

LM2889 R.F. Modulator

National Semiconductor
Application Note 402
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Introduction

Two I/C RF modulators are available that have been especially designed to convert a suitable baseband video and audio signal up to a low VHF modulated carrier (Channel 2 through 6 in the U.S., and 1 through 3 in Japan). These are the LM1889 and LM2889. Both I/C's are identical regarding the R.F. modulation function—including pin-outs—and can provide either of two R.F. carriers with dc switch selection of the desired carrier frequency. The LM1889 includes a crystal controlled chroma subcarrier oscillator and balanced modulators for encoding (R-Y) and (B-Y) or (U) and (V) color difference signals. A sound intercarrier frequency L-C oscillator is modulated using an external varactor diode. The LM2889 replaces the chroma subcarrier function of the LM1889 with a video dc restoration clamp and an internally frequency modulated sound intercarrier oscillator.

Modulation Parameters

In the U.S., either of two R.F. channels is made available so that the user can select a vacant channel allocation in his geographic area, thus avoiding co-channel problems with

older receivers that have inadequate shielding between the antenna input and the tuner.

The characteristics of the R.F. signal are loosely regulated by the FCC under part 15, subpart H. Basically the signal can occupy the standard T.V. channel bandwidth of 6 MHz, and any spurious (or otherwise) frequency components more than 3 MHz away from the channel limits must be suppressed by more than -30 dB from the peak carrier level. The peak carrier power is limited to 3 mVrms in 75Ω or 6 mVrms in 300Ω, and the R.F. signal must be hard-wired to the receiver through a cable. Most receivers are able to provide noise-free pictures when the antenna signal level exceeds 1 mVrms and so our goal will be to have an R.F. output level above 1 mVrms but less than 3 mVrms. Since the distance from the converter to the receiver is usually only a few feet, cable attenuation will rarely be a problem, but mis-termination can change both the amplitude and relative frequency characteristics of the signal.

The standard T.V. channel spectrum has a picture carrier located 1.25 MHz from the lower band edge. This carrier is amplitude modulated by the video and sync signal. In the

