Chapter 7 : FREQUENCY CONVERSION

Mixers convert the frequency band of signals. The audio signal is first converted from *base-band* up to 16 MHz in TRC-10, and then to amateur band, between 28 MHz and 29.7 MHz. This is referred to as *up conversion*.

The received signal, on the other hand, is *down converted* to 16 MHz, where it is filtered and detected.

7.1. Amplitude modulators

Amplitude modulation is also a frequency conversion operation. Modulated signal is obtained right after the microphone amplifier in TRC-10. A diode and a tuned circuit perform amplitude modulation. Consider the circuit given in Figure 7.1.

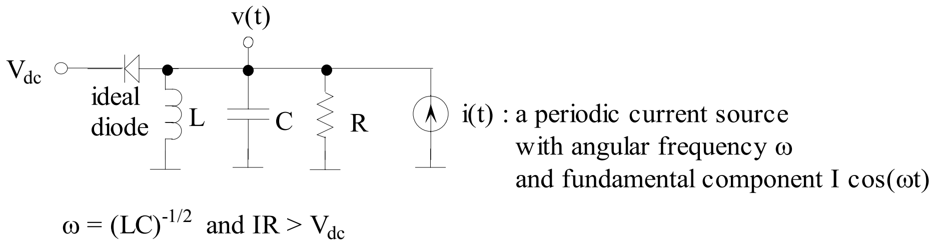


Figure 7.1 A circuit for amplitude modulation

A parallel tuned circuit is driven by a periodic current source and it is connected to a d.c. voltage source through a diode. In this circuit the voltage across the tuned circuit is always

$v(t) = V_{dc} \cos(\omega t)$

as long as $IR > V_{dc}$ and the circuit is tuned to ω.

Principle of operation is straightforward. We enforce $IR > V_{dc}$, so that if there were not any diode, the voltage peak across the tuned circuit would exceed V_{dc}. Whenever the voltage across the tuned circuit exceeds V_{dc}, diode conducts. Since the resistance of the d.c. voltage source is zero the peak of the voltage is *clamped* to V_{dc} (i.e. it cannot be more than V_{dc}).

On the other hand, the voltage that can appear across the tuned circuit can only be in sinusoidal form, if the Q of the circuit is reasonably high. The d.c. current component flows through the inductor and harmonics flow through the capacitor, generating insignificant voltages. Hence the tuned circuit voltage is forced to be a sinusoidal voltage with amplitude of V_{dc} and frequency of ω.

This problem belongs to a family of circuit theory problems, usually called as “nonlinear loading of narrowband circuits”. Exact solution of this problem, given above, can be analytically found by applying a circuit theory technique called *harmonic balance*. We leave the discussion of harmonic balance to advanced texts.

Now if we add a slowly varying (compared to ω) audio signal component to V_{dc} as in Figure 7.2, we obtain an amplitude modulator. OPAMP output impedance is very low and it behaves like a voltage source providing $V_{dc}[1+m(t)]$. Here, IR must be larger than the maximum value of $V_{dc}[1+m(t)]$, of course.

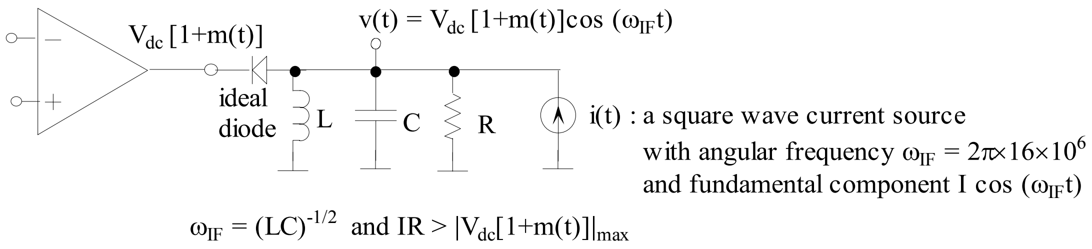


Figure 7.2 Amplitude modulator

7.2. Mixers in telecommunication circuits

7.2.1. Switch mixer

Converting the frequency of a signal requires multiplication operation. Consider a simple mixer comprising a switch in Figure 7.3.

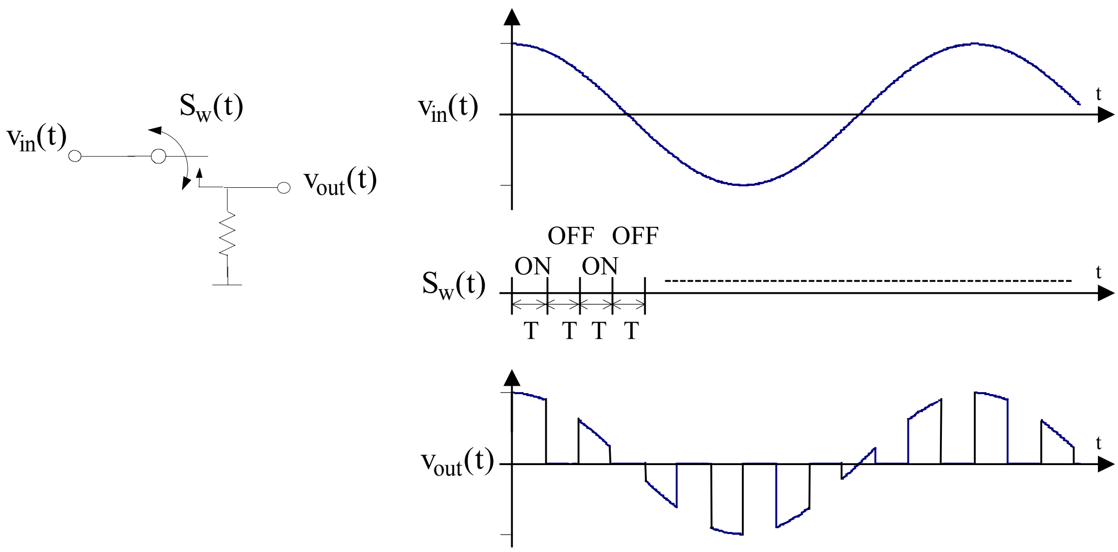


Figure 7.3 Switch mixer

Switch remains closed for T seconds and the opens and remains open for T seconds periodically. When the switch is closed, $v_{out}(t)$ is an exact replica of $v_{in}(t)$ in this circuit. When switch is open, $v_{out}(t)$ is zero.

This operation is equivalent to multiplying $v_{in}(t)$ by a square wave like the one in Section 1.2. This square wave, however, has amplitude of 1 when the switch is closed, and it is zero when the switch is open. Such a square wave can be written as

$$s(t) = a_o + \sum_{n=1}^{\infty} b_n \sin(n\omega_s t)$$

where a_o is 0.5 and $b_n = (1/n\pi)[1-(-1)^n]$ for this square wave (p-p amplitude of this square wave is 1). The period of the square wave is 2T, hence its fundamental angular frequency ω_s is π/T. Assuming that the input signal is a narrow band signal and can be represented by a sinusoid

$v_{in}(t) = V\cos(\omega_i t)$,

output becomes

$v_{out}(t) = v_{in}(t) \times s(t)$

$$= \{V\cos(\omega_i t)\}\{0.5 + \sum_{n=1}^{\infty} b_n \sin(n\omega_s t)\}$$

$$= 0.5V\cos(\omega_i t) + \sum_{n=1}^{\infty} b_n V\sin(n\omega_s t)\cos(\omega_i t).$$

Output signal contains signal components at many frequencies now, as well as a component at the input frequency, 0.5 $V\cos(\omega_i t)$. For n=1, b_1 is $2/\pi$. The next coefficient b_2 is zero and b_3 is $2/3\pi$. The open form of output signal can be written as

$v_{out}(t) = 0.5\ V\cos(\omega_i t) + (2/\pi)\ (V)\ \sin(\omega_s t)\cos(\omega_i t) + (2/3\pi)\ (V)\ \sin(3\omega_s t)\cos(\omega_i t) + \ldots$

$= 0.5\ V\cos(\omega_i t) + (V/\pi)\sin[(\omega_s+\omega_i)t] + (V/\pi)\sin[(\omega_s-\omega_i)t] + (V/3\pi)\sin[(3\omega_s+\omega_i)t] + (V/3\pi)\sin[(3\omega_s-\omega_i)t + \ldots$

The frequency of the input signal $v_{in}(t)$ is now shifted by all odd harmonics of the switch frequency. The signal component $(2/\pi)\ (V)\ \sin(\omega_s t)\cos(\omega_i t)$ in $v_{out}(t)$, for example, can be viewed as a DSBSC AM signal where ω_s is the carrier and ω_i is the modulating frequency (or the other way around). The spectra of $v_{in}(t)$ and $v_{out}(t)$ are given in Figure 7.4.

