INTERMEDIATE-FREQUENCY AMPLIFIERS

INTERMEDIATE-FREQUENCY AMPLIFIERS FOR AM AND FM

The intermediate-frequency (*IF*) amplifier is the circuitry used to process the information-bearing signal between the first converter, or mixer, and the decision making circuit, or detector. It can consist of a very few or a great many component parts. Generally, it consists of an amplifying stage or device to provide gain and a band-pass filter to limit the frequency band to be passed. The signal to be processed can be audio, video, or digital, using Amplitude Modulation, Frequency Modulation, Phase Modulation or combinations thereof. Several examples are shown in Figs. 13 through 19.

IF amplifiers are also used in radio transmitters to limit the occupied bandwidth of the transmitted signal. Certain modulation methods create a very broad frequency spectrum which interferes with adjacent channels. Regulatory agencies, such as the FCC, require that these out-of-band signals be reduced below a certain permissible level, so they must undergo processing through a bandwidth limiting filter and amplifier.

For each application there are certain design restrictions or rules which must be followed to achieve optimum results.

General IF Amplifier Functions and Restrictions

1. Image Rejection The mixer stages in a receiver convert a frequency below or above the local oscillator frequency to an IF frequency. Only one of these frequencies is desired. The IF frequency must be chosen so that undesirable frequencies or images are removed by the RF amplifier filter and are rejected by the mixer. This may mean that two or three different IF frequencies must be used within the same receiver. The IF frequencies in common use range from 0 Hz to approximately 2.0 GHz.
2. Selectivity Selectivity is required to reject as much as possible of any adjacent channel's interfering signal. Generally this means obtaining a band-pass filter characteristic as close to that of the ideal filter as possible to pass the necessary Nyquist bandwidth (the baseband bandwidth from 0 Hz to the highest frequency to be passed) without introducing harmful amplitude or phase distortion.
3. Gain Gain is required to amplify a weak signal to a useful level for the decision making circuit. This gain must be provided by a stable amplifier that introduces a minimum of noise, so as not to degrade the receiver noise figure. All circuit input and output impedances should be properly matched for optimum power transfer and circuit stability.
4. Automatic Gain Control The amplifier gain must vary automatically with signal strength so that the decision making circuit receives a signal of as nearly constant a level as possible. The stages of the IF amplifier must not be overdriven or go into limiting until after the last band-pass filter to prevent "splattering" or broadening and distortion of the signal.
5. Linearity The amplifier must be linear to prevent distortion of the recovered information. AM receivers should be linear in amplitude, whereas FM or PM receivers should be linear in phase.

Selecting the IF Frequency

Image rejection and signal selectivity are the primary reasons for selecting an IF frequency. Most currently manufactured band-pass filters of the crystal or resonator type are standardized so that the designer can obtain off the shelf components at reasonable cost for these standard frequencies. The standard AM broadcast receiver utilizes a 455 kHz IF filter because extensive experience has shown that this rejects all but the strongest images. Assume that the desired signal is at 600 kHz. A local oscillator operating at 1,055 kHz has an image frequency at 1,510 kHz, which the RF input filter easily rejects. Similarly, an FM receiver operating at 90.1 MHz with an IF frequency of 10.7 MHz has an image at 111.5 MHz, which is rejected by the RF amplifier. A single IF frequency is used in both of these cases. A receiver operating at 450 MHz requires two IF frequencies obtained by using first and second mixers, as in Fig. 16. The first IF amplifier may consist of a relatively broadband filter operating at 10.7 or 21.4 MHz, followed by a second converter and IF stage operating at 455 kHz. The first IF filter is narrow enough to reject any 455 kHz images, and the second IF filter is a narrowband filter that passes only the desired signal bandwidth. If the 455 kHz filter had been used as the first IF filter, the 450 MHz RF filter, being relatively broad, would not have eliminated the image frequency, which is 455 kHz above or below the local oscillator frequency.

Television receivers use a video IF frequency of approximately 45 MHz, because this allows a relatively broad RF filter to pass the broadband TV signal, while still rejecting the images. The video signal from the IF amplifier is AM with an FM sound carrier riding on it. Television sound is generally obtained from a beat, or difference frequency, between the video and sound carriers, which is at 4.5 MHz. Satellite receivers use a broadband first IF frequency covering a frequency block from 900 MHz to 2.1 GHz. This is done by a "Low Noise Block" converter (*LNB*). The second mixer is tunable so that any frequency in the block is converted to the second IF frequency, usually fixed at 70 or 140 MHz. The second IF frequency, which drives the detector, has a narrower bandwidth to reduce noise and reject adjacent channel interference.

Crystal, ceramic resonator and SAW filters are mass produced at relatively low cost for the frequencies mentioned previously, so that most consumer products employ one or more of these standard frequencies and standard mass produced filters.

Selectivity

Carson's rule and the Nyquist sampling theorem, on which it is based, state that a certain bandwidth is required to transmit an undistorted signal. The necessary bandwidth for an AM signal is given as

$$\mathrm{BW} = 2f_m \tag{1}$$

Thus an AM broadcast receiver requires 10 kHz of bandwidth to pass a 5 kHz = f_m audio tone. In data transmission systems, the frequency f_m corresponding to the data rate f_b, is given by $f_m = ½ f_b$. The data clock frequency is twice the frequency of the data in ones and zeros. This means that a baud rate f_b of 9,600 bits per second requires a bandwidth of 9.6 kHz.

For FM, the necessary bandwidth required for transmission is given by

$$\mathrm{BW} = 2(f_m + \Delta f) \tag{2}$$

A 15 kHz audio tone (= f_m) and an FM transmitter deviated with a modulation index of 5 requires 2 [15 + (15 × 5)] = 180 kHz of bandwidth. Δf is (5 × 15) and f_m is 15 kHz. Narrow band FM (with a modulation index less than 0.7) is somewhat different in that the bandwidth actually required is the same as that for AM because the higher Bessel products are missing (Eq. 1).

These values are for "double-sideband" transmission. Single-sideband transmission requires half as much bandwidth. The required baseband bandwidth is the same as the value for f_m. This is also known as the "Nyquist bandwidth," or the minimum bandwidth that carries the signal undistorted at the baseband.

Ideally the IF filter, or the baseband filter, need pass only this bandwidth and no more. This requires an "ideal" band-pass or low-pass filter, which does not exist, but is approached by various means. The filter must be as narrow as conditions permit to reduce the noise bandwidth and any adjacent channel interference, because noise power increases linearly with increasing filter bandwidth.

Gain

The IF amplifier must provide sufficient gain to raise a weak signal at the RF input to the level required, or desired, by the decision making circuit or detector. This receiver gain varies from 0 up to 130 dB, most of which must be provided by the IF amplifier. The RF amplifier and mixer circuits preceding the IF amplifier generally provide 20 or more dB of gain so that the IF amplifier generally contributes little to the receiver noise figure. See "noise figure" elsewhere in this article.

Gain is provided by an amplifying device, such as a transistor or vacuum tube (in older equipment). These devices have complex input and output impedances that must be matched to the filtering circuits for best power transfer, stability, and lowest noise. Current practice is to use a "gain stage" which consists of multiple amplifying devices in an integrated circuit package. These packages contain the mixer stages and detectors.

