12

Mixers

Mixers shift signals from one frequency to another. In radios, mixers allow us to move a signal from the operating frequency that we use for transmission and reception to an audio frequency that we can hear. Mixers are fundamentally more complicated than amplifiers and oscillators because there are inputs and outputs at several frequencies. Because of the different frequencies that go in and out, mixers need filters to choose the frequencies that we want. Figure 12.1a shows the schematic symbol for a mixer. It is traditional to call the input the *RF* signal, for radio frequency. In addition, an oscillator signal, the *Local Oscillator*, or *LO*, is applied to the mixer. It is called the local oscillator because it is generated in the receiver itself, that is, locally. The output is called the *IF*, for *Intermediate Frequency*, and it is either at the sum of the RF and LO frequencies or at the difference. We characterize a mixer by a conversion gain G, which is defined the same way as for an amplifier, except that the input and output frequencies are different:

$$G = P/P_{+}, \tag{12.1}$$

where P is the IF output power and P_{+} is the available power from the RF source.

Our receiver uses two mixers (Figure 12.1b). In the first mixer, the RF Mixer, the RF input mixes with the VFO to produce the IF. This IF is in turn the input for the second mixer, the Product Detector. There it mixes with the Beat-Frequency Oscillator, or BFO, to produce the audio output. This type of receiver, with an IF, is called a *superheterodyne* receiver, in contrast to the *direct-conversion* receiver, which has only a product detector. The superheterodyne receiver has the advantage that filters can reduce all of the spurious responses. We will measure the spurious responses in the exercises. The transmitter uses only one mixer (Figure 12.1c), where the VFO and the Transmit Oscillator mix to produce the output RF signal.

12.1 Gilbert Cell

The mixers in the NorCal 40A are called *Gilbert Cells*. This circuit was invented by Barrie Gilbert in 1967, and it is one of the most important circuits in

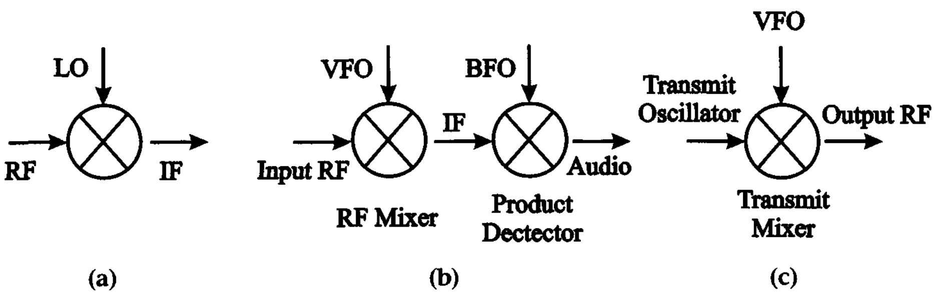

Figure 12.1. (a) Mixer schematic symbol and notation for the inputs and outputs. (b) The NorCal 40A receiver mixers. (c) The NorCal 40A Transmit Mixer.

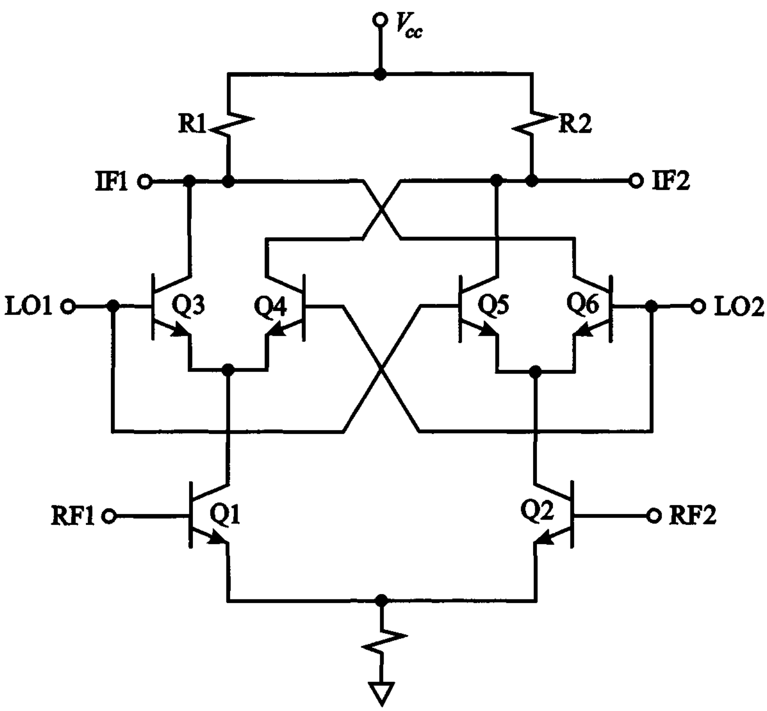

Figure 12.2. The Gilbert-cell mixer.

communications electronics. Figure 12.2 shows the mixer. It is complicated because there are some crossing wires. It may be easiest to start at the bottom, which is a long-tailed pair differential amplifier, with the RF signals as inputs. However, inserted between the collectors and the collector resistors are four cross-coupled transistors. These are driven by the local oscillator. To understand the effect of the local oscillator assume that the voltage at LO1 is large, so that Q3 and Q5 turn on. This connects Q1 to R1 and Q2 to R2, and we have a normal differential amplifier. Now consider what happens if the voltage at LO2 is large, so that Q4 and Q6 are turned on. This connects Q1 to R2 and Q2 to R1. We have a differential amplifier again, but the outputs are interchanged. This changes the sign of the outputs. The effect is that the output is alternately multiplied by +1 and −1 depending on the sign of the local oscillator.

The Gilbert Cell is an *active* mixer, which combines an amplifier with a mixer. We contrast active mixers with diode mixers that have loss. The gain from active mixers provides a big advantage because it reduces the gain that we need in the rest of the circuit. In addition, active mixers are tolerant of large reactances in the IF output lines. This is useful because IF crystal filters have large reactances at stop-band frequencies. The disadvantage of active mixers is that they can overload easily. Diode mixers, in contrast, are less likely to overload, but they require careful attention in the IF circuit and much more local-oscillator power than active mixers.