If we filter out the $(2/\pi)\ (V)\ \sin(\omega_s t)\cos(\omega_i t)$, or equivalently

$(V/\pi)\sin[(\omega_s+\omega_i)t] + (V/\pi)\sin[(\omega_s-\omega_i)t]$,

from $v_{out}(t)$, we obtain a *frequency converted* version of base band signal $v_{in}(t)$, converted to ω_s. All frequency components of $v_{in}(t)$ are shifted by ω_s in frequency.

The circuit in Figure 7.3 *mixes* $v_{in}(t)$ and s(t). This circuit can easily be implemented, using the techniques we discussed in Chapter 6.

Note that a component at input signal frequency is also present in $v_{out}(t)$. In some mixer configurations, a component at switch fundamental frequency is also present.

Such mixers are called *unbalanced* mixers. It is quite undesirable to have either one of these components at the output, because their presence complicates the filtering.

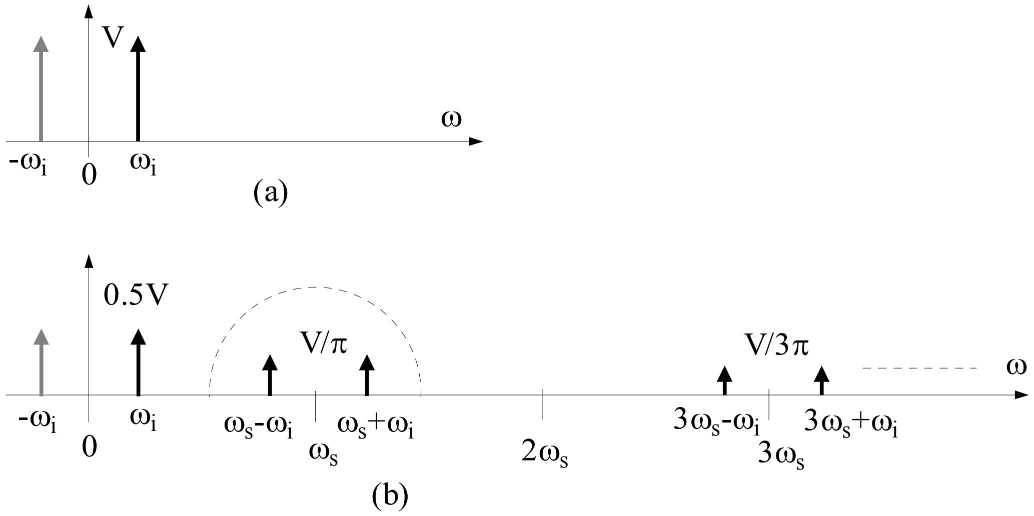


Figure 7.4 Frequency spectrum of (a) $v_{in}(t)$ and (b) $v_{out}(t)$

7.2.2. Double-balanced mixers

Mixers with configurations that provide an output free from any component at input frequencies (ω_i and ω_s in the case for the circuit in Section 7.2.1) are called *double-balanced* mixers. A double-balanced mixer circuit is shown in Figure 7.5.

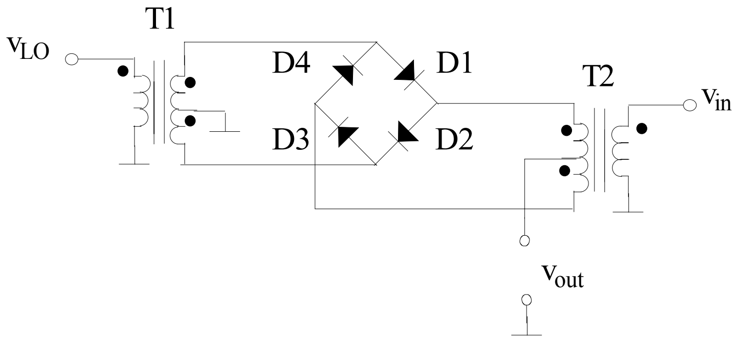


Figure 7.5 Double-balanced mixer

V_{LO} is the periodic signal (usually square wave) which provides the mixing frequency, like s(t) in Section 7.1.1. It is denoted as v_{LO}, because usually a *local oscillator* within the communication device generates this signal. v_{in} and v_{out} are the signal to be mixed and mixed output, respectively.

The dots on the transformer winding show the coupling polarity of primary and secondary voltages. Voltages on the dotted side of the windings have the same phase in each transformer.

This mixer works by switching the route of v_{in} to v_{out}. When the square wave of v_{LO} is in positive phase, D1 and D2 conduct. In this case the secondary current in T2 passes through the upper secondary winding of T2, D2 and lower secondary winding of T1 to ground. v_{out} is the voltage across the upper secondary winding of T2 in this case. Note that the dotted side of this winding is on the grounded side for this route.

When the square wave is in negative phase, D3 and D4 conduct. The secondary current follows the path provided by lower secondary winding of T2, D3 and upper secondary winding of T1 to ground. v_{in} couples to v_{out} across the lower secondary winding of T2. In this case un-dotted terminal of the winding is grounded. The phase of v_{out} in this route is exactly 180 out of phase with the one in the previous route.

The function of the circuit in Figure 7.5 can be modeled as multiplying v_{in} by a square wave of 1's and –1's, rather than 1's and zeros. The average value of this kind of *symmetric* square wave is zero and hence the d.c. term in its expansion does not exist.

7.3. Analog multipliers

A more sophisticated type of double-balanced mixing employs *analog multipliers*. Certain electronic circuit configurations provide means of multiplying two analog signals. *Gilbert cell* is a commonly used analog multiplier configuration. The input/output relations in a Gilbert cell circuit are modeled in Figure 7.6(a).

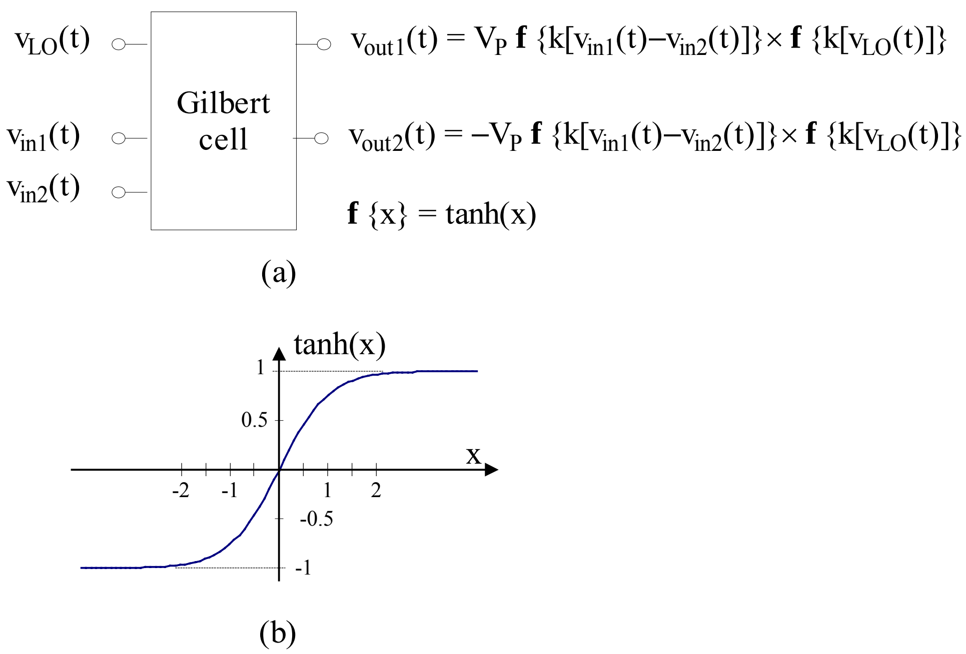


Figure 7.6 (a) Block diagram of *Gilbert cell* and (b) tanh(x)

The output of a Gilbert cell is a product of a function of the difference of two input signals and the same function of the local oscillator signal. The function is *hyperbolic tangent*, or *tanh(·)*, given as

$$\mathbf{tanh(x)} = \frac{e^{x} - e^{-x}}{e^{x} + e^{-x}}.$$

The variation of tanh(x) is depicted in Figure 7.6 (b). Gilbert cell is an electronic circuit and obviously it scales the input voltages in order to convert them into the argument of the tanh(·) function. The scaling factor is k and has a unit of $(\text{volt})^{-1}$.

Note that when x is small, tanh(x) is linear. Indeed for $|x| < 0.3$, $\tanh(x) \approx x$. Therefore for low amplitude levels of $v_{in1} - v_{in2}$ and v_{LO}, $v_{out1}(t) \approx V_p k^2 [v_{in1}(t) - v_{in2}(t)][v_{LO}(t)]$. A direct *analog multiplication* can be performed.

The block diagram of a typical application of using an analog multiplier as a double balanced mixer is given in Figure 7.7 (a).

