In an FM signal, the modulation is the deviation of a carrier from its nominal frequency. The conventional method to demodulate this signal is to convert frequency deviation to phase and detect the change of phase. In the quadrature demodulator, the modulated carrier is passed through an LC tank circuit that shifts the signal by 90° at the center frequency. This phase shift is either greater or less than 90° depending on the direction of deviation. A phase detector compares the phase-shifted signal to the original to give the demodulated baseband signal. You use quadrature demodulators not only for frequency modulation, but also with digital modulation schemes such as FSK (frequency shift keying) and GFSK (Gaussian frequency shift keying).

FM Quadrature Demodulator Block Diagram
The conventional method of FM demodulation for integrated circuits is Bilotti’s quadrature demodulator that uses a phase shift network and a phase detector. Figure 1 shows the block diagram of this quadrature demodulator. The phase detector compares the phase of the IF signal \(v_1\) to \(v_2\), the signal generated by passing \(v_1\) through a phase shift network. This phase shift network includes an LC tank \((L, R_p, and C_p)\) and a series reactance \((C_s)\). The network gives a frequency-sensitive 90° phase shift at the center frequency. The phase detector discussed here is the bipolar
double-balanced multiplier popularized by Bilotti. The output of the multiplier \((I_0)\) is filtered, which results in a DC level that changes as the input frequency changes.

**Figure 1:** Quadrature demodulator block diagram

Quadrature Demodulator Transfer Function
To derive the transfer function of the quadrature demodulator, the phase shift network is first drawn as a small-signal circuit model (Figure 2). The impedance \((Z_p)\) of the parallel combination of \(L, R_p,\) and \(C_p\) is:

\[
Z_p(s) = \frac{R_p L \cdot s}{C_p R_p L \cdot s^2 + L \cdot s + R_p}.
\]  

**Figure 2:** Small-signal model of the quadrature phase-shift network

The ratio of \(v_2\) over \(v_1\) is the ratio of impedances \(Z_p(s)\) over \((Z_p(s) + 1/sC_p)\). Simplifying this ratio,

\[
\frac{v_2}{v_1} = \frac{C_\text{s} L \cdot s^2}{(C_p + C_\text{s}) L \cdot s^2 + \frac{L}{R_p} \cdot s + 1}.
\]  

The resonant frequency \(\omega_0\) of this filter is:
\[ \omega_n = \frac{1}{\sqrt{L(C_z + C_p)}}. \quad (3) \]

The quality factor \( Q \) of the phase shift network is \( R_p/(\omega_n L) \). Next, \textbf{Equation 2} is used to solve for the transfer function from \( v_1 \) to \( v_2 \). The variables \( \omega_n \) and \( Q \) are substituted into \textbf{Equation 2} and \( v_2/v_1 \) is written in terms of \( s = j\omega \): \( \omega_n \approx \omega \):

\[ \frac{v_2}{v_1} = \frac{jQ \cdot \frac{C_z}{C_z + C_p} \cdot \frac{\omega}{\omega_n} \approx jQ \cdot \frac{C_z}{C_z + C_p}}{1 + jQ \cdot \frac{\omega_n^2 (\omega_n - \omega_n)/(\omega_n - \omega_n)}{\omega_n^2 (\omega_n - \omega_n)}} \approx \frac{jQ}{1 + jQ (2\Delta\omega)}. \quad (4) \]

In \textbf{Equation 4}, \( \Delta\omega \) is the deviation from the carrier frequency, and \( 2Q\Delta\omega/\omega_n \) is the normalized deviation. Defining:

\[ a = \frac{2Q\Delta\omega}{\omega_n}, \quad (5) \]

\textbf{Equation 4} can be written as:

\[ \frac{v_2}{v_1} \approx jQ \cdot \frac{C_z}{C_z + C_p} \cdot \frac{1}{\sqrt{1 + a^2}} \cdot \tan^{-1} a \quad (6) \]

Writing \( v_2 \) in terms of \( v_1 \),

\[ v_2 = v_1 \cdot Q \cdot \frac{C_z}{C_z + C_p} \cdot \frac{1}{\sqrt{1 + a^2}} \cdot \left(90^\circ + \tan^{-1} a\right) \quad (7) \]

\textbf{Equation 7} describes the signal at one multiplier input in terms of the signal at the other input. The signal \( v_1 \) is applied to the first input and is in limiting (a square wave). The signal at the second input (\( v_2 \)) is a linear signal. By integrating over half of the period, you get the average value of the multiplier output current:

\[ \bar{i} = \frac{\omega}{\pi} \int_0^{\pi} v_2 \cdot g_m \, dt \quad (8) \]

For a bipolar differential amplifier, \( g_m \) is \( I_o/V_o \) where \( 2I_o \) is the multiplier bias current. Substituting
for $v_2$ and $g_m$,

$$
\frac{i_o}{2I_o} = \frac{\alpha}{\pi} \int_0^{\pi/2} \frac{V_1}{Q} \frac{C_s}{C_s + C_p} \frac{1}{\sqrt{1 + a^2}} \cdot \sin(\omega t + 90^\circ + \tan^{-1} a) \, dt
$$

(9)

where $V_1$ is the peak voltage of the signal $v_1$. Simplifying Equation 9 yields the transfer function for the quadrature demodulator:

$$
\frac{i_o}{2I_o} = \frac{2Q}{\pi} \frac{C_s}{C_s + C_p} \frac{V_1}{2V_T} \frac{1}{\sqrt{1 + a^2}} \frac{a}{\sqrt{1 + a^2}}.
$$

(10)

In Figure 3, the term $a/(1+a^2)$ from Equation 10 is plotted versus the normalized frequency deviation ($a$). This plot is the quadrature demodulator s-curve. As the frequency of the signal applied to the demodulator becomes more positive than the natural frequency of the phase shift network, the filtered output of the multiplier increases. Likewise, the filtered output decreases as the frequency of the input signal decreases.

Figure 3: Plot of normalized demodulator output vs. normalized frequency deviation

Integrated Circuit Implementation

Figure 4 shows an integrated circuit implementation of the quadrature demodulator. The input signal $v_m$ is supplied from a limiting amplifier and is a square wave of known amplitude. The input signal $v_m$ is level shifted, and $v_1$ is applied to transistors $Q_1$ and $Q_2$. The amplitude of $v_1$ is large enough such that $Q_1$ and $Q_2$ are switched completely on or off during each cycle. Capacitor $C_s$ is typically integrated while $C_p$, $L$, and $R_p$ are external components. The component values are chosen such that the amplitude of $v_2$ is less than that of $v_1$ as given by Equation 7. This causes transistors $Q_3$-$Q_6$ to operate as linear devices rather than switches. The output of the multiplier is converted from a differential current to a single-ended voltage $v_o$. The output is filtered by components $R_f$ and $C_f$. 
Figure 4: Integrated circuit implementation of the quadrature demodulator

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