12.2 Mixer Mathematics

In a mixer, the LO drives transistor switches that steer the input to the output alternately in phase and out of phase. Mathematically this is like multiplying the input by a square wave. Notice that this is a linear network, in that we are always multiplying by +1 or −1. However, we say it is *time varying*, because the multiplier changes with time. Linear time-varying networks can shift the frequency of a signal, in contrast to linear *time-invariant* networks that leave the frequency unchanged. In practice, the switches do not change phase instantaneously, and so the input is really multiplied by a waveform that is somewhere between a square wave and a sine wave. But to keep the mathematics simple we will use a square wave. It is easy to sketch the output waveform (Figure 12.3) but not easy to perceive the different frequency components. We start with an input RF cosine signal, which we write as

$$V_{rf}(t) = V_{rf}\cos(\omega_{rf}t). \tag{12.2}$$

The Fourier components for the LO are given in Appendix B, Section 2:

$$V_{lo}(t) = \frac{4}{\pi}\left(\cos(\omega_{lo}t) - \frac{\cos(3\omega_{lo}t)}{3} + \frac{\cos(5\omega_{lo}t)}{5} - \cdots\right). \tag{12.3}$$

We multiply the two expressions

$$V(t) = V_{rf}(t) \cdot V_{lo}(t) \tag{12.4}$$

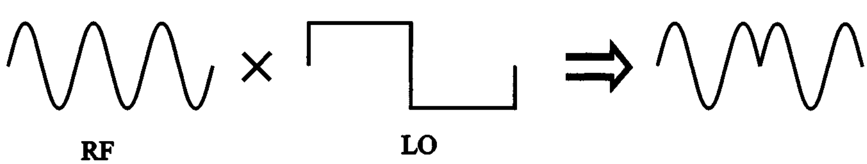

Figure 12.3. Mixing as the multiplication of an input RF cosine wave by a square-wave local oscillator.

and expand to get

$$V(t) = \frac{2V_{rf}}{\pi}\left(\cos(\omega_{lo}t - \omega_{rf}t) - \frac{\cos(3\omega_{lo}t - \omega_{rf}t)}{3} + \frac{\cos(5\omega_{lo}t - \omega_{rf}t)}{5} - \cdots\right)$$
$$+ \frac{2V_{rf}}{\pi}\left(\cos(\omega_{lo}t + \omega_{rf}t) - \frac{\cos(3\omega_{lo}t + \omega_{rf}t)}{3} + \frac{\cos(5\omega_{lo}t + \omega_{rf}t)}{5} - \cdots\right). \tag{12.5}$$

From these sums, we can identify the sum and difference frequency terms V_+ and V_-:

$$V_+(t) = \frac{2V_{rf}}{\pi}\cos(\omega_+ t), \tag{12.6}$$

$$V_-(t) = \frac{2V_{rf}}{\pi}\cos(\omega_- t), \tag{12.7}$$

where

$$\omega_+ = \omega_{lo} + \omega_{rf}, \tag{12.8}$$

$$\omega_- = |\omega_{lo} - \omega_{rf}|. \tag{12.9}$$

I have added absolute-value signs for the difference frequency, because ω_{lo} can be larger or smaller than ω_{rf}, and we work with positive frequencies. These are the mixer products for the fundamental frequency component of the local oscillator. Either the sum or the difference frequency could be used for the IF. We would need a filter to remove the one that we do not want. We can also pick out the sum and difference frequency terms for the third harmonic of the LO, V_{3+} and V_{3-}:

$$V_{3+}(t) = -\frac{2V_{rf}}{3\pi}\cos(\omega_{3+} t), \tag{12.10}$$

$$V_{3-}(t) = -\frac{2V_{rf}}{3\pi}\cos(\omega_{3-} t), \tag{12.11}$$

where

$$\omega_{3+} = 3\omega_{lo} + \omega_{rf}, \tag{12.12}$$

$$\omega_{3-} = |3\omega_{lo} - \omega_{rf}|. \tag{12.13}$$

The theory predicts that these components would be a factor of three below the fundamental components, but usually they would be less than this. We could go on to the higher harmonics, but life is short. This is a complicated set of components. Unfortunately, the real picture is even more complicated. In this analysis, there are no products with even-order harmonics of the LO, and there are no components at the RF or LO frequencies or their harmonics. However, in a real mixer, all these will be present to some extent.

12.3 Spurious Responses

We have seen that when an input sine wave mixes with a square-wave local oscillator, we generate many frequency components. In a receiver, we usually ask this question in another way. An antenna receives a great many signals on different frequencies simultaneously. We ask, what frequencies in addition to the RF frequency give an output at the IF frequency? Signals at these other frequencies interfere with the signal that we want, and the outputs are called *spurious responses*, or spurs for short. For example, consider the frequencies shown in Figure 12.4. In the figure, the LO is below the RF, and the IF is difference frequency. These are the relationships between the frequencies in the RF Mixer of the NorCal 40A, but other choices are possible.

Now consider the difference of f_{lo} and f_{if}. This is called the *image frequency*, because the RF signal and the image sit symmetrically around the IF frequency. We write the image frequency f_i as

$$f_i = f_{if} - f_{lo}. \tag{12.14}$$

We can rearrange the formula to get

$$f_{if} = f_{lo} + f_i. \tag{12.15}$$

This formula means that if a signal at the image frequency is present at the mixer input, it will produce a spurious output at the IF, just like an input at the RF frequency. We suppress this spurious response by adding a band-pass filter that passes an input at the RF but blocks the image (Figure 12.4).

Signals can also cause a spurious response by mixing with the third harmonic of the LO. Consider a signal at the frequency given by the difference

$$f_{3\downarrow} = 3f_{lo} - f_{if}. \tag{12.16}$$

This frequency is indicated on the left in Figure 12.5. We can write

$$f_{if} = 3f_{lo} - f_{3\downarrow}, \tag{12.17}$$

and this means that $f_{3\downarrow}$ will cause a spurious response. Similarly, we can look at

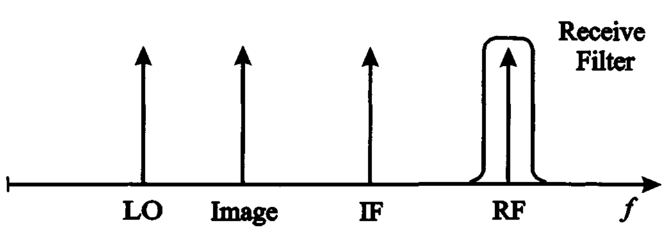

Figure 12.4. Mixer frequencies and the image frequency. A band-pass filter centered on the RF frequency prevents a signal at the image frequency from reaching the mixer.