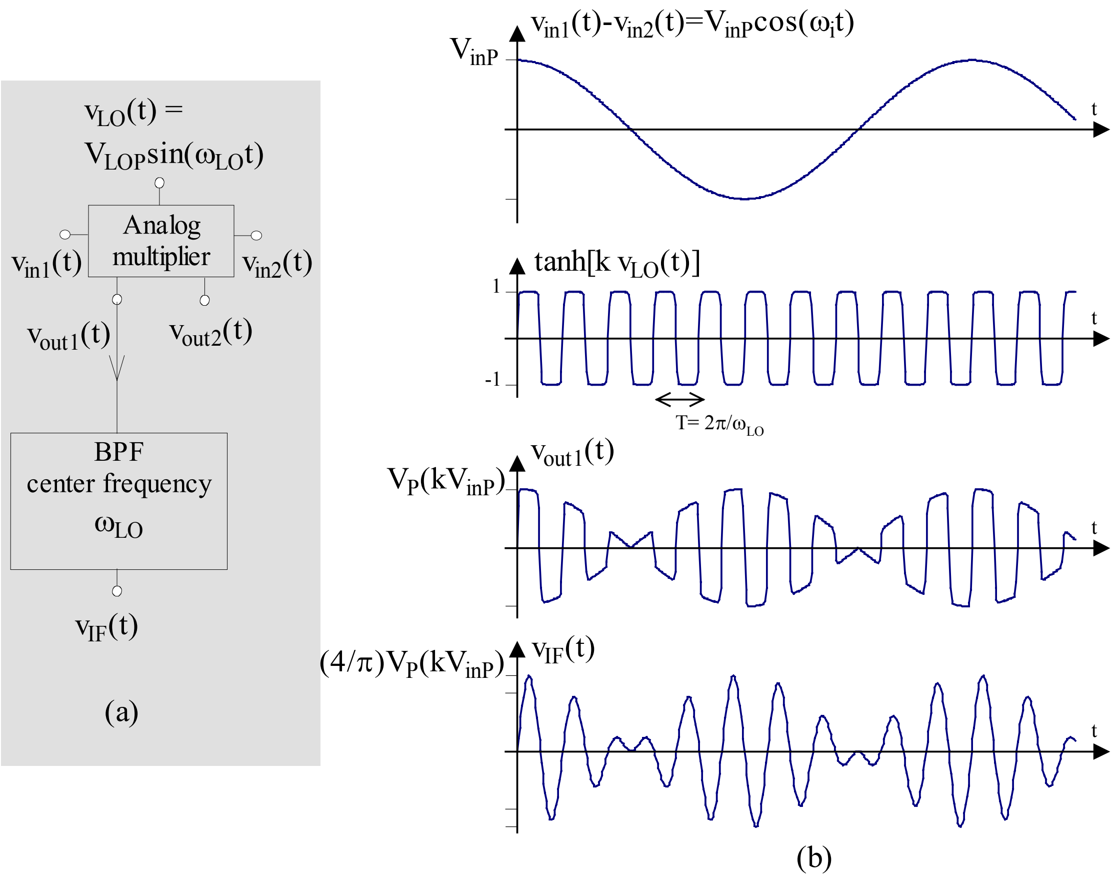


Figure 7.7 Analog multiplier and IF filter outputs with large v_{LO} amplitude (a) block diagram and (b) waveforms.

Base band signals (such as audio, for example) $v_{in1}(t)$ and $v_{in2}(t)$ and a sine wave from the local oscillator $v_{LO}(t)$ are applied as inputs to the analog multiplier. The output of the multiplier is filtered using a BPF centered at the local oscillator frequency. IF signal v_{IF} is obtained at the filter output.

The amplitude of $v_{in1}(t)$ and $v_{in2}(t)$ is normally kept low in order to avoid distortion. A low amplitude sine wave is assumed in the waveforms given in Figure 7.7 (b).

When analog multipliers are used as double-balanced mixers for frequency conversion, a high amplitude sinusoidal $v_{LO}(t)$ signal is employed. Then $\tanh[kv_{LO}(t)]$ is almost like a square wave with ±1 levels, as depicted in Figure 7.7 (b).

With such choices for $v_{in1}(t) - v_{in2}(t)$ and $v_{LO}(t)$, the output becomes

$$v_{out1}(t) \approx V_P k(v_{in1} - v_{in2})\{\sum_{n=1}^{\infty} b_n \sin(n\omega_s t)\} = V_P \sum_{n=1}^{\infty} b_n \sin(n\omega_s t)\{k(v_{in1} - v_{in2})\}.$$

The signal $v_{out1}(t)$ is given in Figure 7.7 (b). When this signal passes through the BPF, all harmonics of $\tanh[kv_{LO}(t)]$ are eliminated and we obtain the IF signal

$$v_{IF}(t) \approx b_1 V_P(kV_{inP})\cos(\omega_i t)\sin(\omega_s t).$$

The fundamental component amplitude in this square wave is $b_1=4/\pi$ (and is larger than 1), like the one in Chapter 1. Therefore the p-p amplitude of $v_{IF}(t)$ is b_1 times larger than the p-p amplitude of $v_{out1}(t)$.

Gilbert cell analog multipliers are difficult to implement at frequencies above few hundred MHz. SA602A is a very good integrated circuit implementation of Gilbert cell, which operates very well up to UHF. Such implementations are active circuits and they provide gain as well as analog multiplication.

7.3.1. Conversion gain

An important parameter in the evaluation of mixers is *conversion gain*. Conversion gain is defined as

$$G_c = P_o/P_i$$

where P_o is the total power delivered to a matched load at output and P_i is, again, the total available power at the input. This expression is similar to the gain of an amplifier, except here the input and output frequencies are different.

7.4. Oscillators

We need oscillators, i.e. periodic signal generators, for frequency conversion. Any telecommunication receiver contains at least one oscillator. We use one fixed frequency oscillator and one variable frequency oscillator (VFO) in TRC-10.

Fixed frequency oscillator is an IC module, which produces a square wave at 16 MHz.

The VFO in TRC-10 is implemented using an OPAMP and discrete elements. VFO provides a sinusoidal signal. Its frequency can be changed between 12 MHz and 13.7 MHz.

7.4.1. Oscillator concept

We discussed *feedback* in Section 3.4 and noted that we always use negative feedback for amplification. In oscillators, we need *positive feedback*.

Consider the simple circuit given in Figure 7.8 (a), where a current source is connected to a *very high Q* parallel tuned circuit. We assume that the current source $\bar{i}(t)$ contains many sinusoidal components at all frequencies and its amplitude is very small. Actually *noise* in electronics is such a signal. Only the current component in the vicinity of ω_o generates a voltage $v_1(t)$ across the tank circuit. The amplitude of $v_1(t)$ is very small also. We expect to observe $v_1(t)$ as

$$v_1(t) \approx V_1\cos(\omega_o t),$$

where ω_o is $(LC)^{-1/2}$ and V_1 is very small. If an amplifier is connected to this node, $v_1(t)$ is amplified as shown in Figure 7.8 (b).

Supply voltage levels limit the output voltage swing of amplifiers. When the amplified signal amplitude AV_1 approaches to supply voltage V^+, output waveform gets distorted. We obtain a clipped waveform. The amplifier is said to be *saturated*. If input voltage amplitude increases further, the output waveform approaches to a square wave.

At this stage, let us assume that the gain is not large enough to saturate the amplifier.

When a *feedback* path to the *positive input* is provided by means of a resistor R_2 as shown in Figure 7.8 (c), $v_1(t)$ is modified. Initially, an additive sample from output increases $v_1(t)$ to

$$v_1(t) \approx V_1\cos(\omega_o t) + [R_1/(R_1+R_2)]\times(AV_1)\cos(\omega_o t).$$

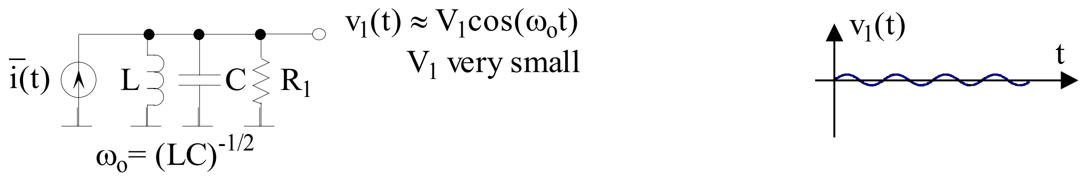
The additive component has an amplitude of $[R_1/(R_1+R_2)]\times(AV_1)$ which is much larger than the signal directly created by $\bar{i}(t)$. This component alone drives the amplifier deep into saturation, since $A\times[R_1/(R_1+R_2)]\times(AV_1) >> V^+$. We immediately have a square wave at the output as $v_2(t)$. The p-p amplitude of this wave is twice the saturation voltage, which is usually slightly less than V^+. This waveform is also shown in Figure 7.8 (c).

