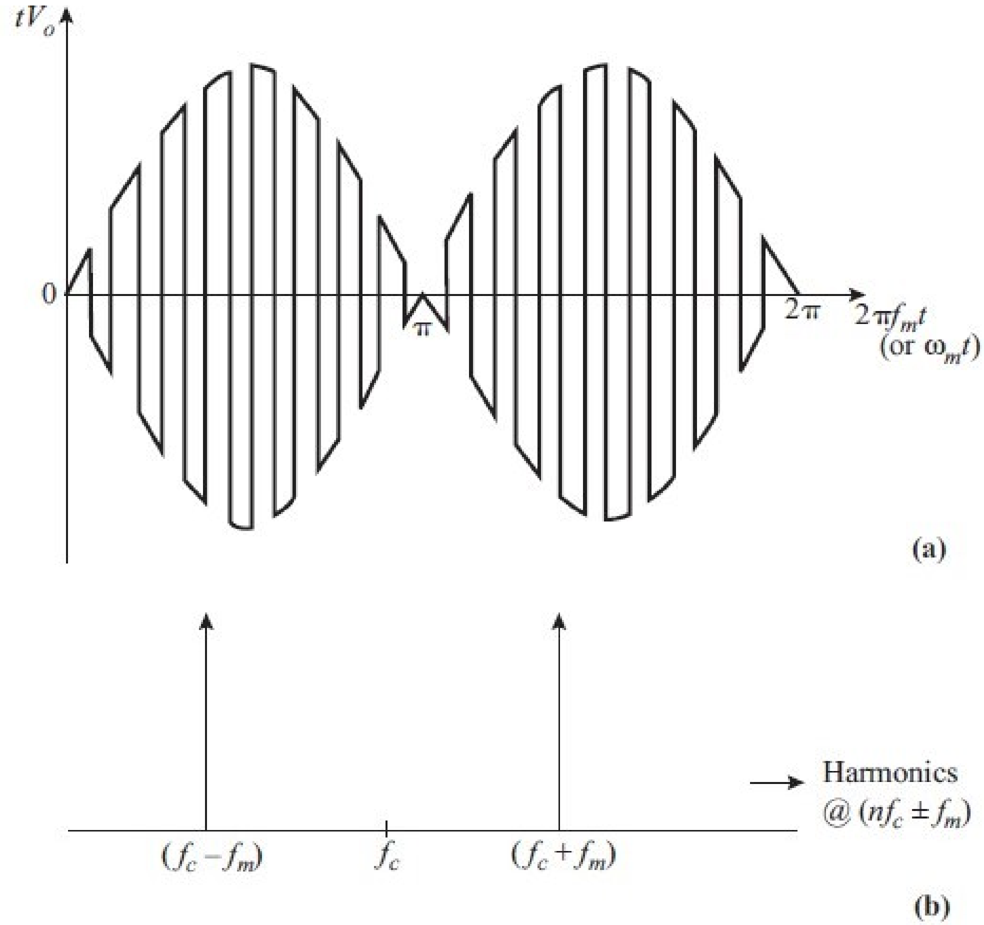


Figure 9.3.3 (a) Time waveform of a DSBSC signal for one cycle of modulation. (b) Spectrum for the signal of (a).

Credit: Electronic Communication by Roddy and Coolen

These circuits have been extensively used for low-frequency telephone applications, where they require balanced input and output transformers and some adjustment of circuit balance for good performance. Care must be exercised if the ring modulator is used at radio frequencies, since the high-level carrier signal may result in the radiation of interference. The doubly balanced diode ring circuit is widely used as a mixer in microwave applications where shielded enclosures prevent radiation.

Credit: Electronic Communication by Roddy and Coolen

Balanced Modulator–Filter Method

Early SSB transmitters used balanced modulator circuits to generate DSBSC signals followed by sideband filters to remove the unwanted sidebands. Such a transmitter is illustrated in Fig. 9.4.1. Initial modulation takes place in the balanced modulator at a low frequency (such as 100 kHz) because of the difficulty of making adequate filters at higher frequencies. The filter is a band-pass filter with a sharp cutoff at each side of the band-pass to obtain satisfactory adjacent sideband rejection. In this case, a single-sideband filter is used, and the carrier oscillator crystal is switched to place the desired sideband in the filter window. Alternatively, two sideband filters (one for each sideband) could be used with a fixed carrier frequency.