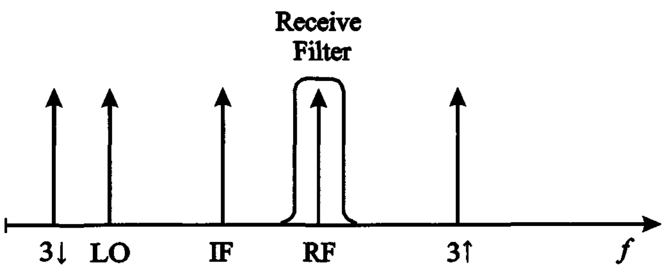

Figure 12.5. Spurious responses caused by mixing with the third harmonic of the LO.

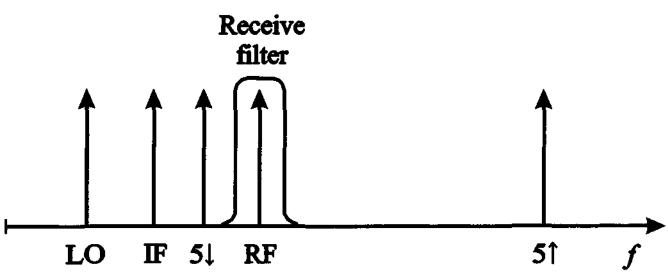

Figure 12.6. Spurious responses caused by mixing with the fifth harmonic of the LO.

the sum

$$f_{3\uparrow} = 3f_{lo} + f_{if}. \tag{12.18}$$

This frequency is on the right in Figure 12.5. It will also produce a spurious response. In the NorCal 40A, the spurs from the third harmonic of the VFO are well away from the RF frequency, and thus the band-pass filter does a good job of suppressing them. However, the lower frequency $f_{3\downarrow}$ is in the frequency band for AM radio stations, which are often close and powerful. It is easy to hear a spurious response in the NorCal 40A from them.

Finally, we consider the spurs from the fifth harmonic of the LO (Figure 12.6). We write

$$f_{5\downarrow} = 5f_{lo} - f_{if}, \tag{12.19}$$

$$f_{5\uparrow} = 5f_{lo} + f_{if}. \tag{12.20}$$

In the NorCal 40A, $f_{5\uparrow}$ is high, and the band-pass filter does a good job of rejecting it. Unfortunately, $f_{5\downarrow}$ is close to f_{rf}, and it is hard for the band-pass filter to reject it. The frequency $f_{5\downarrow}$ is one of the largest spurious responses.

12.4 Broad-Band Receivers

The NorCal 40A VFO can tune only over a range of about 50 kHz. This is fine for the frequency range that it was intended for, but for many applications, much

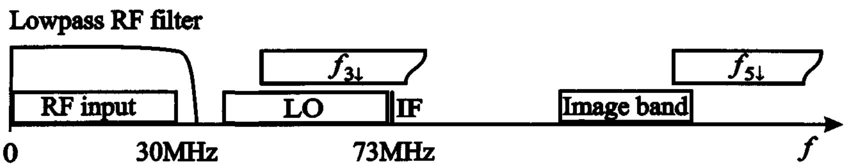

Figure 12.7. RF, LO, IF, and image frequencies for the first IF at 73 MHz in the Kenwood 850 transceiver.

broader bandwidths are needed. However, even assuming that the VFO could be redesigned for a broader bandwidth, the receiver would be limited by spurious responses. If we tried to increase the tuning range significantly, $f_{5\downarrow}$ would enter the pass band of the RF Filter. We can, however, make a receiver cover a much broader bandwidth without spurious responses if we use an IF frequency that is much higher than the RF frequency. At first glance, this does not seem like a great idea. A higher IF frequency takes us farther away from the audio frequencies that are our eventual goal. However, higher IF frequencies have the great advantage that the spurs are even higher in frequency. This means that we can block all of them with a low-pass filter. In practice, these receivers need to use more than one IF frequency, and this allows additional filter stages for better spur rejection.

As an example of a broad-band design, we consider the Kenwood 850, an extremely popular transceiver that covers the frequency range from 100 kHz to 30 MHz. It has three IFs, and so we call this a *triple-conversion* receiver. The first is at 73 MHz, the second at 8.83 MHz, and the third at 455 kHz. Figure 12.7 shows the RF input and LO range for the first IF. We can calculate the LO frequencies that are required by subtraction:

$$f_{lo} = f_{if} - f_{rf}. \tag{12.21}$$

The maximum LO frequency $f_{\overline{lo}}$ comes at the minimum RF frequency $f_{\underline{rf}}$. We write it as

$$f_{\overline{lo}} = f_{if} - f_{\underline{rf}} = 73\,\text{MHz} - 0.1\,\text{MHz} = 72.9\,\text{MHz}. \tag{12.22}$$

The minimum LO frequency $f_{\underline{lo}}$ comes at the maximum RF frequency $f_{\overline{rf}}$. We get

$$f_{\underline{lo}} = f_{if} - f_{\overline{rf}} = 73\,\text{MHz} - 30\,\text{MHz} = 43\,\text{MHz}. \tag{12.23}$$

We can calculate the corresponding maximum and minimum image frequencies as sums:

$$f_{\overline{i}} = f_{if} + f_{\overline{lo}} = 73\,\text{MHz} + 72.9\,\text{MHz} = 145.9\,\text{MHz}, \tag{12.24}$$

$$f_{\underline{i}} = f_{if} + f_{\underline{lo}} = 73\,\text{MHz} + 43\,\text{MHz} = 116\,\text{MHz}. \tag{12.25}$$

The image frequencies are far away from the input RF frequencies, and we can easily remove them with a low-pass filter.

The harmonic spurs $f_{3\downarrow}$ and $f_{5\downarrow}$ are also quite high. We write them as

$$f_{3\downarrow} = 3f_{lo} - f_{if}, \tag{12.26}$$

$$f_{5\downarrow} = 5f_{lo} - f_{if}. \tag{12.27}$$

The minimum spur frequencies come at the minimum LO frequencies:

$$\underline{f_{3\downarrow}} = 3\underline{f_{lo}} - f_{if} = 129\,\text{MHz} - 73\,\text{MHz} = 56\,\text{MHz}, \tag{12.28}$$

$$\underline{f_{5\downarrow}} = 5\underline{f_{lo}} - f_{if} = 215\,\text{MHz} - 73\,\text{MHz} = 142\,\text{MHz}. \tag{12.29}$$

These are well above the RF frequency band, and an input low-pass filter can remove them.