Automatic Gain Control

Receivers must respond to a wide range of input levels while maintaining a nearly constant level at the detector or decision making circuit. The user or operator does not wish to manually adjust the gain to obtain a constant sound or picture level when changing stations. This function is performed by detecting the output level of the IF amplifier and correcting it by a feed-back circuit that adjusts the level to keep it as constant as possible. Because this detected level varies rapidly, it is passed through a low-pass filter (usually an *RC* pair) to integrate or slow down the changes, then amplified by a dc amplifier, and applied to an IF amplifier circuit or gain stage with variable gain characteristics. Some receivers, such as those in an automobile, require relatively rapid acting automatic gain control (*AGC*) circuits, whereas fixed receivers use a much slower AGC time constant. Dual-gate, field-effect transistors use the second gate to control the gain. Bipolar or single-gate, field-effect transistors vary the gain by a bias voltage or current applied to the input terminal along with the signal. Special integrated circuit gain stages for IF amplification are available, such as the Motorola MC 1350, which amplify and provide a variable gain control function.

Band-Pass Filters for IF Amplifiers

Except for block conversions, which convert wide frequency bandwidths, such as those used on satellite receivers, IF amplifiers generally use a narrow band-pass or a low-pass filter to limit the bandwidth to the Nyquist bandwidth. Block conversion, on the other hand, uses a high-pass/low-pass filter pair where the bandwidth to be passed is between the high and low cutoff frequencies.

The traditional band-pass filter requires a resonant element. Although the actual resonator may be a coil and capacitor, ceramic resonator, or SAW filter, the principles are basically the same. Digital filters which do not use resonators have been employed more recently. These are discussed later in brief. They are discussed in more detail elsewhere in this encyclopedia.

The inductance/capacitor resonator was the first used, and is still a comparison standard. Figure 1(a) shows a series resonant circuit and Fig. 1(b) shows a parallel resonant circuit. These circuits pass a signal at the resonant peak and reject a signal off resonance. R_s and R_p are naturally occurring losses that reduce the circuit efficiency. Figure 2 shows the universal resonance curve which is applicable to both series and parallel resonant circuits. It is important to note that the signal rejection never goes to a zero level in the area of interest, but reaches an asymptotic value of about 0.2 or −15 dB. If it is necessary to reject a signal on the shoulders of this curve by 60 dB, then four cascaded stages of this filter must be used to obtain the necessary rejection. Note also that there is a nonlinear phase shift that reaches a maximum in the area of interest of about ±70°. When stages are cascaded, this phase shift is multiplied by the number of stages. A nonlinear phase shift causes distortion in FM receivers.

A frequency f_0, at which the response of a parallel resonant *LC* filter is a maximum, that is, the point at which the parallel impedance is a maximum, is defined as a "pole." A frequency at which the impedance is a minimum, as in the series *LC* circuit, is defined as a zero. Thus the previous assumed four cascaded stages above constitute a four-pole filter, because it contains four resonant poles. The frequency

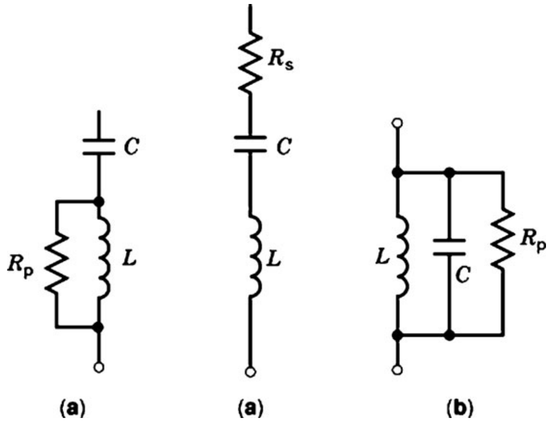

Figure 1. Series and parallel resonant circuits. R_p and R_s are loads that reduce the efficiency of the circuit.

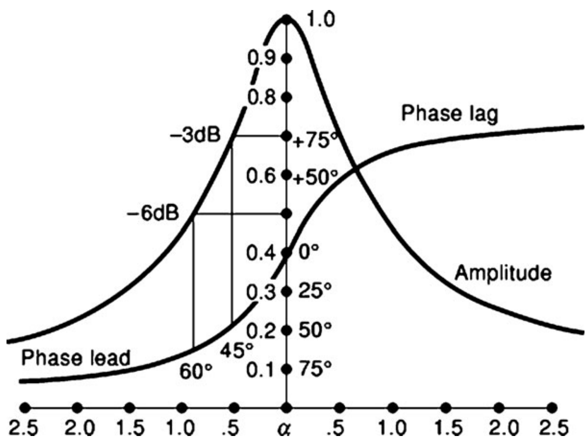

Figure 2. The "Universal Resonance Curve." The phase change is shown for series resonance. The phase reverses for parallel resonance. $\alpha = Q$ (Hz off resonance)/(Resonant frequency).

of resonance is given by Eq. (3). This is the frequency at which $X_c = 1/{-j\omega C}$ and $X_L = j\omega L$ are equal.

$$f_0 = \frac{1}{2\pi\sqrt{LC}} \tag{3}$$

The bandwidth that an analog LC filter passes is altered by the circuit efficiency, or circuit Q, given in Eqs. (4a), (4b), and (4c). Generally the bandwidth is specified as the bandwidth between the −3 dB points.

$$Q = X_c/R_s,\ \text{for a series circuit} \tag{4a}$$

$$Q = R_p/X_c,\ \text{for a parallel circuit} \tag{4b}$$

$$Q = f_0/3\,\text{dB bandwidth} \tag{4c}$$

For simplicity in analyzing the following circuits, the Q determining R is assumed to be a parallel resistance R_p across the inductance.

The amplitude response of the resonant circuit is given by Eq. (5a) and the phase response by Eq. (5b).

$$G(j\omega) = A(j\omega) = \frac{1}{1 + jQ(2\Delta\ f/\ f_o)}$$

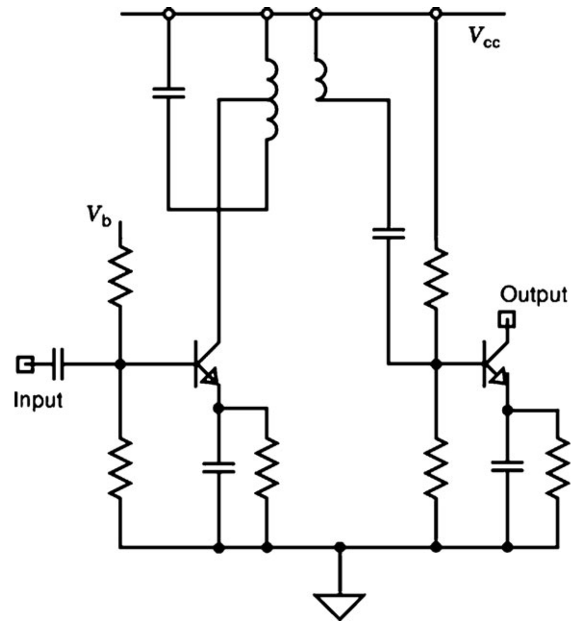

Figure 3. A typical transistor IF circuit with bandpass filter.

$$Tan\varphi = \frac{Q(2\Delta\ f/\ f_o)}{1}$$

Figure 3 shows a typical IF amplifier stage used in earlier transistor radios. In this circuit R'_p (the total shunting resistive load) is actually three resistances in parallel, one the equivalent R_p of the coil itself (representing the coil losses), another the input resistance of the following stage, as reflected, and the third the output resistance of the driving transistor, as reflected. It cannot be assumed that the resulting coil Q and, hence, the selectivity of the circuit is that of the unloaded coil and capacitor alone. Dual-gate, field-effect transistors have the highest shunting resistance values, and bipolar transistors the lowest. The gain is varied by increasing or decreasing the bias voltage V_b applied to the input terminal.