The filtered signal is up-converted in a mixer (the second balanced modulator) to the final transmitter frequency and then amplified before being coupled to the antenna. Linear power amplifiers are used to avoid distorting the sideband signal, which might result in regeneration of the second sideband or distortion of the modulated information signal.

The sideband filters are the critical part of this system. Early sideband filters were expensive and did not have the sharp cutoff characteristic required. The integrated ceramic filters available now offer a very inexpensive and effective solution to this problem.

Credit: Electronic Communication by Roddy and Coolen

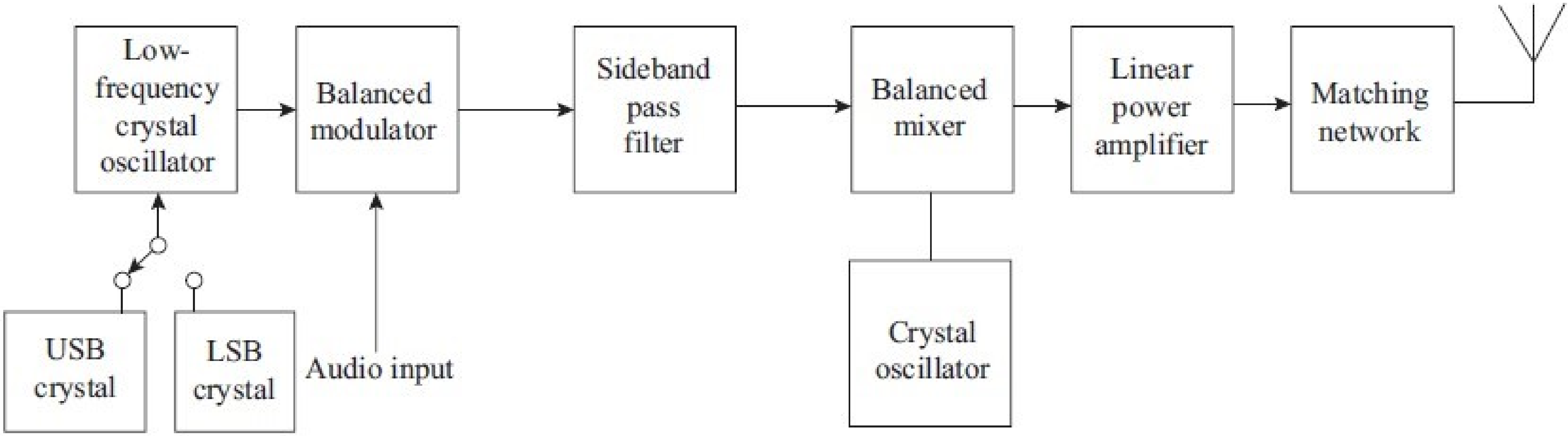


Figure 9.4.1 Single-sideband suppressed carrier transmitter using band-pass filters to eliminate the unwanted sideband.

Credit: Electronic Communication by Roddy and Coolen

Phasing Method

Figure 9.4.2 shows a different means of obtaining an SSBSC signal. This circuit does not have any sideband filters, and the primary modulation can be done at the transmitting frequency. It relies on phase shifting and cancellation to eliminate the carrier and the unwanted sideband.