12.5 Key Clicks

Transmitter pulses have controlled rise and fall times to prevent clicking sounds on nearby frequencies. These are called *key clicks*, and they interfere with other operators. Key clicks point to an important problem in commercial communications. Companies want to have as many channels as possible in their frequency allocation, because they make more money when there are more channels. If the transmitter transitions are too fast, the power spreads in frequency, and the channel spacing must be large to avoid interference. We can understand this effect by using our mixer mathematics. We start with transmitter pulses that have zero rise and fall times (Figure 12.8). We will let the peak voltage be 2 V, and assume that the "on" and "off" time are the same. We let the load resistance be 1 Ω so that we can calculate the power easily. The power when the pulse is on is 2 W, and the average power is 1 W.

To find the frequency components, we multiply the carrier, $2\cos(\omega t)$, by the Fourier series for rectangular pulses with a maximum value of one and a minimum value of zero. We write the pulses as

$$V_r(t) = \frac{1}{2} + \frac{2}{\pi}\left(\cos(\omega_k t) - \frac{\cos(3\omega_k t)}{3} + \frac{\cos(5\omega_k t)}{5} - \cdots\right), \tag{12.30}$$

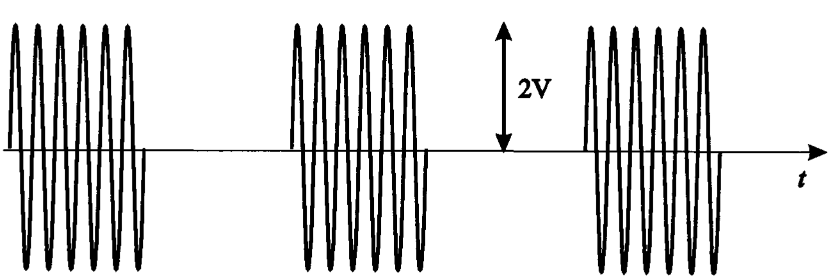

Figure 12.8. Transmitter pulses with zero rise and fall times.

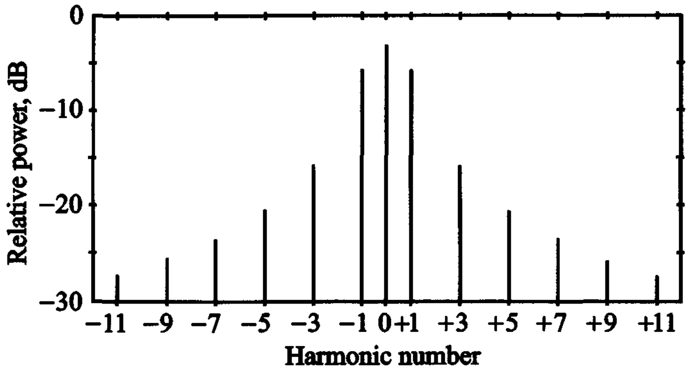

Figure 12.9. Spectrum for a keyed transmitter with zero rise and fall times, in dB relative the total power.

where $\omega_k = 2\pi f_k$ is the keying frequency. These coefficients are derived in Appendix B, Section 2. The product can be written as

$$\begin{aligned} V(t) = \cos(\omega t) &+ \frac{2}{\pi}\left(\cos(\omega t - \omega_k t) - \frac{\cos(\omega t - 3\omega_k t)}{3} + \frac{\cos(\omega t - 5\omega_k t)}{5} - \cdots\right) \\ &+ \frac{2}{\pi}\left(\cos(\omega t + \omega_k t) - \frac{\cos(\omega t + 3\omega_k t)}{3} + \frac{\cos(\omega t + 5\omega_k t)}{5} - \cdots\right). \end{aligned} \tag{12.31}$$

The largest frequency component is at the carrier frequency ω. The power in this component is 1/2. In addition there are sidebands around the carrier frequency that are caused by keying the transmitter on and off. These are shown in Figure 12.9. The horizontal axis is frequency, and the vertical axis is power in dB relative to the total power transmitted. This plot of frequency components is called a *spectrum*.

We get an interesting result if we add the power for each component. This is just the total power, which is one watt:

$$1 = \frac{1}{2} + 2 \cdot \frac{2}{\pi^2}\left(1 + \frac{1}{9} + \frac{1}{25} + \frac{1}{49} + \cdots\right). \tag{12.32}$$

The factor of 2 for the right term comes from the fact that we are summing two identical series, one for the upper sidebands and one for the lower sidebands. We can invert this formula to find an expression for the sum

$$\frac{\pi^2}{8} = 1 + \frac{1}{9} + \frac{1}{25} + \frac{1}{49} + \cdots = \sum_{n\,\mathrm{odd}} \frac{1}{n^2}. \tag{12.33}$$

This is an elegant way to derive this formula.

For isolated frequency components such as transmitter harmonics, it makes sense to specify a maximum power level. For example, in the United States, the FCC requires that for transmitters like the NorCal 40A, with power levels of 5 W or below, the spurious components should be at least 30 dB below the carrier.

However, this kind of specification does not make much sense for keying sidebands, because a receiver may pick up many components at one time. The effect of any one component may be negligible, but the cumulative effect of all the components may be a problem. For keying sidebands, the corresponding question is: How wide does the channel need to be so that the power spilling out at frequencies above the channel is 30 dB below the transmitted power? Since the sidebands are symmetrical, the power below the channel would also be 30 dB down. We write the proportion of the sideband power at higher frequencies p as

$$p = \frac{2}{\pi^2}\left(\frac{1}{(n+1)^2} + \frac{1}{(n+3)^2} + \cdots\right), \tag{12.34}$$

where n is the number of the keying harmonic that marks the channel boundary. For large harmonic numbers, we can approximate this sum by an integral

$$p \approx \frac{2}{\pi^2} \cdot \frac{1}{2} \int_n^\infty \frac{dx}{x^2} = \frac{1}{n\pi^2}. \tag{12.35}$$

The factor of 1/2 before the integral takes into account the fact that only odd keying harmonics are generated. We can rewrite this as

$$n = \frac{1}{\pi^2 p}. \tag{12.36}$$

To find how large the bandwidth needs to be, we set $p=0.001$ and get $n=101$. This means that for 10-Hz keying, we would need a large channel that extends 1 kHz above and 1 kHz below the carrier frequency.