Manufacturers of amplifying devices often provide the impedances, or admittances, of their products on their data sheets. Formerly this was done in the form of h parameters. The more common practice today is to provide the information in the form of S parameters. These values can be converted to impedances and admittances, but the manual process is rather complicated. An easier method is to use the various software programs (see references) to make the conversion. Matrix algebra and "h" and "S" parameters are discussed elsewhere in this article and also in references (3, 4). Unfortunately, S parameters for band-pass filters are rarely available.

Figure 4(a) shows the equivalent circuit of the transistor as the tuned LC sees it. The transistor amplifies a current which is passed through a relatively low driving resistance R_s to the outside. At the same time, the attached LC sees an equivalent shunting resistance R_c and capacitance C_c which must be added in parallel to R_p, L, and C. The input to the following stage, assumed to be an identical transistor, has a relatively low shunting resistance R_i, and capac-

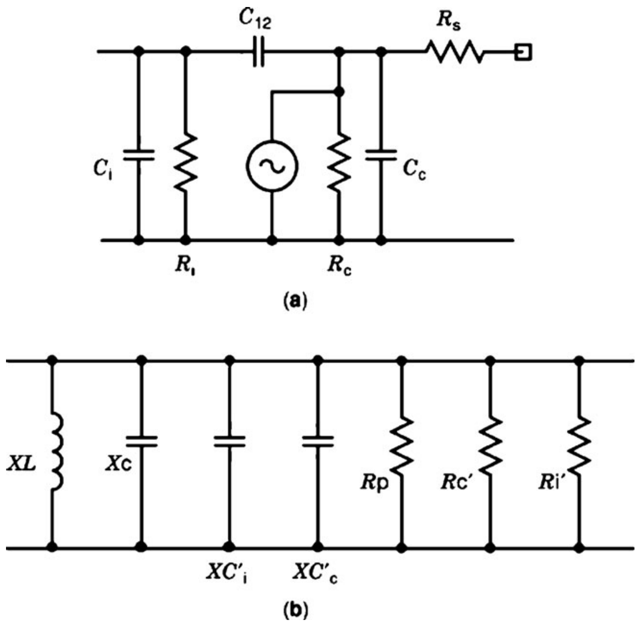

Figure 4. (a) The equivalent circuit of a transistor amplifier. (b) The equivalent circuit of Fig. 3 when all of the shunting loads are included.

itance C_i which must be added. Unless the added capacitances are large compared to the resonant C, they merely add to it without greatly detuning the circuit. When tuned, the total C plus L determine the frequency and the resulting total R'_p determines the Q of the LC circuit, hence the bandwidth. Thus the complex components are tuned out, and the remaining design problem consists of matching the real or resistive part of the input and output impedances to the best advantage.

The desired result is to couple the output of the driving stage to the input of the following stage with the least loss by matching the differing impedances. An additional desired result is to narrow the band of frequencies passed by a filter. These objectives are accomplished by transforming the input and output impedances to a higher or lower shunting impedance that maintains the desired bandpass characteristic of the filter. A low driving or load impedance can be stepped up to a very high impedance which maintains the circuit Q at the desired value.

Impedance matching enables the designer to change the actual impedance to a different apparent value which is optimum for the circuit. Figure 5 shows how impedances are matched by transformer action. A transformer with a 3:1 turn ratio is shown as an example. The output impedance relative to the input impedance is given by Eq. (6). N_i and N_o are the input and output numbers of turns on the winding.

$$Z_i/Z_o = (N_i/N_o)^{1/2} \tag{6}$$

Thus 90 Ω at the input is seen as 10 Ω at the output. The auto transformer [tapped coil in Fig. 5(b)] has the same relationship.

When all of the reactances and resistances from the tuned circuit and the transistor input and output, as modified by the step-up/step-down process of the impedance

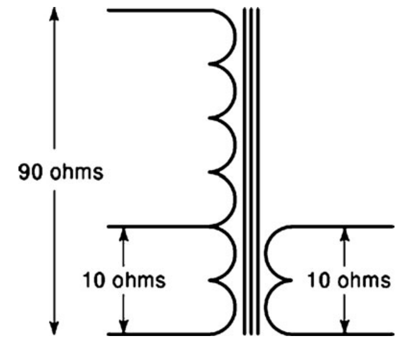

Figure 5. Impedance step up or down using a transformer.

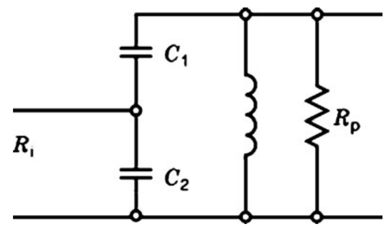

Figure 6. Impedance matching utilizing capacitors.

matching networks are added, the network in Fig. 4(b) results. To calculate the resonant frequency and circuit Q from these reactances and resistances in parallel is complicated unless they are converted to admittances. Software is available at reasonable cost to perform these calculations. See references.

Stock or mass produced IF transformers often do not have the desired turns ratio to match the impedances. An additional Z match circuit using capacitors enables the available transformers to match almost any impedance. This capacitor divider circuit is often used instead of a tapped coil or transformer as shown in Fig. 6.

The formulas for calculating the matching conditions with capacitors are more complex than those used for transformer coupling, because there are more variables. In this circuit R_i, is assumed to be lower than R_p. Although R_p is the equivalent parallel resistance of the LC circuit in Fig. 6, it could also be the reduced resistance or reflected R_{p2} at a transformer tap. N in these equations is equal to the loaded resonator Q, or to a lower arbitrary value if total shunting R_p is lowered by transformer action as in equation (6) or if the component ratios become unwieldy.

$$N = R_p/X_L = Q \tag{7}$$

$$X_{C2} = \frac{R_1}{\sqrt{\dfrac{R_1(N^2+1)}{R_p - 1}}} \tag{8}$$

$$X_{C1} = \frac{R_p N}{N^2+1}\left(1 - \frac{R_1}{N X_{C2}}\right) \tag{9}$$

$$X_{C2} \approx [(R_1 R_p/Q]]^{1/2} \tag{10}$$

$$X_{C1} \approx R_p/Q = X_L \tag{11}$$

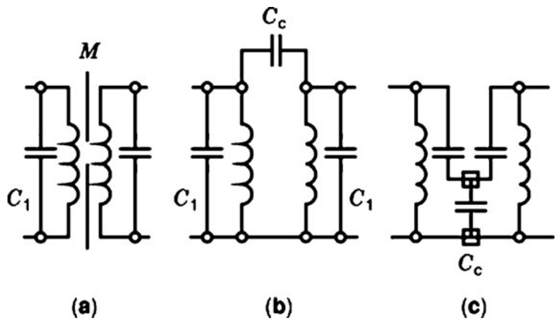

Figure 7. Double-tuned circuits coupled together.

Equations (8) and (9) calculate the reactances of the two capacitors. Note that NX_L is the same as QX_L. Starting with a value of $N = Q$, find X_{C1}, then X_{C2}.