When we have a square wave at the output, the feedback signal is also a square wave. However, the tank circuit filters the fundamental component out of this square wave, producing a $v_1(t)$ as

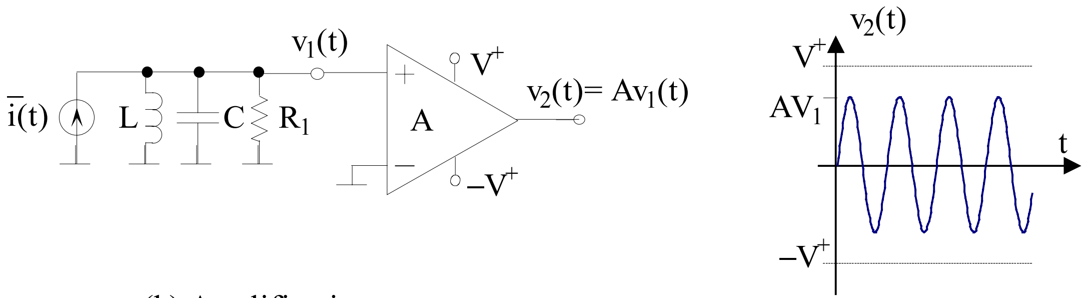
$$v_1(t) \approx V_1\cos(\omega_o t) + [R_1/(R_1+R_2)] \times b_1(V^+)\cos(\omega_o t).$$

Here, $b_1 = 4/\pi$ is the coefficient of the fundamental component in a square wave. Since the output of the amplifier cannot change any more, the circuit operation is stabilized with a square wave at its output. The tank circuit determines the frequency of this signal.

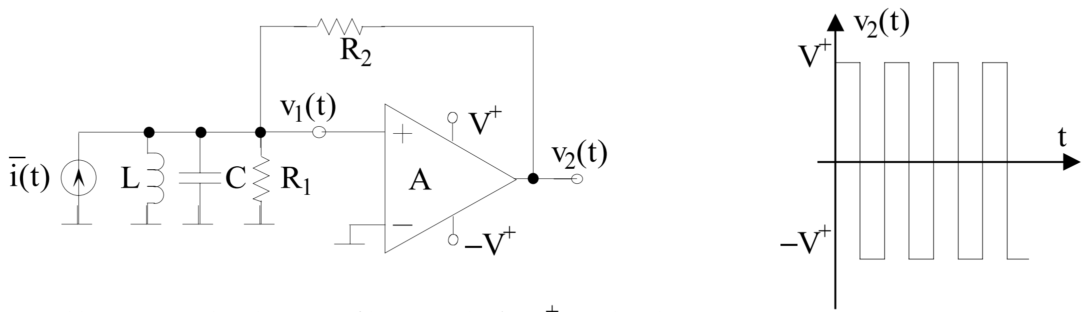
The input signal $v_1(t)$ is predominantly the feedback signal. If we remove the current source from the circuit, the output is still a square wave. Neither $v_2(t)$ nor $v_1(t)$ are affected, as shown in Figure 7.8 (d).



(a) Tank circuit driven by a wide band current source



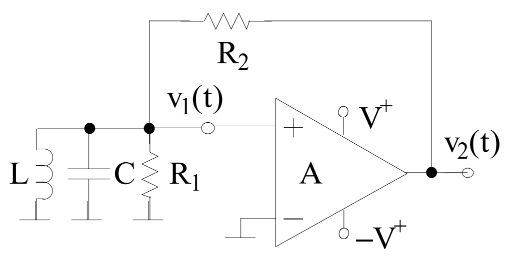
(b) Amplification

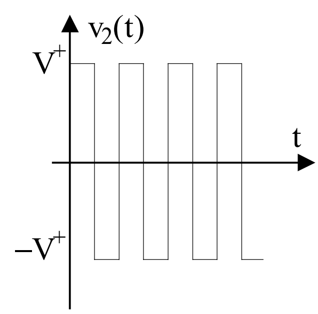


$v_1(t) \approx V_1\cos(\omega_o t) + [R1/(R1+R2)]\, b_1V^+\cos(\omega_o t)$

$[R1/(R1+R2)]\, b_1V^+ >> V_1$

(c) Positive feedback





$v_1(t) \approx [R1/(R1+R2)]\, b_1V^+\cos(\omega_o t)$

(d) current source removed

Figure 7.8 Oscillator concept

Practically we never include the current source to start the oscillation, because it is not necessary. There is *always* noise in electronic circuits. Also we do not need to remove the current source, because we cannot. It is *always* there.

It is possible to deduce from the above discussion that if

$[R_1/(R_1+R_2)]A=1$,

we have a sustained oscillation, once it starts. In this case the output waveform is sinusoidal and amplifier works in linear region all the time. This condition, i.e. the product of amplifier gain and the feedback ratio being unity is *oscillation criterion*.

In the circuit of Figure 7.8, we have the amplitude limiting mechanism of saturating amplifier. Since the fundamental component of the saturated output is b_1V^+, input amplitude is always $[R_1/(R_1+R_2)]\ b_1V^+$. As long as the gain is large enough to keep the amplifier in saturation with this input, oscillation is sustained. The larger gain of the amplifier helps oscillations to start easily. The feedback ratio $[R_1/(R_1+R_2)]$ and gain A must be such that

$$[R_1/(R_1+R_2)]A\times(V^+) > V^+ \qquad \Rightarrow \qquad [R_1/(R_1+R_2)]A > 1,$$

in order that oscillation starts up in this circuit.

There are many subtle subjects in the theory of oscillators, such as *frequency* and *amplitude stability*, *phase jitter*, etc. We leave these topics to advanced texts.

7.4.2. Frequency control

The frequency determining parameters are C and L in the oscillator circuit of Figure 7.8 (d). We must change the value of one of these components, if we want to vary the frequency. It is possible to use either a variable capacitor or a variable inductor.

One of the simplest ways of changing the capacitance is to use a semiconductor device called *varactor diode* (also called *varicap diode* and *tuning diode*). The symbol of the varactor diode is shown in Figure 7.9 (a). Varactor diodes are used with a reverse voltage bias on them. The capacitance across the cathode and anode depends on the level of reverse bias voltage. The variation of the diode capacitance with respect to reverse diode voltage is given in Figure 7.9 (b).

The circuit in Figure 7.9 (c) delineates the way a varactor diode is used in a circuit. The potentiometer R_{tune}, connected between two supply voltages, contrrol the reverse voltage bias on the diode. The capacitance C_o blocks the d.c. bias voltage on the diode. The large series resistor only serves to isolate the tuned circuit elements from R_{tune}, so that the Q of the resonant circuit remains high. The series combination of diode capacitance and C_o appears across the tank circuit. The total capacitance of the tank circuit becomes

$$C + C_DC_o/(C_D+C_o).$$

Adjusting the potentiometer can now vary the resonance frequency of the tank circuit.

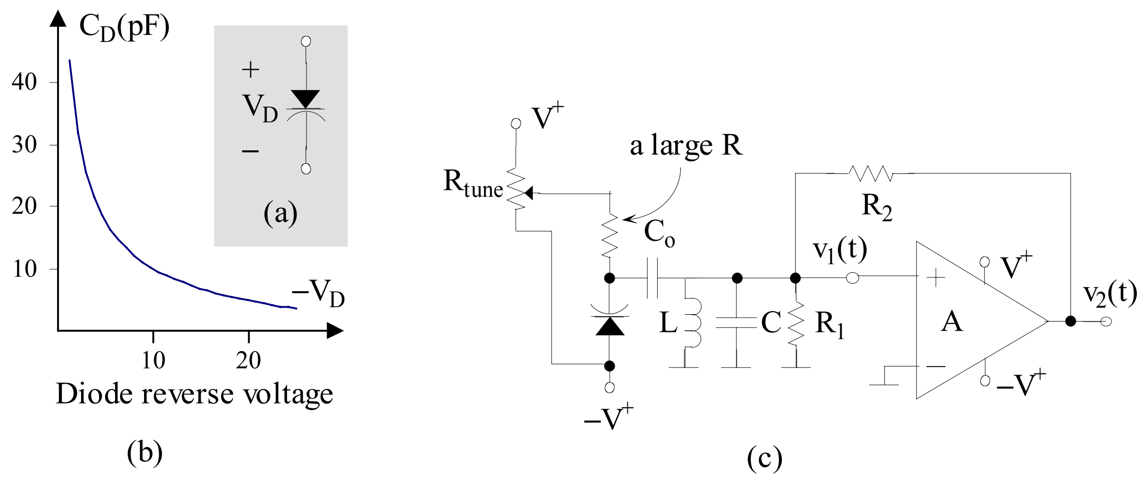


Figure 7.9 Varactor diode (a) symbol, (b) capacitance variation and (c) use in the circuit

7.4.3. Phase considerations in oscillators

We assumed that the OPAMP is ideal in Section 7.4.1 when discussing the concept of oscillation. There, we implicitly assumed that the amplifier provides a gain without any effect on the phase of the signal. However, when OPAMPs are used for RF amplification, the open loop transfer function must be considered in calculating the gain, as discussed in detail in Section 4.5.1. The LM7171 based non-inverting TX amplifier analyzed in Section 4.5.1 turned out to have a voltage gain of $5.1\ \angle\text{-}31°$ between the positive input pin 3 and output. If the OPAMP had indefinitely large open loop gain as in the case of an ideal OPAMP, the overall gain of the amplifier would be $5.7\angle 0°$ instead of $5.1\ \angle\text{-}31°$. This phase shift does not affect the performance of the TX amplifier functionality.