We can reduce the bandwidth dramatically if we increase the rise and fall times. The pulse shapes do not have simple mathematical forms, but we can approximate them by exponentials with a time constant τ (Figure 12.10a). To find the bandwidth that we would need for this waveform, we need to see how the frequency components are affected. One way that we can do this is to consider the frequency response of an RC network that produces these exponential waveforms (Figure 12.10b). This is the delay network we studied in Problem 3. We can

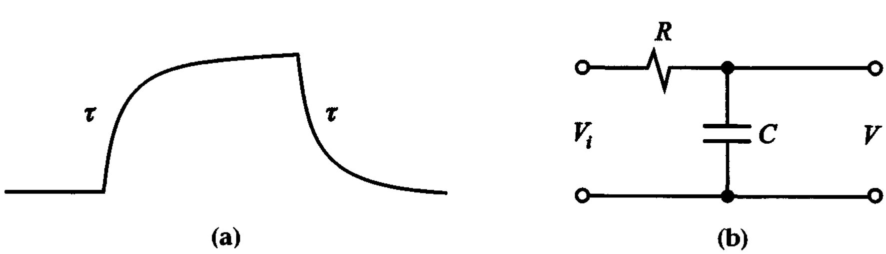

Figure 12.10. Pulse with exponential rise and fall (a). RC network with exponential waveforms (b).

relate the output voltage V to the input voltage V_i by the phasor relation

$$\frac{V}{V_i} = \frac{1}{1 + j\omega\tau}, \tag{12.37}$$

where the time constant is $\tau = RC$. For the nth keying harmonic, we write

$$\frac{V}{V_i} = \frac{1}{1 + jn\omega_k\tau}, \tag{12.38}$$

where n is the number of the keying harmonic. In our calculations, we are interested in the power rather than the voltage, so that we can consider

$$\left|\frac{V}{V_i}\right|^2 = \frac{1}{1 + (n\omega_k\tau)^2}. \tag{12.39}$$

This lets us modify the formula for the outside power p (Equation 12.34):

$$p = \frac{2}{\pi^2}\left(\frac{1}{(n+1)^2}\cdot\frac{1}{1+(n+1)^2(\omega_k\tau)^2} + \frac{1}{(n+3)^2}\cdot\frac{1}{1+(n+3)^2(\omega_k\tau)^2} + \cdots\right). \tag{12.40}$$

We can also approximate this sum by an integral, given by

$$p \approx \frac{1}{\pi^2}\int_n^\infty \frac{1}{x^2}\cdot\frac{1}{1+(x\omega_k\tau)^2}\,dx. \tag{12.41}$$

We will assume that $(x\omega_k\tau)^2 \gg 1$ so that we can write

$$p \approx \frac{1}{\pi^2}\int_n^\infty \frac{dx}{x^4(\omega_k\tau)^2} = \frac{1}{3\pi^2 n^3(\omega_k\tau)^2}, \tag{12.42}$$

or

$$n = (3p(\pi\omega_k\tau)^2)^{-1/3}. \tag{12.43}$$

For $p = 0.001$ we get

$$n \approx (f_k\tau)^{-2/3}. \tag{12.44}$$

Commercial transmitters often use a time constant of $\tau = 3$ ms. For $f_k = 10$ Hz, this gives

$$n = 10. \tag{12.45}$$

This reduces the required channel width dramatically, from 2 kHz to 200 Hz.

FURTHER READING

A good reference for practical details and measurements is "Mixers, modulators, and demodulators," by David Newkirk and Rick Karlquist. This is Chapter 15 in the *ARRL Handbook*, published by the American Radio Relay League. *Mastering Radio Frequency Circuits through Projects and Experiments* by Joseph Carr, published by McGraw-Hill, has a good chapter on the SA602AN mixer that is used in the NorCal 40A. A comprehensive advanced reference is *Microwave Mixers,* by Stephen Maas, published by Artech House.

PROBLEM 28 - RF MIXER

This is the first chance you get to use the SA602AN integrated circuit. This is a classic IC that was developed by the Signetics Corporation. Philips bought Signetics, and so it is now sold under the Philips name. The letters "SA" refer to a Philips product. The "A" suffix indicates that this is an improved version of an earlier part numbered 602. The SA602AN is used in many different communications systems, and it is very worthwhile to study the manufacturer's data sheets for this IC in Appendix D. It contains an amplifier, mixer, oscillator, and regulator, and it operates at frequencies as high as 500 MHz. The current consumption is quite low, only 2.4 mA, and this makes it quite suitable for battery-operated equipment. It costs only about $2. The NorCal 40A uses SA602ANs for the RF Mixer, the Transmit Mixer, and the Product Detector. The block diagram is shown in Figure 12.11. The supply voltage goes to pin 8 and should be between +4.5 and +8 V. There should be a bypass capacitor to ground nearby to prevent oscillations in the supply line. Pins 6 and 7 are the base and emitter for an oscillator transistor. Alternatively, you can connect an external oscillator to pin 6. An external oscillator should have a voltage of at least 200 mVpp. The RF inputs are pins 1 and 2. The LO and the RF signal mix in a Gilbert Cell, and the IF output appears at pins 4 and 5.

We can apply the input at either pin 1 or pin 2. The unused pin will need a bypass capacitor to ground. The input impedance at either pin is nominally 1.5 kΩ, but it actually varies considerably with frequency. Alternatively, the signal can be applied differentially between the two pins. The DC-bias voltage at the inputs is 1.4 V. The bias voltage is generated internally, so that we do not need any external bias circuitry. The circuits applied to the input pins should not draw any current, so that they do not upset

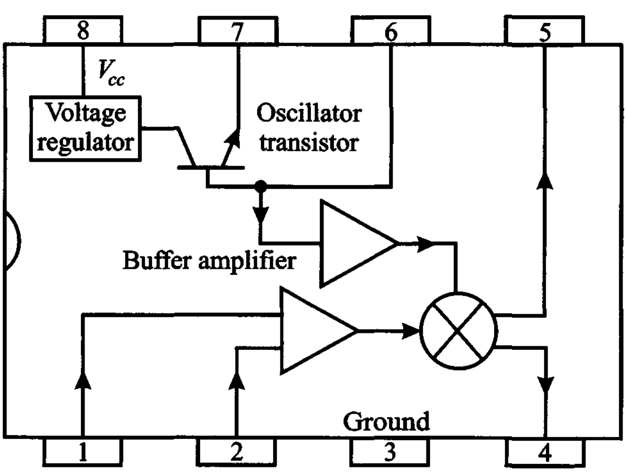

Figure 12.11. Block diagram for the SA602AN mixer.

the bias voltage. The NorCal 40A uses this voltage as the reference voltage for the RIT comparator.