Assume cosinusoidal signals for both carrier and modulation and that the circuit shown produces the lower side frequency, given by

$$e_{\mathrm{LSF}} = E_{L\,\max} \cos(\omega_c - \omega_m)t \tag{9.4.1}$$

The standard trigonometric identity for the difference of two angles gives

$$e_{\mathrm{LSF}} = E_{L\,\max}[\cos \omega_c t \cos \omega_m t + \sin \omega_c t \sin \omega_m t] \tag{9.4.2}$$

but

$$\sin \omega_c t = \cos\left(\omega_c t - \frac{\pi}{2}\right) \tag{9.4.3}$$

$$\sin \omega_m t = \cos\left(\omega_m t - \frac{\pi}{2}\right) \tag{9.4.4}$$

Credit: Electronic Communication by Roddy and Coolen

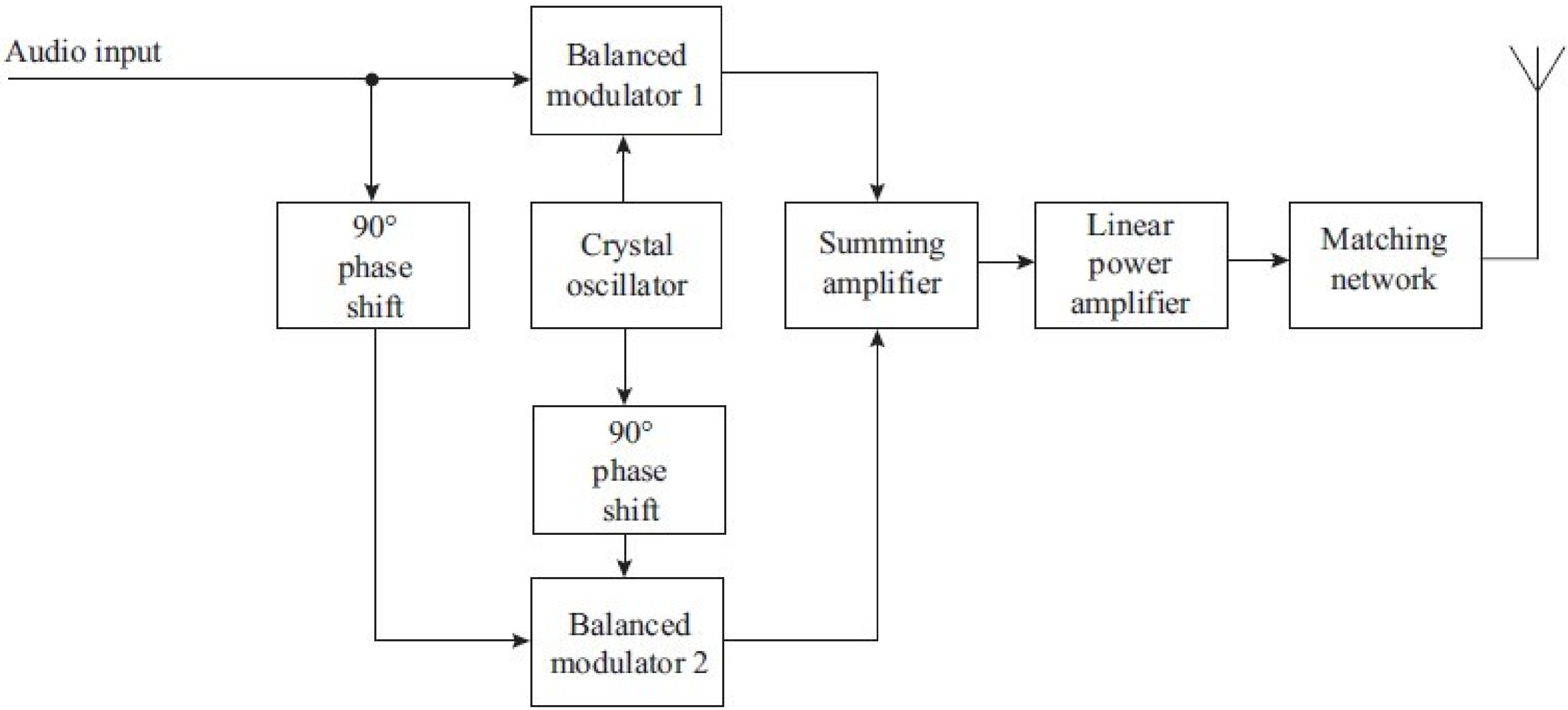


Figure 9.4.2 SSB suppressed carrier transmitter using phase shift to obtain cancellation of sidebands.