A similar phase shift, however is critical if we use the same amplifier in an oscillator. The concept of oscillation relies on the consideration that a constructive addition of the *fed back signal* and the available signal takes place at the positive input terminal, as depicted in Figure 7.8 (c). The sum of addition is maximum when the phases of both signals match. Hence the oscillator configuration in Figure 7.8, seeks the frequency where the phases of the output and that of the input are the same. In other words, there must be no phase difference between the output and positive input at the oscillation frequency.

We use a series capacitor, C140, in the positive feed back loop of the oscillator amplifier in TRC-10, instead of a resistor only. The value of this capacitor must be chosen such that the phase shift provided by this series combination compensates for the shift in the amplifier in the vicinity of the required oscillation frequency.

7.5. Bibliography

Many texts are available on mixers. However the treatment of the topic at introductory electronics textbooks is usually very weak. In *The Electronics of Radio* mixing and SA602A are discussed at an excellent level and detail.

P. R. Gray, P.J. Hurst, S. H. Lewis and R.G. Meyer (Wiley 2001) discuss Gilbert cell in *Analysis and Design of Analog Integrated Circuits, 4th Edition*, in Chapter 10.

Oscillators are extensively discussed in literature. A very good account of OPAMP oscillators is given in D. L. Terrell's *OPAMPS: Design, Application, & Troubleshooting*, Prentice Hall (1992).

7.6. Laboratory Exercises

16 MHz oscillator

1. Study the data sheet of 16 MHz oscillator module. Note that the power supply circuit of this module is connected to +15 V through TX/RX switch. We want this oscillator to operate only when we are transmitting. Therefore we switch its power off during reception. Install and solder D5 and C66. Install and solder the C66 end of R55. Using a piece of wire, make a soldered jumper connection from the free end of R55 to +15 V supply. Switch the power on. Measure and record the d.c. voltage across D5. It must be about 5 V. Switch the power off.

2. Install the IC8 module taking care of correct pin positions, and solder. Switch the power on.

3. Connect the oscilloscope probe to output pin, and observe the square wave at 16 MHz generated by the module. Measure and record the p-p amplitude (must be about 4-5 V average amplitude). Measure and record the frequency using a frequency counter.

Amplitude modulator

4. Amplitude modulator of TRC-10 is given in Figure 7.10.

   The operation concept of this circuit is given in Section 7.1. There are few deviations from the circuit of Figure 7.2, in this circuit:

   a. The16 MHz oscillator and a 1.5 K resistor, which couples the output voltage of oscillator to the circuit, replace the current source.
   b. Parallel capacitance is about 100 pF and is a combination of C57, C58, C60 and series C62. We need a voltage division between C60 and C62 to have a proper signal amplitude at IC7 input.
   c. D1 is a real diode rather than an ideal one.
   d. There is a R52-C59 section between the audio signal and the modulator.

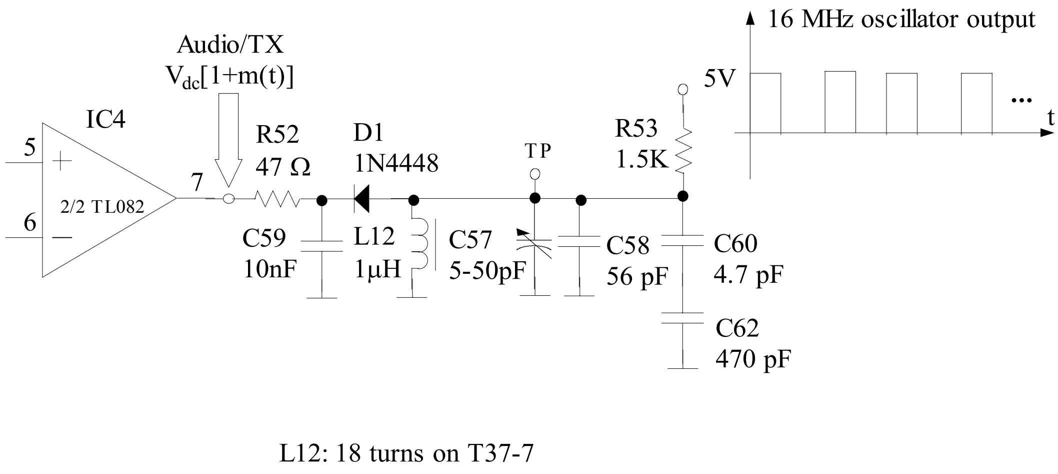


Figure 7.10 Amplitude modulator of TRC-10

Calculate the Norton equivalent of the circuit formed by 16 MHz oscillator output and R53. Re-draw the tuned circuit after replacing this part of the circuit by the Norton equivalent and replacing C57, C58, C60 and C62 by their equivalent capacitance. What is the resonant frequency? What is the Q of the circuit? What is the maximum amplitude of the square wave current that can flow into the tank circuit? What is the maximum amplitude of the voltage that can appear across the tuned circuit if there were not any diode?

5. A good voltage source output is required at the analog side, for proper operation of this modulator. This requirement also includes a very low output impedance. (recall that the output impedance of an ideal voltage source is zero!). Analog voltage is presented at the output of a TL082 OPAMP. TL082 has very low output impedance at low frequencies. What is it at 1 KHz (approximately)? This circuit has current components at 16 MHz and at its overtones. We do not have any data at these frequencies on TL082, but we know that the output impedance can be considerably high (How?). Hence the output impedance of TL082 by means of R52 and C59 to such a form that it presents reasonably low impedance at all frequencies.

   Calculate the Thevenin equivalent circuits Both the voltage and the impedance seen at the diode side at 1 KHz, 16 MHz and 48 MHz.

6. Diode D1 is a 1N4448 signal diode. It is not ideal. We know that if a diode is not ideal, we may suffer certain problems when the voltage across it is small. We also discussed that a possible solution is to bias it.

   The rigorous analysis of this modulator circuit (which is not difficult but is definitely beyond the scope of this book) shows that there is always some bias current which improves the diode performance in this circuit. Hence consider D1 ideal, as long as the modulation index is not close to 100%.

7. Install and solder R52, R53, C57, C58, C59, C60, C62 and D1. Wind 18 careful and loving turns on a T37-7. Install and solder L12 after carefully trimming and soldering the leads.

8. Check the connections both visually and using a multimeter. Make sure that there are no short circuits and all your solders are well. Making good solder joints are particularly important in RF circuits.

9. Compensate both of your probes.

10. Connect the probe of channel 1 at the output of the microphone amplifier (an appropriate place) and the probe of channel 2 across L12. Switch on the power. Set potentiometer R31 to somewhere at mid-range. Adjust R28 such that approximately 1.2 V d.c. is observed on channel 1.

    Now tune the tank circuit to the frequency of the 16 MHz oscillator by adjusting C57. You must observe a good (no distortion) sine wave across the tuned circuit (on channel 2) when it is properly tuned. Measure and record the amplitude of the sine wave. Compare this amplitude with the d.c. voltage at the output of microphone amplifier.

    Switch off the power.

11. Connect the signal generator to the input of microphone amplifier as in Figure 3.28. Connect the probe of channel 1 across D3 and D4.

12. Switch on the power. Adjust the signal generator to a sine wave of 1 KHz frequency. Increase the amplitude sufficiently, such that a clipped sine wave of approximately 0.5-0.6 V peak amplitude appears across D3 and D4.

    Connect channel 1 probe back to the output of microphone amplifier output. Adjust the scope settings so that you can observe the clipped sine wave on channel 1. Adjust R31 so that you observe 0.6 V peak a.c. voltage superimposed on a 1.2 V d.c., on channel 1 at the output of amplifier.

    Keep the oscilloscope triggered by the signal on channel 1 and adjust the scope settings so that you can observe the signal on channel 2 connected across L12. It must be an approximately 4 V p-p amplitude AM signal at 16 MHZ. Re-adjust C57 for maximum swing, if necessary.

    Adjust R28 for 50% amplitude modulation. Record the peak and valley amplitude. (You may need to refresh your knowledge on AM given in Chapter 1). Switch the power off.