We can take the IF output from either pin 4 or pin 5. Alternatively, since the two IF outputs are 180° out of phase, we can take the difference between the two pins to get an output that is twice as large. Aside from the larger voltage, this approach has other advantages. Often outside signals that couple into the lines will cause similar voltages on both lines, and this means that they do not affect the voltage difference. In addition, even-order spurious components tend to appear on both lines at similar levels, so that if we take the difference, these spurs will drop out. The output impedance for each pin is 1.5 kΩ, or 3.0 kΩ if we use a differential output. The DC voltage at the outputs is about 1.2 V below the supply voltage (6.8 V for an 8-V supply). To avoid upsetting the DC bias, the external input and output circuits should not draw any DC current.

In the RF Mixer, we do not use the internal oscillator but inject the VFO from the outside. We will use the internal oscillator as a fixed crystal oscillator for the Transmit Mixer and the Product Detector, however. First install the Receiver Mixer U1, together with the two bypass capacitors C5 and C8. Install the RF Gain pot R2. You may need to move the transformer T2 out of the way so that R2 and T2 do not touch each other. R2 serves as an attenuator at the input to keep the radio from overloading.

The RF Mixer output goes to the IF Filter, which removes the sum frequency response, and from there to the Product Detector U2. In the previous problem we applied the reference voltage for the RIT comparator to this pigtail. This time we will use the input of U2 to provide the reference. Install U2 and the bypass capacitor C15, being careful to orient both parts correctly. C15 is a very large capacitor (2.2 μF) to suppress coupling between the RIT comparator and the Product Detector.

A. Now we measure the conversion gain of the RF Mixer. Connect the power supply and turn it on. Connect the counter to the VFO loops and use the VFO Tune and RIT pots to adjust the VFO frequency to 2.1 MHz. The VFO Tune serves as a coarse adjustment and the RIT as a fine adjustment. It is important to leave the counter connected for the rest of this lab. The counter has 30 pF of capacitance, and the cable adds more. Removing the counter would pull the VFO frequency up enough to push a signal right out of the pass band of the IF Filter. Make sure that the RF Gain pot R2 is fully clockwise. This gives maximum gain. Connect the function generator to the Antenna jack J1. The amplitude setting should be for a 50-mVpp sine wave. Look over your work in Problem 14 to find the center frequency for your IF Filter. Use this frequency for your IF. Now calculate the correct frequency f_{rf} for the function generator, and set the function generator to this frequency. Be careful to get the frequency right, or else you will not likely see a response. Connect the 10:1 scope probe to pin 1 of U2 to see the IF. Because the IF is not at the same frequency as the function generator, you cannot use the sync output from the function generator to trigger the scope. You should set up the scope for internal triggering. Adjust the tuning capacitors for the RF Filter (C1 and C2) for maximum IF output. You should vary the frequency slightly to make sure that you are at the frequency of minimum loss for the IF Filter. Now find the conversion gain in dB for the RF Mixer, assuming that the input impedance of U2 is 1.5 kΩ.

B. For the remaining scope measurements, you will probably find it convenient to switch in the low-pass filter if one is available to reduce the noise on the screen. This will reduce the scope signal level, and you will need to take this reduction into account. We can control the signal level with the RF Gain pot R2. To see the range of attenuation that it offers, rotate it fully counterclockwise. How much attenuation in dB is provided by the pot? You will probably want to increase the function generator amplitude setting to its maximum value to make the IF output visible at all. The IF will be small, and it can be challenging to get a good trace on the scope. The mixer bias gives this signal a relatively large DC offset, and you will need to switch the scope coupling to AC to get rid of the DC component. Otherwise the trace sails off the scope screen. You may also see multiple traces, indicating that the triggering is not consistent. If you have trouble with the triggering, you may be able to eliminate the traces you do not want by adjusting the triggering level. After you make the attenuation measurement, you should return the function generator to its original amplitude setting, 50 mVpp. Restore the RF Gain pot to the clockwise position. Make sure that you do not forget to do this, or the remaining measurements will be off.

C. Next we look for spurious responses. There are many spurs, but for now we will measure two of the largest ones. Calculate the image frequency f_i for the VFO. Set the function generator to f_i and look for the IF response. You will need to set the function-generator amplitude to its maximum output value to see it. By how many dB is the image response suppressed compared to a signal at the RF frequency?

D. The other spur that we will look for is $f_{5\downarrow}$, from the fifth harmonic of the LO. This is tougher to find than the image. This spur is close to the normal receive frequency, and for this reason, the RF Filter does not reject it well. Calculate $f_{5\downarrow}$ and $f_{5\uparrow}$. Set the function generator to $f_{5\downarrow}$. This time, you need to be more careful with the function-generator level, because the output saturates if you set the amplitude too large. You might try an amplitude setting of 1 Vpp. It is important that the VFO frequency be precisely 2,100 kHz during this measurement, because we are mixing with the fifth harmonic. A 100-Hz error in the VFO frequency will cause a 500-Hz shift in f_{5-}, and this will push us clear out of the pass band of the IF Filter. It is easy to get this large a shift if something comes near the air-variable capacitor C50 or the big inductor L9. You should vary the function-generator frequency slightly to make sure that you are at the peak of the response. By how many dB is the $f_{5\downarrow}$ spur suppressed compared to a signal at the RF frequency?

E. Can you find the error in Figure 4 of the data sheet for the SA602AN in Appendix D?

PROBLEM 29 – PRODUCT DETECTOR

The Product Detector is the second mixer in the receiver (Figure 12.12). The input to the Product Detector is the IF signal, after it has passed through the IF Filter. The local oscillator for the Product Detector is called the Beat-Frequency Oscillator, or BFO. The output is at the difference frequency, which is an audio frequency. We use an audio

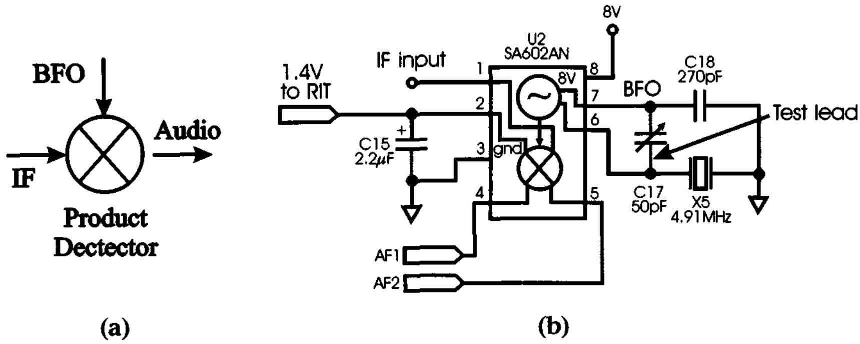

Figure 12.12. (a) Product Detector inputs and outputs, and (b) the schematic for the NorCal 40A Product Dectector.

frequency of 620 Hz. This matches the resonant frequency of the tuned speakers that you made for Problem 17.