Credit: Electronic Communication by Roddy and Coolen

Therefore,

$$e_{\text{LSF}} = E_{L\,\max}\left[\cos \omega_c t \cos \omega_m t + \cos\left(\omega_c t - \frac{\pi}{2}\right)\cos\left(\omega_m t - \frac{\pi}{2}\right)\right] \tag{9.4.5}$$

The first term on the right of Eq. (9.4.5) is the result of balanced modulator 1, which multiplies the two unshifted signals. The second term is the result of balanced modulator 2, which multiplies the two signals, each shifted by $-90°$. The $-90°$ shift for the carrier is easily accomplished by feeding the signal through a controlled current source (transconductance amplifier) into a capacitor. The phase shifting network for the baseband signal must accurately provide a constant $90°$ phase shift over a wide frequency range. Such circuits are tricky to build.

The carrier signal is canceled out in this circuit by both of the balanced modulators, and the unwanted sidebands cancel at the output of the summing amplifier. It is left as an exercise for the student to expand the outputs of the two balanced modulators into sideband form and show that the cancellation does occur on summing. The two outputs are summed to produce the lower sideband signal.

Examination of the trigonometric identity shows that if the two outputs are subtracted instead of added the upper sideband will result, since

$$\begin{aligned} e_{\text{USF}} &= E_{U\,\max} \cos(\omega_c + \omega_m)t \\ &= E_{U\,\max}\,[\cos \omega_c t \cos \omega_m t - \sin \omega_c t \sin \omega_m t] \end{aligned} \tag{9.4.6}$$

Credit: Electronic Communication by Roddy and Coolen

While the system is more complex than one using filters, the individual circuits are quite straightforward, and by using integrated-circuit balanced modulators, very little adjustment is required. Only a simple band-pass filter to remove any harmonics is required in the output before application to the final transmitter amplifier.

It should be noted that the modulation signal is usually a broad band of frequencies of varying amplitudes, which the modulator system must not distort. If the capacitor–transconductance amplifier combination causes such distortion, complete cancellation of the unwanted sideband will not occur, and the wanted sideband will have distortions introduced into it. The third method described next eliminates this problem.

Credit: Electronic Communication by Roddy and Coolen

Third Method

The third method of generating SSBSC modulation is attributed to D. K. Weaver and was developed during the 1950s. It is similar to the phase shifting method presented previously, but it differs in that the modulating signal is first modulated on a low-frequency subcarrier (including phase shifts), which is then modulated onto the high-frequency carrier.

The circuit connections for generating an LSB signal are shown in Fig. 9.4.3. Modulators $BM1$ and $BM2$ both have the unshifted modulating signal as inputs. $BM1$ also takes the low-frequency subcarrier with a 90° shift introduced in it from the oscillator signal. $BM2$ takes the subcarrier signal directly from the oscillator. Assuming unity magnitudes and cosinusoidal single-frequency modulation, the output from $BM1$ becomes

$$\begin{aligned} e_{BM1} &= \cos\left(\omega_o t + \frac{\pi}{2}\right)\cos \omega_m t \\ &= \frac{1}{2}\left[\cos\left(\omega_o t + \omega_m t + \frac{\pi}{2}\right) + \cos\left(\omega_o t - \omega_m t + \frac{\pi}{2}\right)\right] \end{aligned} \tag{9.4.7}$$

and the output of $BM2$ becomes

$$e_{BM2} = \cos \omega_o t \cos \omega_m t + \frac{1}{2}[\cos(\omega_o t + \omega_m t) + \cos(\omega_o t - \omega_m t)] \tag{9.4.8}$$