    Since the probe of channel 2 was connected across the tank circuit during tuning, the probe tip capacitance Cp (please refer back to Laboratory exercise 3 of Chapter 3) also appears in parallel with the tank circuit. Hence the tank circuit is tuned to a lower frequency than what is required when the probe is not connected. We shall therefore adjust C57 after installing TX mixer.

13. What is the expected frequency content of this signal when the modulating signal is a pure sine wave? How many sine waves are there in the modulated signal? What are they?

VFO

14. VFO of TRC-10 generates sine wave at frequency range of 12-13.7 MHz, approximately. VFO is an OPAMP oscillator circuit, which uses a LM7171 as the gain element. It has a basic oscillator part and a frequency variation sub-circuit. The *variable frequency* property is provided by two parallel varactors and their d.c. bias circuits. The basic oscillator circuit is depicted in Figure 7.11.

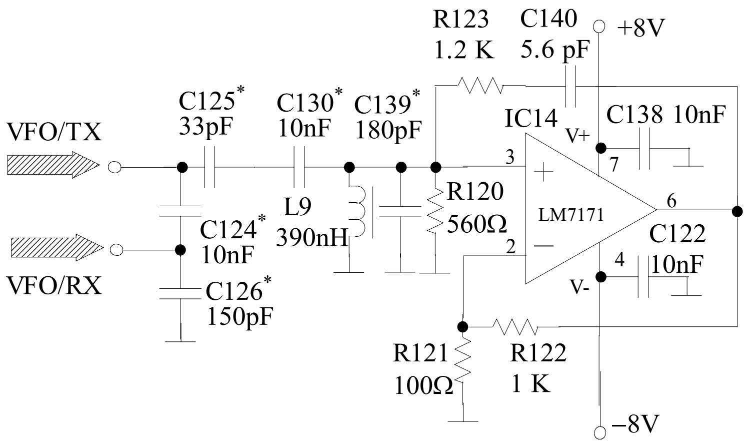


Figure 7.11 Basic oscillator circuit

The gain providing part is the OPAMP with resistive feedback in Figure 7.11. This part is shown in Figure 7.12, separated from the rest of the VFO circuit.

R121 and R122 are the resistors of the feedback block. C138 and C122 are supply by-pass capacitors.

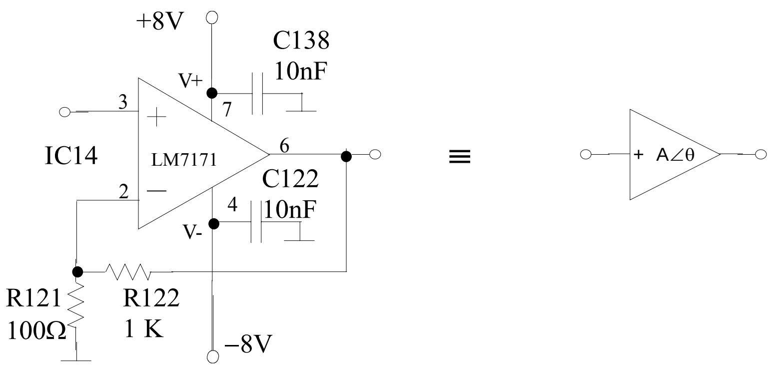


Figure 7.12 Gain block of VFO

15. Calculate the gain A and its phase θ at 16 MHz for this amplifier, making use of the data in the datasheet.

16. The parts of VFO basic circuit are shown in Figure 7.13.

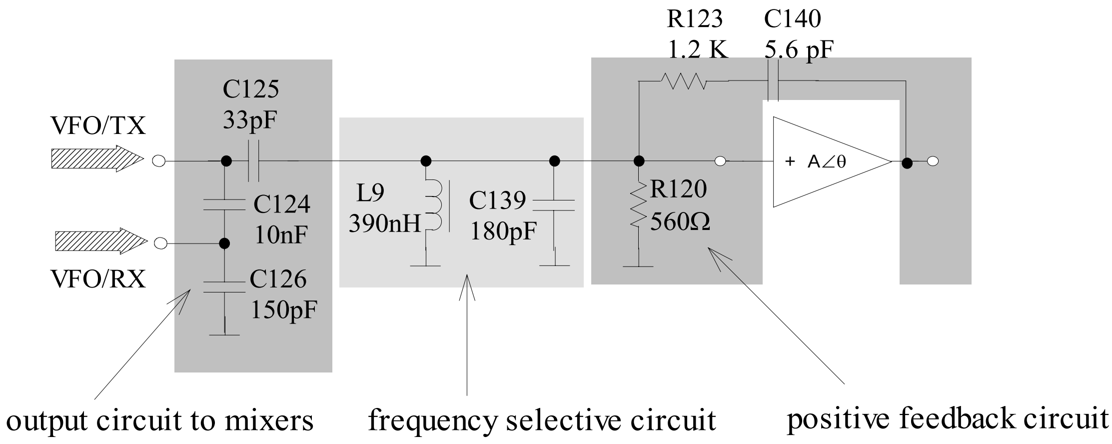


Figure 7.13 Blocks in oscillator circuit

The output circuit must both attenuate the oscillator output to that level required by SA602A. Also it must decouple the circuits d.c. wise.

The output of the OPAMP is a square wave. It is fed back by R120, R123 and C140 filtered into a sine wave across the tuned circuit. Series C140 compensates for the phase θ calculated in 15. What must be the exact value of C140 so that the voltages at pins 2 and 3 of the OPAMP have the same phase?

Calculate the positive feedback ratio.

17. The frequency selective circuit has an inductor L9 and a parallel capacitor C139. However other capacitors in the feed back circuit and the output circuit also affect the resonance. The equivalent parallel capacitance can be calculated from Figure 7.14. C130 is a large d.c blocking capacitor used in varactor circuit.

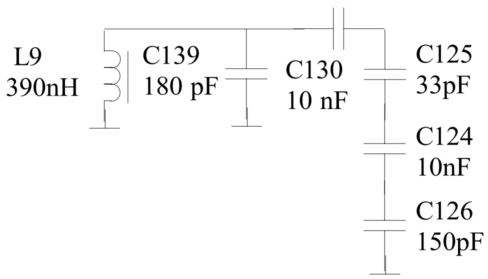


Figure 7.14 All capacitors in parallel with L9

The equivalent parallel capacitance is approximately 250 pF. This capacitance sets the resonance frequency to about 16 MHz with 390 nH. Any other capacitance that appears in parallel with C139 pulls this frequency downwards. There is additional parallel capacitance due to PCB connections of this circuit.

18. Install and solder IC14, taking care of correct placement of the pins. Install and solder the supply by-pass capacitors C122 and C138. Install and solder R120, R121 and R122. Do not install R123 and C140 yet.

19. Switch the power on. Measure the supply voltages and make sure that they are all right. Measure the d.c. levels of the OPAMP output, and both inputs, which must be zero nominally (remember that the presence of input bias currents and offset voltage can set the input to few mV and output to few hundred mV d.c., as discussed in Chapter 3). Switch the power off.

20. We use a T37-7 core for L9. Calculate the number of turns necessary for L9. Wind the inductor, install and solder. Install and solder C124, C125, C126, C130, C139, C140 and R123. Check the joints.

21. Switch the power on. Connect the probe of the oscilloscope across the OPAMP output. The lead of R123 (or R122) connected to pin 6 may be a better place to hook up the probe. Observe the square wave signal. Measure and record its amplitude and its approximate period. Measure the frequency of this square wave.

    It is possible to observe the sine wave across the tank circuit with the oscilloscope. We can measure the sine wave across the VFO/RX and VFO/TX pigtails (i.e. across C126 and C124), safely. The capacitive loading created by the probe is insignificant when compared with the large capacitor C126. Measure and record the amplitude of these sine wave signals. Switch the power off.

22. Frequency variation for VFO is maintained by two parallel varactors in a single case, KV1360NT. The d.c. reverse bias voltage is adjusted appropriately to change the total capacitance of two diodes, which are connected in parallel with C139 through a large (d.c. blocking) capacitor C130. The varactor circuit of VFO is shown in Figure 7.15. The only function of C130 is to isolate the d.c. bias voltage of varactor diodes from the tank circuit. Calculate the reactance of C130 at 16 MHz. Can this reactance be ignored when compared with the reactance of varactor diode?