The BFO uses the internal oscillator circuit in the SA602AN (Figure 12.12b). It is a Clapp oscillator like the VFO, except that the resonator is a crystal instead of an LC circuit, and the transistor is a BJT rather than a JFET. The divider circuit consists of C17 and C18. C17 is a variable capacitor, and this allows the BFO to be set precisely. We will put the BFO 620 Hz above f_{if}.

A. Install C17, C18, and X5. Plug in the power supply and turn it on. Use short test leads to probe pin 6 to see if there is an oscillation. You may need to adjust C17 to see it, because the circuit often does not oscillate when C17 is in the low-capacitance part of its range. Then move the test leads from the scope to a counter. Adjust the capacitor C17 for the minimum oscillation frequency, and record this frequency.

B. For comparison, calculate the minimum oscillation frequency that we should be able to reach with the BFO, assuming that the range of the variable capacitor C17 is 7 pF to 70 pF. For the crystal X5 use the inductance L and capacitance C that you measured in Problem 14. Set the oscillation frequency 620 Hz above the center frequency of your IF Filter by adjusting C17. If you cannot reach this frequency, get as close as you can.

C. Measure the temperature coefficient for the BFO in the same way that you did for the VFO. Try to make sure that the BFO components and the thermometer both get plenty of hot air, so that their temperatures will be similar. The frequency shift should be small, and it is a good idea to measure the frequency to the nearest hertz.

Next we look for the output audio signal. Install a 3-kΩ resistor in the C19 holes. Leave room to hook probes on each end. This resistor will act as the load for the Product Detector. Set the function generator to a 7,010-kHz sine wave, with an amplitude of 50 mVpp. Put a 10:1 scope probe at one end of the resistor. You will need to set the scope channel for AC coupling, because the SA602AN has a large DC voltage on its outputs. Adjust the

VFO Tune pot and the RIT pot until you see an audio signal on the scope. You will need to slow down the trace to see this. The waveform may be difficult to interpret because you are seeing the raw mixer output with all the frequency components.

Next we set the RF Gain pot for an attenuation of 20 dB, so that we do not overload the Product Detector. Start with the RF Gain pot set to minimum attenuation (fully clockwise). Move the scope probe to pin 1 of the Product Detector, and readjust the trace speed so that you can look at the IF. Tune the VFO for maximum scope voltage. Then adjust R2 to reduce the voltage by a factor of ten.

We will use the multimeter for AC voltage measurements for the rest of this lab. The advantage of using the multimeter is that both inputs of the multimeter are isolated from ground, and you do not short out the output the way a scope ground lead would. In addition, a multimeter usually only measures the voltage for frequencies up to about 100 kHz. Thus the sum frequency at 10 MHz and the higher-order products are ignored. Make sure that the multimeter leads do not get near the VFO inductor or air-variable capacitor, or they will pull the VFO frequency.

D. Measure the gain of the receiver from the Antenna jack J1 through the Product Detector. The input is the available power P_+ from the function generator, and the output is the power P delivered to the 3-kΩ load. You will need to add 20 dB to this number as an allowance for the loss in the RF Gain pot. You should note that the multimeter measures rms voltage, not peak-to-peak voltage.

E. Next we look for the $f_{3\downarrow}$ spur for the RF Mixer. This is in the AM band, and it is the one that causes the most trouble in this receiver. Calculate the frequency of the $f_{3\downarrow}$ spur for $f_{rf} = 7{,}010$ kHz. Set the function generator to $f_{3\downarrow}$, and use the maximum voltage setting. There is no danger of saturating the output this time, and you need all the voltage you can muster to get through the RF Filter. Set the RF Gain pot fully clockwise for minimum attenuation. The voltage will be small. You should vary the function-generator frequency slightly to make sure that you are at the peak of the response. This requires patience, because a multimeter is slow. By how many dB is the $f_{3\downarrow}$ spur suppressed compared to a signal at the RF frequency?

F. Finally, we investigate the spur at f_{if}. In principle, this spur should not be there, but in practice there is some leakage through the RF Mixer. Set the function generator to f_{if}. Use as big a voltage as you can without saturating the output. The bigger the output, the more it will stand out from the noise, and the easier it will be to measure it accurately. You can check for saturation by making sure that the output is proportional to the input. Try cutting the input voltage in half, and verify that the output also drops by a factor of two. If it does not drop this much, the output is saturated, and you should start over at the lower input voltage. By how many dB is the f_{if} spur suppressed compared to a signal at the RF frequency? One thing that makes this spur different from the others is that it does not vary when you tune the radio, because it does not depend on the VFO frequency. You can go right from the bottom of the tuning range to the top without affecting the spurious response. Remove the 3-kΩ resistor that you installed as the load for the Product Detector.

PROBLEM 30 – TRANSMIT MIXER

The Transmit Mixer takes the signal from the VFO and mixes it with its internal oscillator, the Transmit Oscillator (Figure 12.13). Its output is a low-level signal that is amplified by the transmitter–amplifier chain. In the receiver mixers, we take differential outputs, but in the Transmit Mixer, we only use one of the outputs. If we do this, we cut the output voltage in half and the output resistance in half. The available power is proportional to the voltage squared and inversely proportional to the resistance. Consequently, our available power drops by a factor of two.

A. To start, you should install all of the components in Figure 12.13 except for C31. This leaves the VFO disconnected initially, which makes it easier to measure the Transmit-Oscillator frequency. You will also need to insert the shorting plug to turn on the transmitter. You should adjust C34 to set the Transmit-Oscillator frequency to the center frequency of your IF Filter. This aligns the transmitter and receiver frequencies and sets the Transmit Oscillator 620 Hz below the BFO. If you study the resonators for the two oscillators, you will see that the Transmit Oscillator has an additional inductor, L5, which is an 18-μH inductor. Calculate how much you would expect this inductor to lower the oscillation frequency.