If Q is large in Eq. (7), the equations reduce to the approximate values in Eqs. (10) and (11). Unless Q is less than 10, these approximate equations are accurate enough for general use. As an example, let $R_i = 100\ \Omega$ and $R_p = 10{,}000\ \Omega$ with $Q = 100$. Then using Eq. (10), X_{C2} becomes 10 Ω and X_{C1} becomes 100 Ω. C_2 is approximately ten times as large as C_1. Note the similarity of this ratio to Eq. (6). If a transformer is involved, N becomes much smaller and the full formulas Eqs. (7), (8), and (9) should be used.

Equations (7) through (9) apply for $R_1 < R_p$, and $N > (R_p/R_1 - 1)^{1/2}$.

Double-Tuned Circuits

When two identical LC circuits are coupled together, as in Fig. 7, a number of responses are possible as in Fig. 8. The amplitude response depends on the coupling coefficient K. Undercoupling results in a two-pole filter with the sharpest selectivity. Critical coupling results in the narrowest bandwidth with the highest gain. Transitional coupling is slightly greater than critical coupling and results in a flat topped response with a wider bandwidth. Overcoupling results in a double-humped response with sharper skirts and broad bandwidth. The coupling coefficient is calculated using Eqs. (5a–d). Equation (12a) applies to mutual inductive coupling and Eqs. (12b–d) to capacitive coupling.

$$K = M/(L_1L_2)^{1/2} \tag{12a}$$

$$K_c = 1/(Q_1Q_2)^{1/2} \tag{12b}$$

$$K = [C_c/(C_c + C_1)]^{1/2} \tag{12c}$$

$$K = [C_1/(C_c + C_1)]^{1/2} \tag{12d}$$

Equation (12a) calculates the coupling coefficient for two identical LC tuned circuits that are coupled together by leakage inductance [Fig. 7(a)], often obtained by using shielded coils with holes in the sides of the shield cans to allow the magnetic fields to interact. The size of the hole determines the value of the mutual inductance M. Because this is difficult to control, a coupling capacitor is often used as shown in Fig. 7(b,c). The critical coupling value is given by Eq. (12(b)). The coupling coefficients for Fig. 7(b,c) are given in Eqs. (12 c,d).

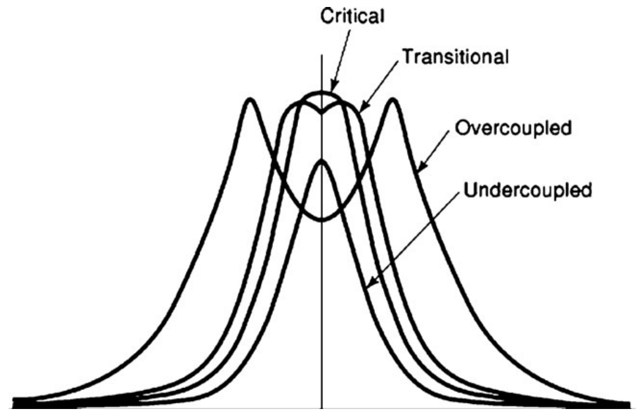

Figure 8. The effects of various coupling coefficients.

The amplitude response curves in Fig. 8 do not yield any information as to the phase shifts that take place through the filter. In AM circuits, phase is generally of little concern, and most attention is paid to the amplitude ripple and linearity. In FM circuits, nonlinear phase shift, or a related term "differential group delay," becomes more of a problem and efforts are made to keep the phase shift as linear as possible. In data transmission circuits using phase modulation, any nonlinearity must be avoided. For these reasons, the coupling coefficients are carefully adjusted, and cascaded IF amplifier stages are used to get the desired transfer function for the IF amplifier.

CASCADING IF AMPLIFIER STAGES AND FILTERS

All filtering actions between the RF receiver input and the decision making circuit are parts of the IF amplifier bandpass filter. Because the decision making circuit is at baseband, or 0 Hz, all filtering before the decision making circuit is part of the IF bandpass filtering, which should be treated as a whole.

A single LC circuit seldom has the desired bandpass characteristic for an IF amplifier. Cascading IF amplifier stages with differing coupling and Q values enables the designer to obtain the desired transfer response. One combination of LC filters uses an overcoupled, double-tuned stage followed by a single tuned stage with a lower Q. The result is a three-pole filter with relatively steep skirt slopes. Cascading these stages results in filters with responses resembling Butterworth, Chebycheff, elliptical, or equal-ripple filters which are noted for rejecting adjacent channel interference. See Figs. 9 and 10.

When additional filtering is required at the baseband, simple RC filters, low-pass LC filters, or digital FIR filters are used. These and other filters are discussed in greater detail elsewhere in this encyclopedia.

Crystal and Ceramic Filters

Figure 11(a) shows the equivalent circuit of a crystal or a ceramic resonator. These devices have both a pole and a zero whose frequencies are located relatively close to each other. Quartz crystals have Q values from 2,000 to 10,000 or more depending on the mechanical loading of the crys-

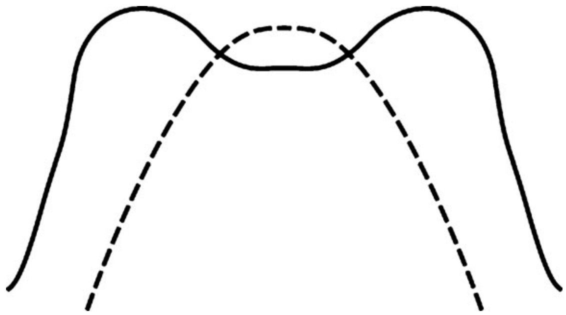

Figure 9. Two amplifier stages cascaded. The first stage is overcoupled; the second stage with a lower Q is critically coupled.

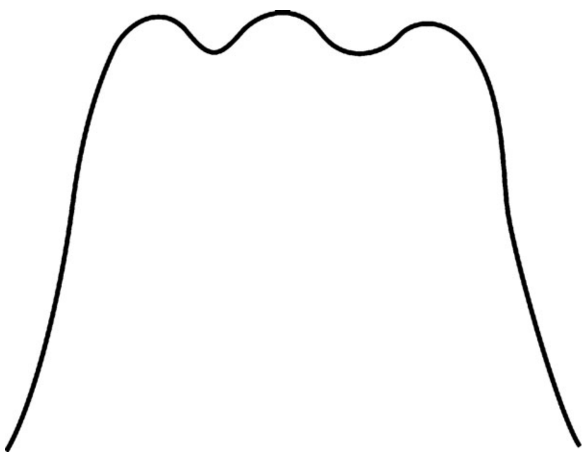

Figure 10. The overall response of the two cascaded stages in Fig. 9.

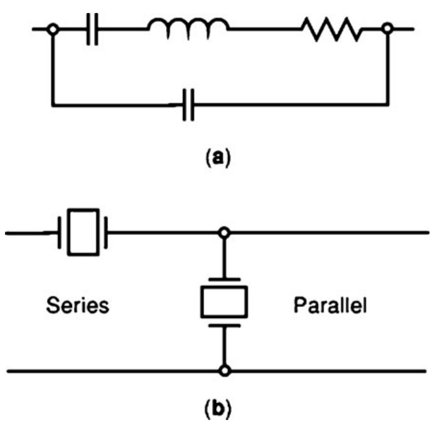

Figure 11. (a) The equivalent circuit of a crystal or ceramic resonator. (b) Two resonators are coupled together to form a bandpass filter.

tal. Ceramic resonators usually have Q values between 100 and 400. The higher the Q, the narrower the filter bandpass. When two of these devices are connected as shown in Fig. 11(b), the result is a band-pass filter with steep skirts, as in Fig. 12. These devices are always used in pairs to make a two-pole filter, which then is combined in a single container with other pairs to create a filter with as many as eight or more poles. They usually have excellent adjacent channel rejection characteristics.