Credit: Electronic Communication by Roddy and Coolen

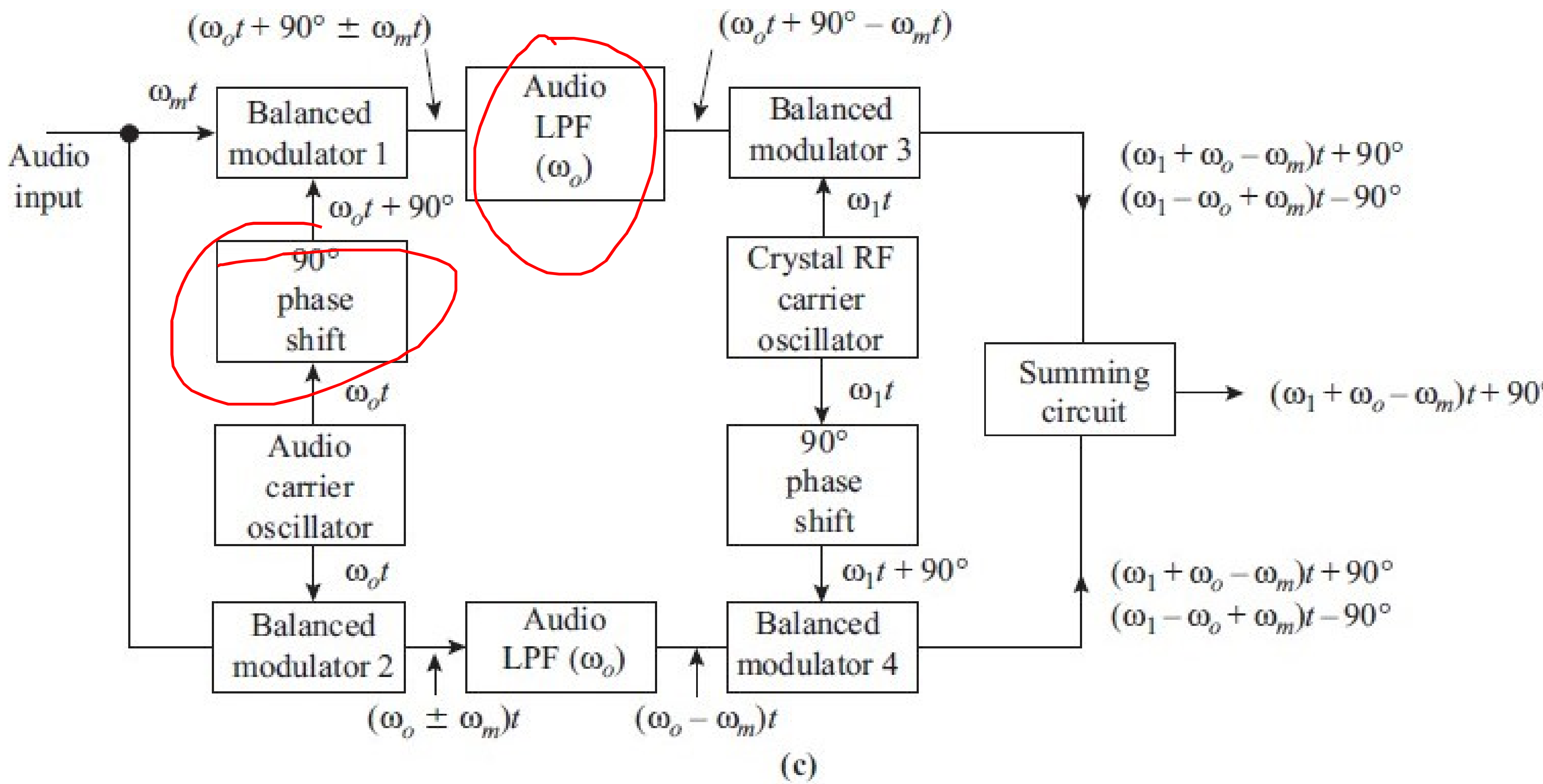


Figure 9.4.3 The "third method" of generating an SSBSC signal.

Credit: Electronic Communication by Roddy and Coolen

Low-pass filters with a cutoff frequency set at the subcarrier frequency f_o removes the sum (the first) term from each of the above signals, leaving only the second (difference) terms as inputs to $BM3$ and $BM4$. These signals are the lower sidebands on f_o. They are identical except that the signal applied to $BM3$ is shifted by $+90^\circ$ from that applied to $BM4$. This process eliminates the need to provide a wideband 90° phase shifting network for the baseband signals, as was the case for the phase shifting method.