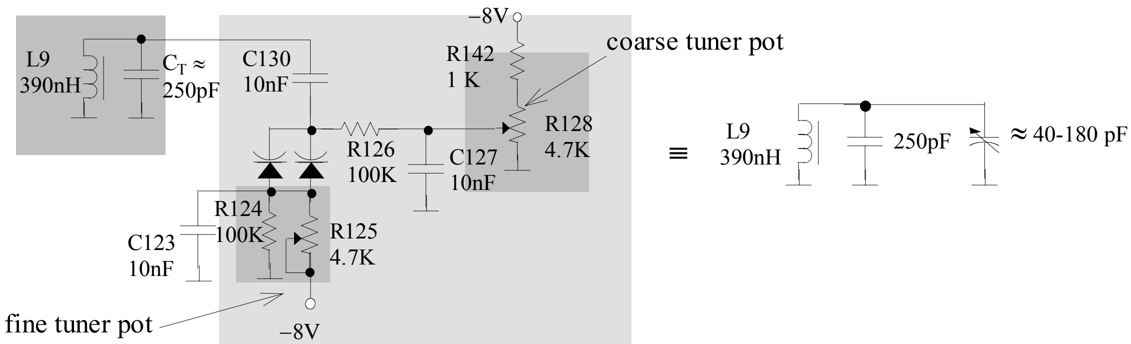


Figure 7.15 Tuner circuit

Calculate the d.c. potential range that can be produced at the cathode of the diodes by the *coarse tuner potentiometer*, and the potential range at the anode by *fine tuner potentiometer*. C123 and C127 are large capacitors to provide a stable d.c. bias voltage for the varactors.

The approximate variation of the capacitance of KV1360NT with respect to reverse bias voltage is given in the datasheet in the appendix. Determine the tuning range of total varactor capacitance in this circuit.

23. Install and solder C123, C127, R126, and R124. Install and solder the PCB connectors J24 and J25. Install and solder the varactor diodes D10, taking care of correct polarity. Check the connections.

24. R125 is a panel potentiometer for fine tuning. Short the middle pin and one of the side pins of R125. Cut two 15 cm pieces of wire and solder them to the respective pins of R125. Crimp the contacts for PCB jack to their other ends. Mount R125 on the panel, using a pair of pliers. Use anti-slip washer. Take care not to round the nut. Fit an *enumerated skirt* and a knob on the pot swindle. Fit the jack.

25. Install and solder R142. R128 is a panel potentiometer for coarse tuning. Cut three 15 cm pieces of wire and solder them to the respective pins of R128. Crimp the contacts for J24 PCB jack to their other ends. Mount R128 on the panel similar to R125. Again fit an enumerated skirt and a knob on the pot swindle. Fit the jack J24.

26. Hook up a frequency counter to the R122 leg connected to pin 6 of LM7171. Switch on the power. Find the tuning range of the oscillator. We want the tuning range to cover 12-13.7 MHz range. If this range is not covered, modify L9 for proper frequency adjustment.

27. Set fine adjustment pot to minimum. Set each step of the enumerated skirt of *coarse tuner* potentiometer to a pointer, and measure the corresponding frequency using the counter. Make a record of your measurements by filling the following table.

0	1	2	3	4	5	6	7	8	9

Mixers

28. We use three SA602A integrated circuits produced by Philips, as mixers in TRC-10. SA602A has a Gilbert cell made up of transistors. SA602A is a very popular and low-cost analog multiplier. It can be used in applications up to 500 MHz.

    There is also SA612A, which is almost exactly the same as SA612A, with slightly relaxed specifications. As far as TRC-10 is concerned, SA612A is a direct replacement for SA602A. The data sheet of SA602A is given in Appendix D.

    Examine the data sheet carefully. The following discussion is based on the information given in the data sheet.

    As in the case of an OPAMP, internal circuit configuration is not relevant here. All we need to know is how to use this IC in our circuits. The information about the power considerations and the input/output specifications of SA602A as a block is sufficient for our purposes. The block diagram and pin configuration of SA602A is given in Figure 7.16.

SA602A has an internal amplifier for oscillator. In other words, a local oscillator can also be directly implemented within this mixer. This amplifier input is pin 7 and oscillator output is pin 6. However we use external oscillators in TRC-10. The external oscillator signal is applied through pin 6.We do not use pin 7 at all, and this pin is *not connected* (NC) in our circuits.

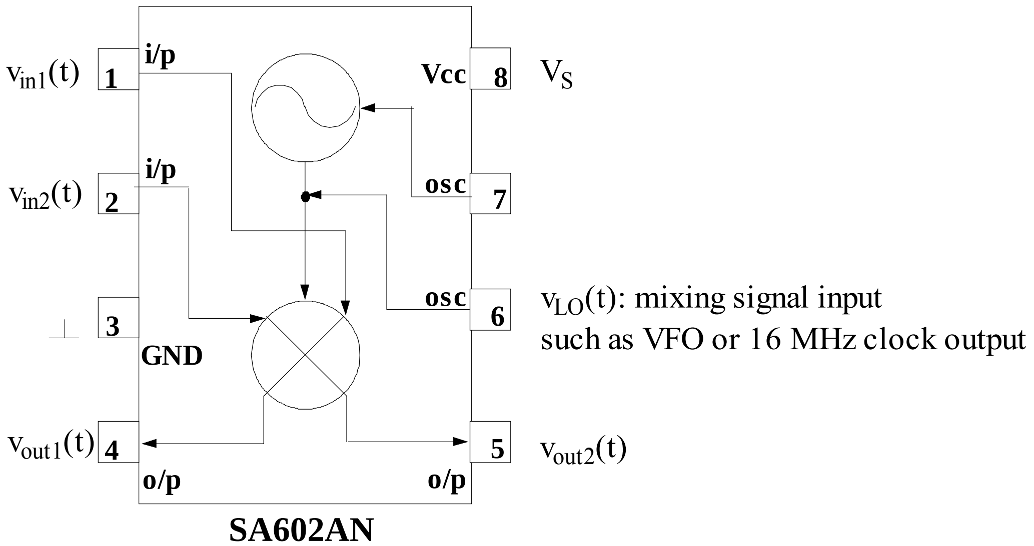


Figure 7.16 SA602A block diagram

SA602A operates from a single supply voltage and supply pin is pin 8. The ground pin is pin 3.

SA602A has two input pins. Either one can be used. Input pins have a bias voltage of about 1.4 V d.c. The external circuit must not disturb this voltage level. A d.c. blocking capacitor in series with the input pin is necessary.

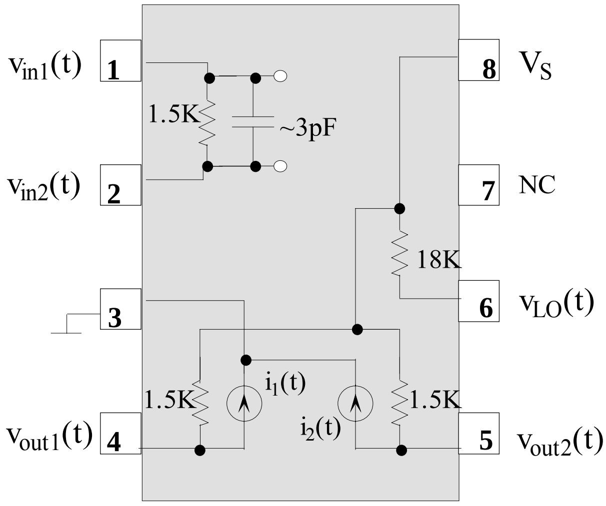
When SA602 is operated from a single input, the other input pin must be very effectively by-passed to ground, with capacitors.

We can also use both input pins of SA602A *differentially*. The voltage difference between the two pins is used in mixing, in that case.

When used as a mixer local oscillator must be at least 200 mV p-p, and must not exceed 300 mV. 200 mV p-p amplitude a square wave shaped internal $v_{LO}(t)$ signal, as we discussed in Section 7.3.

The two input terminals have impedance of approximately 1.5 K resistance in parallel with about 3 pF capacitance. The impedance of the local oscillator input terminal (between pin 6 and 3) is about 18 K.

An equivalent circuit of output terminals is given in Figure 7.17. The two outputs of SA602A are connected to the supply pin through 1.5 K resistors. The output signals are supplied by symmetrical *current sources* $i_{in1}(t)$ and $i_{in2}(t)$. The current sources have a d.c. component of 0.8 mA, as well as the signal components described in Section 7.3 and delineated in Figure 7.17. The signal component amplitude is also 0.8 mA. The scaling constant k is $(52\text{ mV})^{-1}$, approximately.



$i_1(t) \approx (0.8\text{ mA})\{1+\tanh[k(v_{in1}-v_{in2})]\times\tanh[kv_{LO}]\}$

$i_2(t) \approx (0.8\text{ mA})\{1-\tanh[k(v_{in1}-v_{in2})]\times\tanh[kv_{LO}]\}$

$k \approx 1/(2\times 26\text{ mV})$

Figure 7.17 Equivalent circuit of SA602A input and output

The circuit configuration and the d.c. component in the current sources yields a d.c. voltage level of $V_S - (0.8\text{ mA})\times(1.5\text{ K}) = V_S - 1.2$ V approximately. External circuits must not disturb this voltage.

The specifications of SA602A are applicable when the IC terminals are properly terminated. The external circuit must provide 1.5 K resistance to output pins.