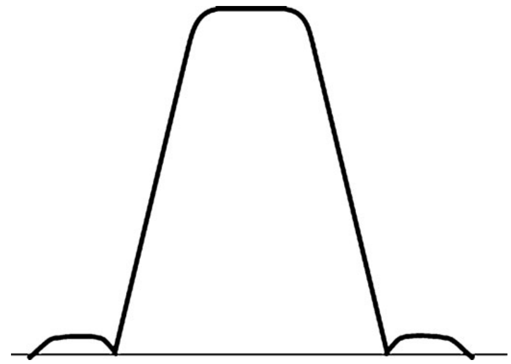

Figure 12. The frequency response of the two-pole crystal resonator in Fig. 11(b).

When using these devices, care must be taken to carefully match the specified impedance. Any impedance mismatch seriously alters the response curve of the filter. The impedance matching techniques previously discussed enable the designer to obtain a very close match which optimizes the circuit performance. Typical input and output impedances range from 50 to 4,000 Ω. Crystal filter makers often build in transformer or other tuned matching circuits so that the user does not need to provide a matching circuit outside the crystal filter.

Surface acoustic wave (*SAW*) filters utilize a crystal oscillating longitudinally with many fingers or taps placed along the surface. They are made with very broad bandpass characteristics, which makes them well suited for TV IF amplifiers, spread-spectrum IF filters, and other uses requiring wide bandwidths. They have losses which are typically about 8 to 20 dB, so they must have amplifiers with adequate gain ahead of them if the receiver noise figure is not to be degraded. They are not suitable for narrowband or low-frequency applications.

Baseband IF Filtering

IF filters with specific response characteristics are sometimes very difficult to obtain, whereas the desired characteristic is easily and inexpensively obtainable at the baseband. This concept is often applied to transmitters where a sharp cutoff filter is obtained with simple components, such as the switched filter. An eight-pole equivalent at baseband becomes a 16-pole filter at the modulation IF frequency. For example, a sharp cutoff filter for voice with a 4 kHz audio cutoff results in a bandpass filter 8 kHz wide at RF after modulation, with the same sharp cutoff. The same cutoff characteristics at RF are almost impossible to obtain in a crystal filter, which is also very costly and beyond the budget for a low-cost transmitter, such as a cordless telephone. By using baseband filtering, a poor-quality RF filter that rejects only the opposite image is used. Similarly, a wideband or poor-quality IF filter is used ahead of a detector if the undesired signal components are filtered off later at baseband by a sharp cutoff filter.

Switched capacitor filters are available as packaged integrated circuits that are used at baseband and some lower IF frequencies. They have internal operational amplifiers with a switched feedback capacitor, the combinations of which determine the filter characteristics. Because they depend on the speed of the operational amplifiers and the values of the feedback capacitors, they seldom function much above 100 kHz. They can be configured as Bessel, Equal-ripple and Butterworth filters. Typical of this type of filter are the LTC 1060 family manufactured by Linear Technology Corporation and the MAX274 from Maxim. As Bessel type filters, they perform well out to about 0.7 times the cutoff bandwidth, after which the phase changes rapidly and the Bessel characteristic is lost.

Digital signal processing (*DSP*) at baseband is widely used to reduce the component count and size for baseband filters in very small radio receivers, such as cordless and cellular telephones. Almost any desired filter response is obtained from DSP filters without inductances and capacitors which require tuning. FIR filters have a flat group delay response and are the best choice for FM or PM filtering.

Amplifying Devices for IF Amplifiers

Transistors in one form or another are the standard for IF amplifiers. The single, bipolar or field-effect transistor (*FET*) used as an individual component, was formerly the preferred device. For very high Q circuits, the dual-gate FET performs best, because it is the most stable and offers the lowest shunt resistance. Single-gate, FET devices often have too much drain-to-gate capacitance for good stability. Modern bipolar transistors usually have good stability, but higher shunt resistances than dual-gate FETs. Stability is discussed later in this section.

MMIC devices are stable and have good gain, but the shunt impedance is too low for any bandpass filter except a crystal filter matched to 50 Ω.

The most recent practice for IF amplifiers is to use integrated circuit blocks containing more than one transistor in a gain stage. These are then packaged together in an integrated circuit with other circuit components to form an almost complete radio. Integrated circuits of this type are shown later.

Typical Consumer IF Amplifiers

Consumer radio and TV equipment is mass produced at the lowest possible cost consistent with reasonable quality. Manufacturers of integrated circuits now produce single-chip IF amplifiers that are combined with mass produced stock filters to produce a uniform product with a minimum of adjustment and tuning on the assembly line. In the examples that follow, some circuit components inside and outside the IC have been omitted to emphasize the IF amplifier sections.

Figure 13 shows a single-chip AM receiver that uses the Philips TDA 1072 integrated circuit and ceramic IF filters at 455 kHz. The input impedance of the ceramic filter is too low to match the output impedance of the mixer, so that a tuned matching transformer is used to reduce the passed bandwidth and to match the impedances. The input impedance of the IF amplifier was designed to match the average impedance of the ceramic filters available. This integrated circuit has a built-in automatic gain control that keeps the received audio output level relatively constant at

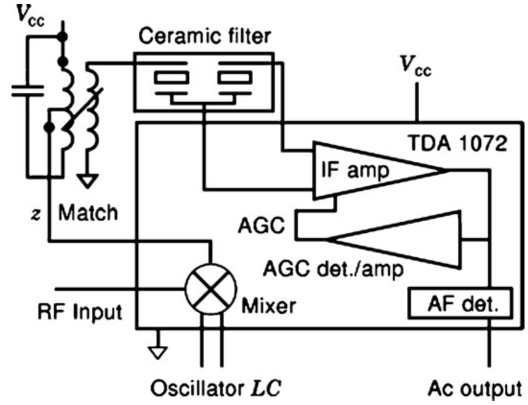

Figure 13. An integrated circuit with built-in IF amplifier that comprises an almost complete AM radio.

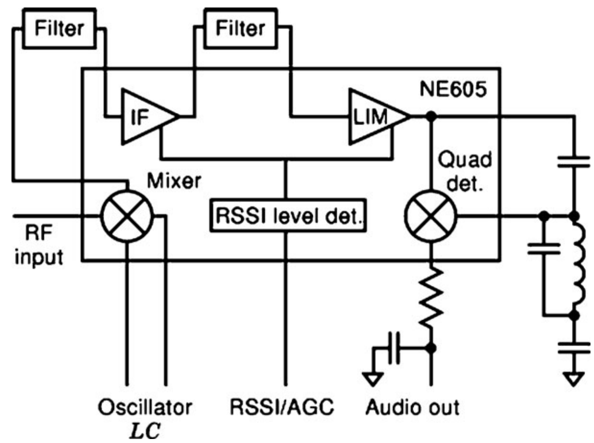

Figure 14. An integrated circuit for FM use that comprises an almost complete FM receiver.

250 mV as long as the input signal level to the chip exceeds 30 μV.