Credit: Electronic Communication by Roddy and Coolen

The high-frequency oscillator signal at f_1 is applied directly to $BM3$, but it is shifted by $+90°$ before being applied to $BM4$. The output from $BM3$ becomes

$$e_{BM3} = \cos \omega_1 t \cos\left((\omega_o - \omega_m)t + \frac{\pi}{2} \right) \tag{9.4.9a}$$

$$e_{BM3} = \frac{1}{2}\left[\cos\left(\omega_1 t + \left((\omega_o - \omega_m)t + \frac{\pi}{2} \right) \right) + \cos\left(\omega_1 t - \left((\omega_o - \omega_m)t + \frac{\pi}{2} \right) \right) \right] \tag{9.4.9b}$$

$$e_{BM3} = \frac{1}{2}\left[\cos\left((\omega_1 + \omega_o)t - \omega_m t + \frac{\pi}{2} \right) + \cos\left((\omega_1 - \omega_o)t + \omega_m t - \frac{\pi}{2} \right) \right] \tag{9.4.9c}$$

and the output of $BM4$ becomes

$$e_{BM4} = \cos\left(\omega_1 t + \frac{\pi}{2} \right) \cos(\omega_o - \omega_m)t \tag{9.4.10a}$$

$$e_{BM4} = \frac{1}{2}\left[\cos\left(\left(\omega_1 t + \frac{\pi}{2} \right) + (\omega_o - \omega_m)t \right) + \cos\left(\left(\omega_1 t + \frac{\pi}{2} \right) - (\omega_o - \omega_m)t \right) \right] \tag{9.4.10b}$$

$$e_{BM4} = \frac{1}{2}\left[\cos\left((\omega_1 + \omega_o)t - \omega_m t + \frac{\pi}{2} \right) + \cos\left((\omega_1 - \omega_o)t + \omega_m t + \frac{\pi}{2} \right) \right] \tag{9.4.10c}$$

Credit: Electronic Communication by Roddy and Coolen

The first terms in Eqs. (9.4.9c) and (9.4.10c) are identical lower sidebands on an offset carrier frequency $f_c = f_1 + f_o$. The second terms are the upper sidebands on an offset carrier at $f_c = f_1 - f_o$, but are 180° out of phase with each other. The oscillator frequency f_1 must be adjusted so that the output carrier frequency f_c and the desired sideband fall in the correct position in the output frequency spectrum.

Usually, f_o is chosen to fall at the midpoint of the modulating signal baseband, so that $f_o = W/2$. The result is that both the USB and LSB spectrums are centered on f_1, with the carrier position f_c for the LSB located at the upper edge of the pass-band and that for the USB at the lower edge. This is illustrated in Fig. 9.4.4.

The outputs from $BM3$ and $BM4$ are added in a summing amplifier to produce the final output. The first two terms add, but the second two cancel, leaving the output as

$$e_{\text{out}} = \cos\left((\omega_1 + \omega_o - \omega_m)t + \frac{\pi}{2}\right) \tag{9.4.11}$$