Connect the Antenna jack J1 to the scope with a 50-Ω load. Now install the capacitor C31 (5 pF). This capacitor couples the TX VFO output to the Transmit Mixer. Adjust the filter capacitor C39 to peak the transmitter output. Now set the Drive Adjust pot R13 for an output of 30 Vpp. Connect the output to the counter, again with the 50-Ω load in parallel. Adjust the output frequency to 7,010 kHz with the VFO Tune pot. The RIT pot should have no effect on this frequency. You should repeat the scope adjustments of C39 and R13 as a final transmitter adjustment.

B. The coupling capacitor C31 has two additional functions. First, the VFO and the Transmit Mixer input have different DC voltages, and the capacitor ensures that they stay that way. Also, the TX VFO voltage is too large for the input of the Transmit Mixer, and the capacitor acts as a potential divider to reduce it. Calculate the voltage attenuation factor for the potential divider. Measure the TX VFO voltage. Use the TX VFO voltage and the voltage attenuation factor to calculate the input power to

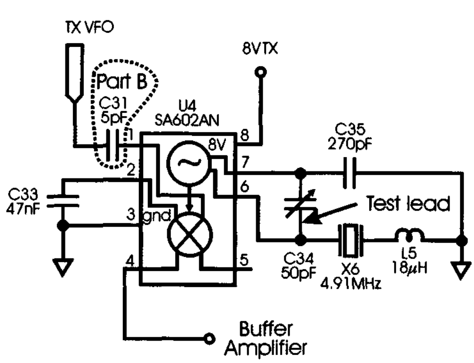

Figure 12.13. The Transmit Mixer and the Transmit Oscillator. C31 should not be installed until after the frequency is set in Part A.

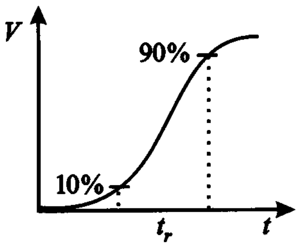

Figure 12.14. Measuring the rise time t_r of a pulse.

the Transmit Mixer. Calculate the gain in dB of the entire transmitter chain, starting from the Transmit-Mixer input and ending with the 30-Vpp output.

Now we look at the keying waveform. Replace the shorting plug with a keying relay cable, and drive it with a function generator set for a 10-Hz, 5-Vpp square wave. These pulse lengths are similar to those used in actual communications. You should use a sync cable to trigger the scope. Adjust the scope until you see the keying pulses on the screen. It is important that keyed transmitters not start or stop suddenly, because this causes interference. Operators on nearby frequencies hear annoying clicks. It is traditional to specify the rise and fall times of the keying envelope, which is the outline of the scope pattern. Previously we have characterized rise and fall times in terms of t_2, the time that it takes the trace to come a factor of two closer to its final value. The rise and fall waveforms for a transmitter are not exponentials, however, and so we will use a different definition. It is traditional to define the rise time of a pulse as the time it takes to go from 10% to 90% of the final value (Figure 12.14). The fall time is defined in a similar way. In the NorCal 40A, the rise time is determined by the Transmit-Oscillator build-up. The fall time is controlled by the Driver capacitor C56. In addition, the rise and fall times are influenced by the input–output characteristic of the Power Amplifier. It is not easy to calculate these times, and we will measure them instead.

C. Measure the rise and fall times of the keying envelope. It takes some thought to display the beginning and ending parts of the keying waveform with a time scale that allows accurate measurements. There is a substantial keying delay, and you may find that the leading edge of the pulse dances around the screen. Here are some things that you might consider to help make your measurements. You can use the trigger slope switch to determine whether the scope triggers at the beginning or the end of the pulse. You can also use the trigger level control to set where the sweep starts on the waveform. If you see several traces on the screen at the same time, try the holdoff control to control the delay between sweeps. This often helps in getting rid of traces that you do not want. When you finish the measurements, leave the holdoff control at the normal minimum position. Otherwise the delay reduces the number of sweeps and dims the trace.

D. The Transmit Mixer, in addition to producing the transmitter signal, has spurious frequency components that are of the form

$$f_{mn} = m f_{vfo} + n f_{to}, \tag{12.46}$$

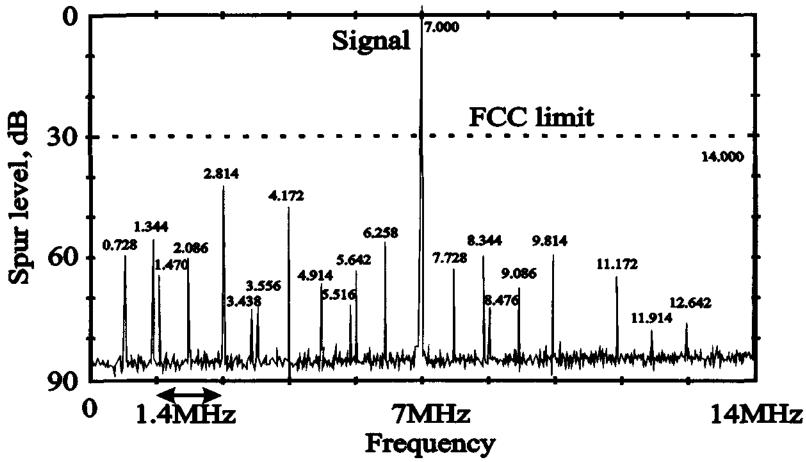

Figure 12.15. Frequency components for the NorCal 40A. The vertical axis shows the power on a dB scale, at 10 dB per division, and the horizontal axis shows the frequency, with a scale of 1.4 MHz per division. The different frequency components appear as lines on the plot, and the frequency is noted at each component.

where f_{to} is the frequency of the transmit oscillator, and m and n are integers. Figure 12.15 is a spectrum-analyzer plot that shows the power levels of the different frequency components from the NorCal 40A. Each of the components that is larger than 80 dB below the carrier has the frequency listed beside it. To make the plot, the spectrum analyzer measured the power at 14-kHz intervals from 0 to 14 MHz. Therefore, the frequencies are not precise, and you should allow for errors of up to 14 kHz. Find n and m for each of the lines. You should use the lowest values of n and m that give a frequency within 14 kHz of the line.