Figure 14 shows a single-chip FM radio based on the NXP (Phillips) NE605 integrated circuit that uses ceramic IF filters at 10.7 MHz. The input and output impedance of the IF amplifier sections is approximately 1500 Ω to match the ceramic filter impedance, so that no matching transformer is required. The audio output is maintained constant at 175 mV for all signal levels at input levels from −110 dBm to 0 dBm. An automatic frequency control (*AFC*) voltage is obtained from the quadrature detector output.

AGC is available from all FM integrated circuits so that the gain of the mixer and RF stages is controlled at a level that does not allow these stages to be saturated by a strong incoming signal. Saturation or nonlinearity before filtering results in undesirable signal spreading. The NE605 has a "Received Signal Strength Indicator" (*RSSI*) output which is amplified and inverted if necessary to provide an AGC voltage or current for the RF amplifier and Mixer.

Figure 15 shows a TV IF amplifier using the Motorola (Freescale) MC44301/2 Video IF integrated circuit with a SAW filter at 45 MHz. The SAW filter band-pass is approximately 6 MHz wide to pass the video and sound. The circuit has both AFC and AGC features built in. Unlike the IF amplifiers used for AM and FM audio broadcast

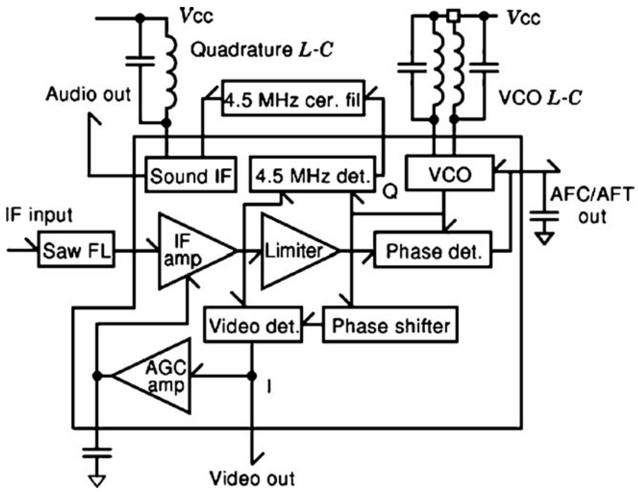

Figure 15. An integrated circuit for a TV IF amplifier that detects both the AM video and the FM subcarrier sound.

applications, the TV IF amplifier includes a phase-locked loop and a synchronous detector which locks the frequency of an internal oscillator to the IF frequency. This locked, or synchronous oscillator output is then mixed with the information-bearing portion of the signal to create a baseband signal. This is one of a family of 0 Hz IF amplifiers which are becoming more popular in radio designs, because they permit most or additional signal processing at baseband. In the TV case, the video and sound carriers are both passed by the SAW filter. They beat together at 4.5 MHz in the detector, providing a second IF stage with the sound information. This 4.5 MHz IF information is then filtered by a ceramic filter approximately 50 kHz wide to remove any video components, limited, and detected as a standard FM signal to provide the TV sound. Then the video portion, consisting of signals from 15 kHz to approximately 4.25 MHz, is further processed to separate the color information at 3.58 MHz from the black and white information. The video output level is detected to provide the AGC voltage.

The phase-locked oscillator, operating at the IF frequency, also provides automatic frequency control to the first mixer stage local oscillator.

Figure 16 shows a dual conversion receiver for communications utilizing the Motorola (Freescale) MC13135 integrated circuit. When the receiver is operated at 450 MHz or 850 MHz, as mentioned previously, single conversion IF stages do not offer the necessary image rejection. This receiver is for narrowband AM or FM as opposed to wideband FM for entertainment purposes. The first IF filter is a low-cost ceramic filter at 10.7 MHz. The second filter is a multipole crystal or ceramic filter with a band-pass just wide enough to pass the audio with a small FM deviation ratio. Radios of this type are used for 12.5 kHz and 25 kHz channel spacings for voice-quality audio. Analog cellular telephones, aircraft, marine, police, and taxicab radios are typical examples.

Direct Conversion and Oscillating Filters

Direct conversion converts the RF frequency directly to the baseband by using a local oscillator at the RF frequency. The TV IF amplifier with the detector circuit given in Fig. 15 illustrates some of the reasons. Conversion to baseband occurs at the IF frequency or directly from the RF frequency.

There is a noticeable trend in integrated circuit design to utilize synchronous detection and restore the carrier by a phase-locked loop, as in Fig. 15, or by regenerative IF amplifiers, to achieve several desirable features not obtainable from classical circuits with square law detectors.

In the case of direct RF to baseband conversion, there is no IF stage in the usual sense, and all filtering occurs at the baseband. For this reason, direct conversion receivers are referred to as zero Hz IF radios. Integrated circuits for direct RF conversion are available that operate well above 2.5 GHz at the RF input.

It was discovered in the 1940s that the performance of a TV receiver is improved by using a reconstructed synchronous or exalted carrier, as occurs in the TV IF amplifier described in Fig. 15. The carrier is reduced by vestigial sideband filtering at the transmitter and contains undesirable AM signal components. By locking an oscillator to, or synchronizing it with the carrier, and then using it in the detector, a significant improvement in the received signal is achieved. Before using circuits of this type, the intercarrier sound at 4.5 MHz in earlier TV sets had a characteristic 60 Hz buzz due to the AM on the carrier. By substituting the recovered synchronous carrier instead, this buzz was removed. Figure 15 is an example.

The earliest direct conversion receivers using locked oscillators or synchronous detectors were built in the 1920s, when they were known as synchrodyne or homodyne receivers. The theory is relatively simple. A signal from the RF amplifier is coupled to an oscillator causing a beat or difference frequency. As the frequencies of the two sources come closer together, the oscillator is pulled to match the incoming signal. The lock range depends on the strength of the incoming signal. Then the two signals are mixed to provide a signal at the baseband, which is further filtered by a low-pass filter. In this way, a relatively broad RF filter is used, whereas the resulting AM signal bandwidth after detection and baseband filtering is very narrow. The Q of the oscillator tank circuit rises dramatically with oscillation, so that Q values of 6,000 to 10,000 are not unusual and selectivity is greatly improved. AGC is obtained from the audio signal to maintain a constant input signal to insure a good lock range. An undesirable characteristic is the whistle or squeal that occurs between stations. Later receivers used a squelch circuit to make the signal audible only after locking occurred. High-quality receivers for entertainment and communications, which use this principle have been produced in the 1990s. They offer higher sensitivity, better fidelity and more controlled response. Integrated circuits for receivers of this type (direct conversion) are now being produced for paging, direct broadcast TV, and for cellular and cordless telephones. The MAX 2101 and MAX 2102 integrated circuits are typical examples.

Oscillating filters and phase-locked loops are similar in principle. An IF frequency is applied to a phase/frequency detector that compares the IF carrier frequency with the oscillator frequency. An error voltage is created that changes the oscillator frequency to match or become coherent with that of the incoming IF carrier frequency. In some cases the phase-locked loop signal is 90° out of phase with the carrier, so that a phase shifter is used to restore the phase and make the signal from the oscillator coherent with the incoming signal. See Figs. 15 and 19 where phase-locked loops and phase shifters are employed.