UP-conversion to 29 MHz with SA602A

29. Install and solder IC7, the TX mixer, which converts 16 MHz signal to 28-29.7 MHz, taking care of correct pin alignment. Make sure that pin 8 corresponds to 8 V supply pad.

30. Switch the power on. Using a multimeter, measure and record the d.c. supply voltage at pin 8, d.c. voltages at two input pins and d.c. voltages at two output pins. Compare them with the expected values given in Exercise 28. If you cannot observe these voltages at respective pins, then there must be something wrong in your circuit.

31. Check the connections both visually and using a multimeter. Make sure that there are no short circuits and all your connections are properly soldered. Making good solder joints are particularly important in RF circuits.

32. Install and solder C61 and C63. Desolder the ground plane constructed resistors connected to C64, 51 Ω and 1.5 K. Install and solder the free lead of C64.

33. Calculate the voltage amplitude that appears across pin 1 of IC7, using the values of C60 and C62, and the measurement made in Exercise 12.

34. Connect the probe of channel 1 the output of the microphone amplifier. Switch on the power on. Adjust the signal generator to a sine wave of 1 KHz frequency. Adjust the amplitude such that a clipped sine wave of approximately 1.0V pp appears across D3 and D4.

    Connect the probe of channel 2 across the pin 6 of IC9 (rather on TP on R58). This is a comfortable point to connect probes. You must observe a few hundred mV amplitude AM signal at approximately 29 MHz. Re-adjust C57 for maximum amplitude.

    If necessary, re-tune C65 for maximum amplitude on channel 2 (*this should not be necessary if you have not played with C65 after it is tuned*).

    Re-adjust R28 for 50% amplitude modulation, if necessary. Record the peak and valley amplitude.

35. Connect the probe of channel 2 to the output of IC 10 (possibly at IC10 side of R61), instead of the output of IC9. Observe the AM signal.

    Both oscillators must be active and connected to respective mixers in order that the AM signal is generated. Make sure that the jumper connection for +15 V supply made in Exercise 1, is still connected.

    Adjust C69 for maximum amplitude, if necessary. This must be an AM signal of approximately 6 V peak without distortion. Adjust R28 again if necessary for maximum amplitude. Measure and record the peak amplitude at IC 10 output and the modulation index at this amplitude.

36. Remove the supply jumper to 16 MHz oscillator. Place the free end of R55 into the hole and solder. Now the 16 MHz oscillator can be active only when the S2 is thrown to TX.

RX mixer

37. All the input and output circuits of the RX mixer are already installed and used. Install and solder IC11 (another SA602A). Take care of correct placement. Install the supply bypass capacitor C96, and input bypass capacitor C93.

38. Repeat Exercise 30 for IC11.

39. We need a very weak (low power) signal at the antenna input to test the RX chain. The signal generator output does not provide such low level signals. We must use an attenuator at the antenna input. We use the circuit given in Figure 7.18.

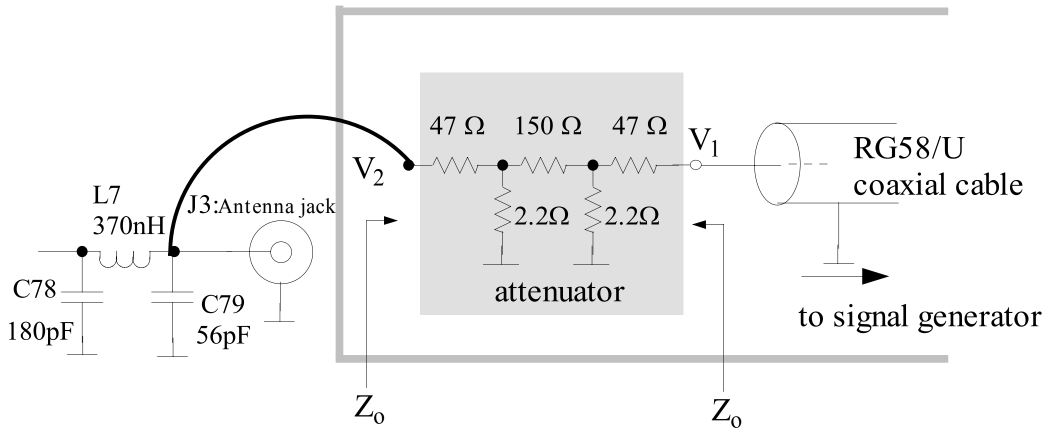


Figure 7.18 Signal generator connection for RX sensitivity measurement

Show that the impedance Z_o, at both ends of this attenuator is approximately 50 Ω. Calculate the attenuation from V_1 to V_2 in dB, considering that the attenuator is terminated by 50 Ω at antenna jack side.

40. The circuit enclosed in gray bracket must be mounted using ground plane construction technique. The cable connecting the signal generator is the coaxial cable with a BNC on one end. Find a clear spot on the PCB near the harmonic filter and construct the circuit in gray bracket. Connect attenuator output to J3-C79-L7 node using a very short piece of insulated wire or connect the free lead of 47 Ω directly.

41. Set the signal generator to 50% *internal AM modulation* (this produces an AM signal at the output with 1 KHz modulating signal), and the carrier frequency to 29 MHz. Set the output to 20 dBm.

42. Connect the Channel 1 probe to IC12 output (R93 is a good place to connect) and Channel 2 probe to Audio/RX pigtail (across C108). Switch on the power.

43. Set the RF gain pot to maximum gain and audio gain pot to a medium position.

44. Initially you will not see any signal at all. This is because your VFO is off tuned. Using the coarse tuning pot of VFO try to tune to 29 MHz. As you pass 29 MHz, you will hear the whistle out of the speaker. Use the fine tuner pot to tune the VFO to exactly 29 MHz (or maximum sound level).

45. Measure and record the peak and valley amplitude at the output of IC12, and the amplitude and the frequency of the signal across C108. Using this data, calculate the composite gain (conversion gain + amplification) up to the output of IC12.

46. Adjust C92 for maximum amplitude on Channel 1, and C104 for maximum amplitude on Channel 2, if necessary.

47. Gradually decrease the signal generator output until you cannot hear any thing. Record this signal generator output in dBm. Calculate the power delivered to the antenna input in dBm. This is the *sensitivity* of your receiver (subjectively, at least).

48. Install J20 and solder. Make the connections for S2/2, install and solder. Remove the jumper from +15 bus to pin 1 of IC5,you connected in Exercise 3.13.

49. Switch the power off. Disconnect the signal generator and the probes. De-solder and remove the ground plane constructed circuit.

You now have a working transceiver, TRC-10! You need an antenna.

7.7. Problems

1. Calculate the forward gain of the amplifier and the feedback ratio in the VFO circuit at the VFO frequency.

2. What is v(t) in the circuit given in Figure 7.1, if $IR < V_{dc}$?

3. Consider the circuit in Figure 7.19(a). Find the transfer function $H(\omega) = V_2(\omega)/V_1(\omega)$. Determine the frequency at which $\angle H(\omega) = 0$, when R3 = R4 and C3= C4. What is $|H(\omega)|$ at that frequency?

4. The circuit shown in Figure 7.19(b) is called the *Wien-bridge* oscillator. The circuit given in Figure 7.19(a) provides the positive feedback. The circuit oscillates at the frequency ω_o where $\angle H(\omega_o) = 0$, if the condition $(1+R2/R1)H(\omega_o) > 1$ is satisfied. What is the minimum value of R2/R1 so that the circuit can oscillate, if R3 = R4 and C3= C4?

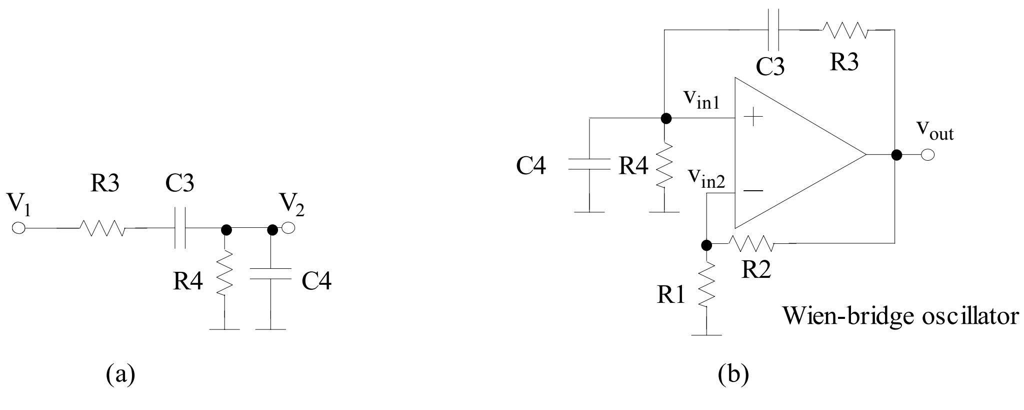


Figure 7.19 Circuits for (a) problem 3 and (b) problem 4.