Credit: Electronic Communication by Roddy and Coolen

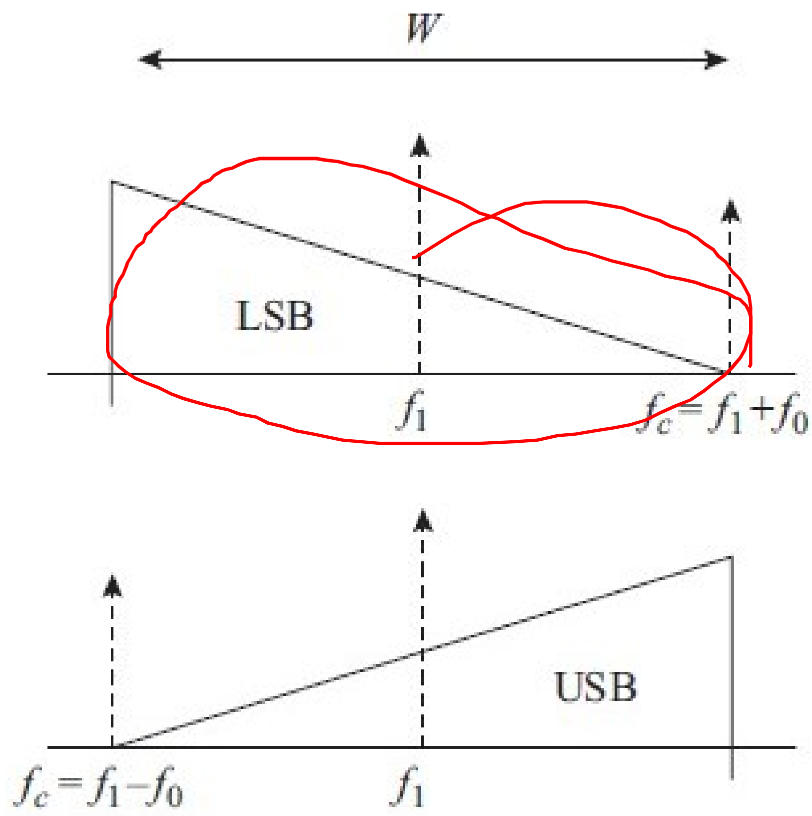


Figure 9.4.4 Output spectra for the "third-method" circuit (a) for LSB and (b) for USB.

Credit: Electronic Communication by Roddy and Coolen

This is the lower sideband on the carrier frequency ($f_1 + f_o$). The $+90°$ shift in the output is of no consequence since the original carrier has been eliminated. This signal may be applied to a linear power amplifier and antenna for radiation. Modulation in *BM*3 and *BM*4 may take place at the final transmission frequency so that no further conversion is needed.

If the output from *BM*3 (or *BM*4) is inverted before the input to the adder, the phasing becomes such that the first terms cancel and the second terms add, giving the upper sideband on the carrier frequency ($f_1 - f_o$).

Credit: Electronic Communication by Roddy and Coolen

Vestigial Sideband Transmission

- It has been stressed that the major advantage of single sideband is the bandwidth saving that accrues from its use, although the power saving cannot be ignored.
- Some demodulation complications arise from the use of SSB, as opposed to AM systems in which a carrier is sent.
- The greater the bandwidth occupied by a signal, the greater is the spectrum space that can be saved by sending one sideband instead of both.
- Finally, the more information that must be sent in a given time, i.e., per second, the larger the bandwidth required to send it.

Credit: Electronic Communication System by Kennedy and Davis

Vestigial Sideband Transmission

- We may now tum to the question of transmitting the video signals required for the proper reception of television, noting that the bandwidth occupied by such signals is at least 4 MHz.
- Bearing in mind filter characteristics, a transmitted bandwidth of 9 MHz would be the minimum requirement if A3E video transmissions were used.
- The use of some form of SSB is clearly indicated here to ensure spectrum conservation. So as to simplify video demodulation in the receiver, the carrier is, in practice, sent undiminished. Because the phase response of filters, near the edges of the flat bandpass, would have a harmful effect on the received video signals in a TV receiver, a portion of the unwanted (lower) sideband must also be transmitted.