Synchronous oscillators and phase-locked loops (*PLL*) extend the lower signal to noise ratio, and they also have a bandwidth filtering effect as well. The noise bandwidth of the PLL filter is the loop bandwidth, whereas the actual signal filter bandwidth is the lock range of the PLL, which is much greater. Figure 17 shows the amplitude and linear phase response of a synchronous oscillator. The PLL is not always the optimum circuit for this use because its frequency/phase tracking response is that of the loop filter. The locked oscillator (6) performs much better than the PLL because it has a loop bandwidth equal to the lock range without sacrificing noise bandwidth, although with some phase distortion. Some authors hold that the synchronous oscillator and locked oscillator are variations of the PLL in which phase detection occurs in the nonlinear region of the oscillating device and the frequency-change characteristic of the voltage-controlled oscillator (*VCO*) comes from biasing of the oscillator. Both the PLL and the locked oscillator introduce phase distortion in the detected signal if the feedback loop is nonlinear. A later circuit shown in Fig. 18 has two feedback loops and is considered nearly free of phase distortion (5). This circuit has the amplitude/phase response given in Fig. 17.

Phase-locked loops have been used for many years for FM filtering and amplification. They are commonly used with satellite communication links for audio and video reception. A 74HC4046 phase-locked loop integrated circuit operating at 10.7 MHz (the FM IF frequency) is used to make an FM receiver for broadcasting (7). The phase-locked loop extends the lower signal-to-noise limit of the FM receiver by several dB while simultaneously limiting bandwidth selectivity to the lock range of the PLL. The detected audio signal is taken from the loop filter.

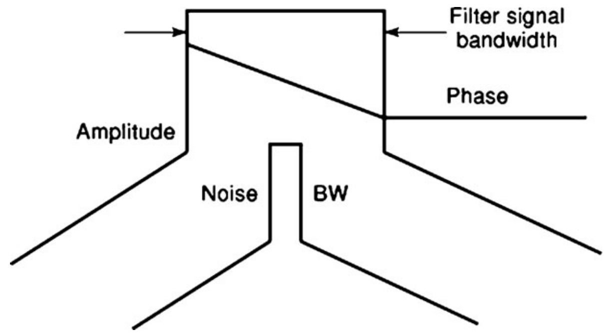

Figure 17. The filtering effect of a synchronous oscillator used as a band-pass filter. The filter bandpass is the tracking range. The noise bandwidth is limited by the Q of the oscillating circuit.

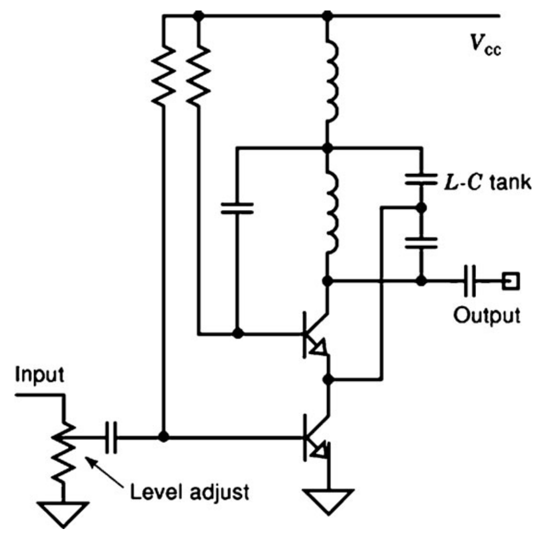

Figure 18. Schematic diagram of a synchronous oscillating band-pass filter.

AM STEREO (C-QUAM)

AM stereo radio is another application of the phase-locked oscillator at the IF frequency. AM Stereo radio depends on two programs transmitted at the same time and frequency. They arrive at the receiver detector circuitry through a

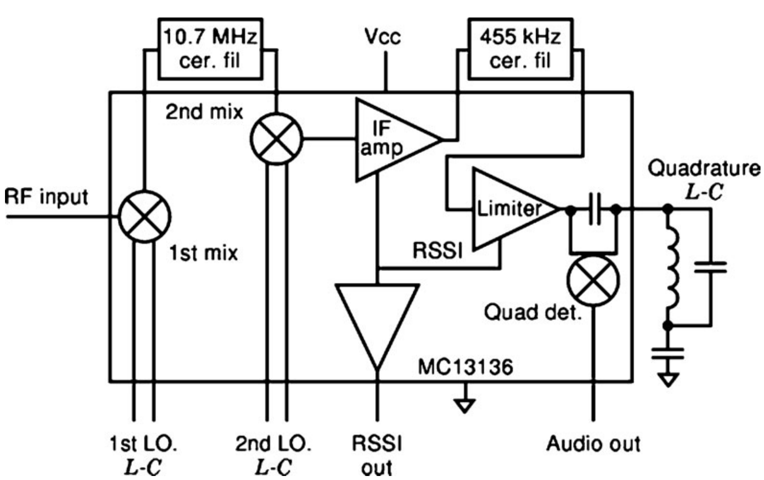

Figure 16. A dual conversion integrated circuit use for high-frequency voice communications.

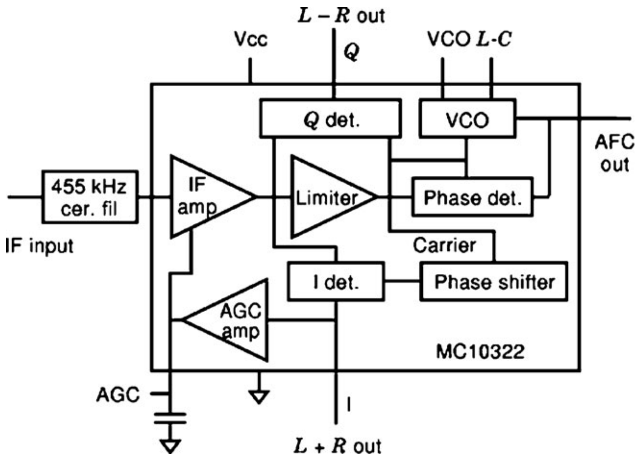

Figure 19. An integrated circuit to detect AM Stereo (C-QUAM). The $L + R$ and $L - R$ signals are in quadrature to each other.

common IF amplifier operating at 455 kHz. The normal program heard by all listeners is the $L + R$ program. The stereo information ($L - R$) is transmitted at the same frequency, but in quadrature phase to the $L + R$ program. Quadrature, or orthogonal transmission, is used because the orthogonal channels do not interfere with one another. Each program section requires a carrier coherent with its own sideband data. The $L + R$ program, which has a carrier, uses an ordinary square law detector or a synchronous detector. This is the program heard over monaural radios. To obtain the $L - R$ program which is transmitted without a carrier, a phase-locked loop is used at the IF frequency to lock a voltage-controlled oscillator to the carrier of the $L + R$ program. This carrier is then shifted 90° in phase and becomes the carrier for the $L + R$ segment. The output of the PLL has the proper phase for the $L - R$ detector, so that phase shifting is not necessary. The $L - R$ detector is a coherent or synchronous detector that ignores the orthogonal $L + R$ information. By adding, and inverting and adding, the left and right channels are separated. Figure 19 shows a simplified block diagram of the C-QUAM receiver.

The Motorola (Freescale) MC1032X series of integrated circuits is designed for AM Stereo use. The MC10322 and MC10325 have most of the components required, including the IF amplifiers, for a complete AM Stereo receiver in two integrated circuit packages.

Subcarriers

Subcarriers carry two or more signals on the same carrier. They differ from the orthogonal signals used with C-QUAM in that they are carried as separate signals superimposed over the main carrier information, as in the video sound carrier in Fig. 15. In Fig. 15, a frequency-modulated subcarrier at 4.5 MHz is carried on top of the main video signal information, which extends from 0 to 4.25 MHz. This is an example of an AM/FM subcarrier. Nondigital satellites utilize a frequency-modulated video carrier with as many as 12 subcarriers at frequencies ranging from 4.5 MHz to 8.0 MHz. Normal FM Stereo broadcasting utilizes a FM/AM subcarrier at 38 kHz to carry the *L–R* portion of the stereo program. FM stations also frequently carry additional subcarriers at 67 and 92 kHz. These FM/FM subcarriers carry background music, ethnic audio programs, and digital data.

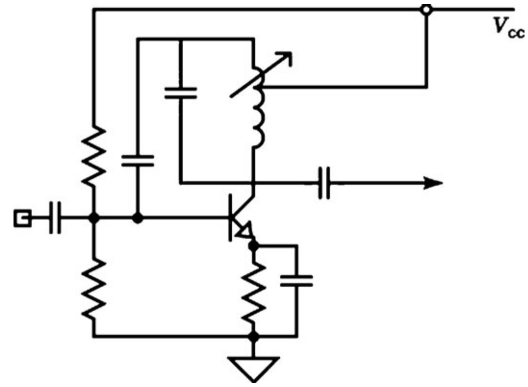

Figure 20. Neutralization or unilateralization of a transistor amplifier to prevent oscillation due to feedback.

To detect a subcarrier, the signal is first reduced to the baseband, then a band-pass filter is used that separates only the subcarrier frequencies. Then the subcarrier frequencies are passed to a second detector which is of the proper type for the subcarrier modulation. This is seen in Fig. 15 where a 4.5 MHz filter is used. This is followed by a limiter and quadrature detector as is appropriate for an FM signal. In the case of a 67 kHz FM/FM subcarrier, the filter is 15 kHz wide at 67 kHz. Detection is accomplished by a discriminator, quadrature detector, or PLL.

Cellular and Cordless Telephones

Analog cellular telephones employ the circuits shown in Figs. 14 and 16. Digital telephones utilizing GMSK also use these circuits. Digital telephones using QAM or PSK employ circuits similar to that used for C-QUAM with digital filtering and signal processing instead of audio filtering at the baseband. The PLL for digital receivers is a more complex circuit known as the "Costas Loop," which is necessary to restore a coherent carrier for digital data recovery. Some cellular phones are dual mode, that is, they transmit and receive analog voice or digital GMSK modulation using circuits similar to Figs. 14 and 16.

Neutralization, Feedback, and Amplifier Stability

Earlier transistors and triode vacuum tubes had considerable capacitance between the output element (collector or plate) and the input side of the device. (See Fig. 4.) Feedback due to this capacitance is multiplied by the gain of the stage so that enough signal from the output was often coupled back to the input to cause the stage to oscillate unintentionally, as opposed to the planned oscillation of the locked oscillator, synchronous oscillator, or PLL. To prevent this, feedback of an opposite phase was deliberately introduced to cancel the undesired feedback. A neutralized IF amplifier is shown in Fig. 20. Transistors and integrated circuits made since 1985 are rarely unstable and generally do not require neutralization unless seriously mismatched. A solution better than neutralization is usually to improve the matching of the components and the circuit layout.

By carefully controlling the feedback, a regenerative IF amplifier that operates on the verge of oscillation can be constructed. This greatly increases the Q of the tuned circuit, thus narrowing the IF bandwidth. Circuits of this type were once used in communications receivers for commercial and amateur use where they were called Q multipliers.

The maximum stable gain (*MSG*) achieved from a potentially unstable amplifier stage without neutralization is obtainable from S parameters and is calculated from Eq. (13). This equation assumes that the input and output impedances are matched and that there is little or no scattering reflection at either the input or output. The stability factor K, usually given with the S parameters, must be >1. A failure to match the impedances can result in an unstable amplifier, but does not necessarily do so. A higher gain is obtained, but at the risk of instability.

$$\mathrm{MSG} = S_{21}/S_{12} \tag{13}$$

The most frequent cause of amplifier instability or oscillation is poor circuit board layout or inadequate grounding and shielding, not the device parameters. The wiring, whether printed or hand wired, forms inductive or capacitive coupling loops between the input and output terminals of the amplifying device. This is particularly noticeable when high gain ICs, such as the NXP SA636, are used. These integrated circuits have IF gains of over 100 dB and require very careful board layouts for best results. Undesirable feedback greatly decreases the usable gain of the circuit.

Software Radio

Digital radios, or radios based on digital signal processing (*DSP*), offer some technical advantages over their analog predecessors. Digital radios are used for digital modulation, and also for AM and FM. One receiver simultaneously detects both digital and analog modulation, thus they can be used for cellular telephones in environments where multiple modulation standards are used. As a class, they belong to the zero Hz IF frequency group.

The typical receiver consists of a conventional RF front end and a mixer stage that converts the signal to a lower frequency, as in the dual conversion radios discussed previously (Fig. 16). The signal at this stage is broadband, but not broadband enough to include the image frequencies. Then the signal is fed to an analog-to-digital (*ADC*) converter which is sampled at several times f_{m}. This converts the portion of interest of the signal to the baseband (or 0 Hz) instead of a higher IF frequency. The actual filtering to remove unwanted interfering signals then takes place at baseband by baseband by digital filtering. Digital signal processing and decimation are covered elsewhere in this encyclopedia. The ADC performs the same functions as the oscillating detectors shown previously.

Noise figure, amplification, and AGC considerations of the first IF amplifier are the same as those for a conventional receiver. The ADC and the DSP filters function best with a constant signal input level.

The term "Software Radio" has been adopted because the tuning function is done in software by changing the sampling frequency at the ADC. The sampling frequency is obtained from a digitally controlled frequency synthesizer instead of tuned LC circuits.

Spread-Spectrum Radios

The spread-spectrum receiver also uses a conventional front end with a wideband first IF stage. The same conditions apply as to software radios and dual conversion receivers. The first IF stage must have the necessary bandwidth to accommodate the spread bandwidth to amplify it with minimum added noise, and to match the output to the despreading circuitry. Spread spectrum is covered elsewhere in this encyclopedia. Although usually associated with digital reception, spread-spectrum technology is also used for analog audio.

Computer-Aided Design and Engineering

For IF filter design, the admittances rather than the impedances are easiest to use, because most components are in parallel as shown in the equivalent circuit of Fig. 4(b). Unfortunately, most available data is in the form of S parameters which are very difficult to convert manually to impedances or admittances. Parameters for the filters are rarely available, so that calculated values based on assumed input and output impedances must be used unless test equipment capable of measuring return losses or standing waves is available. Then the S parameters are measured or calculated.

Smith and Linville charts have been used by some authors to design IF amplifiers, but these methods are not totally satisfactory for IF amplifier design, because a high Q circuit has its plot near the outer edge of the circle and changes are difficult to observe. The network admittance values shown in Fig. 4 would be used.

Computer programs that handle linear or analog designs, such as the various "Spice" programs, are readily available. Other programs which concentrate on filter design can simplify filter design. They have outputs which interface with the Spice programs if desired. Most semiconductor manufacturers provide scattering parameters (S parameters) or Spice input data on disk for use with these programs. Some design software sources are listed in the bibliography.

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