Credit: Electronic Communication System by Kennedy and Davis

Vestigial Sideband Transmission

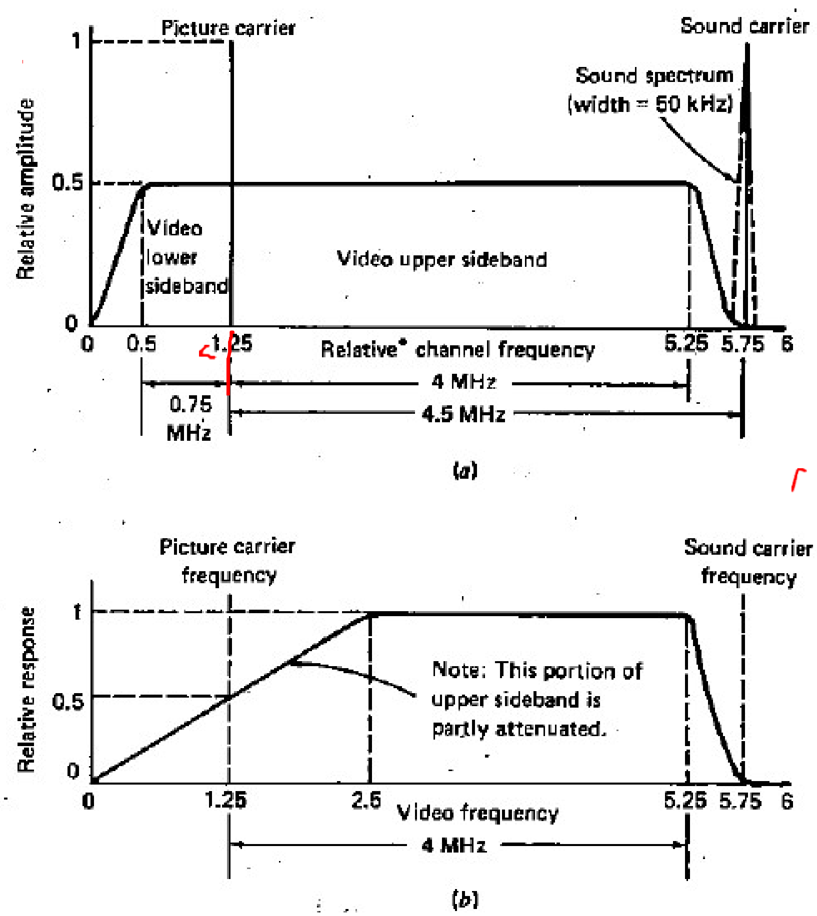


FIGURE Vestigial sideband as used for TV video transmission.
(a) Spectrum of transmitted signals (NTSC);
(b) corresponding receiver video amplifier frequency response.

Credit: Electronic Communication System by Kennedy and Davis

Vestigial Sideband Transmission

- The result is vestigial sideband transmission, or C3F, as shown in Figure .
- Note that the frequencies shown there, like the ones used in text, refer strictly only to the NTSC TV system in use in the United States, Canada and Japan.
- The principles are the same, but the frequencies are somewhat different in the PAL TV system used in Europe, Australia and elsewhere, and again different in the French SECAM system.

Credit: Electronic Communication System by Kennedy and Davis

Vestigial Sideband Transmission

- By sending the first 1.25 MHz of the lower sideband, it is possible to make sure that the lowest frequencies in the wanted upper sideband are not distorted in phase by the vestigial 'sideband filter.
- Because only the first 1.25 MHz of the lower sideband is transmitted, 3 MHz of spectrum is saved for every TV channel.
- Since the total bandwidth requirement of a television channel is now 6 MHz instead of 9 MHz, clearly a great saving has been made, and more channels consequently can be accommodated.

Credit: Electronic Communication System by Kennedy and Davis

Vestigial Sideband Transmission

To complete the illustration, Figure *a* shows the location, in frequency of the frequency-modulated sound transmissions that accompany the video.

It should be noted that these transmissions have nothing to do with the fact that the modulation system for video is C3F, and would have been there regardless of the video modulation system.

All these signals occupy frequencies near the video transmissions simply because sound is required with the pictures, and it would not be very practical to have a completely separate receiver for the sound, operating at some frequency remote from the video transmitted frequencies.

Credit: Electronic Communication System by Kennedy and Davis

Vestigial Sideband Transmission

- Figure *b* shows the video frequency response of the television receiver.
- Attenuation is purposely provided for the video frequencies from 0 to 1.25 MHz
- Extra power is transmitted at these frequencies.
- These frequencies would be unduly emphasized in the video[output of the receiver if they were not attenuated appropriately.

Credit: Electronic Communication System by Kennedy and Davis