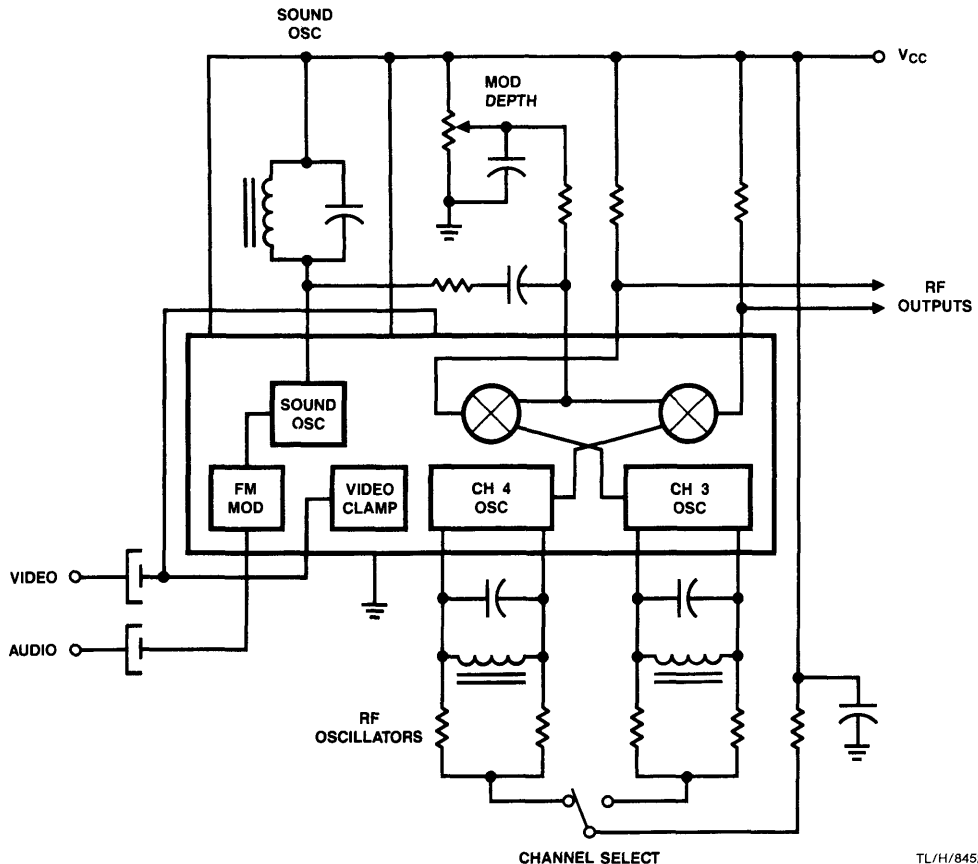
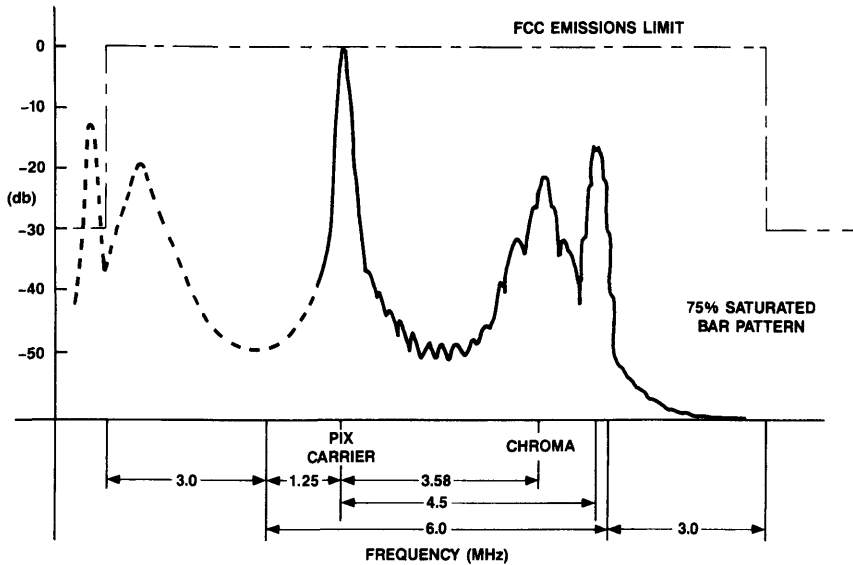


FIGURE 1. The LM2889 Block Diagram With External Components

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Modulation Parameters (Continued)



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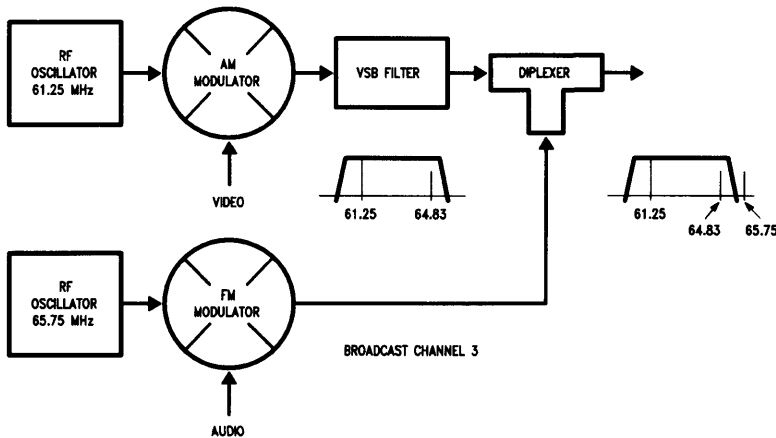
FIGURE 2. Television Channel R.F. Spectrum

case of a color signal, a second subcarrier is added 3.58 MHz above the picture carrier. The sound or aural carrier is 4.5 MHz above the picture carrier and is frequency modulated with the audio signal to a peak deviation of 25 kHz. This audio signal has pre-emphasis above 2.1 kHz (a 75 μ s time constant). Similar modulation methods and standards are used in Japan and Europe.

With the picture carrier located near one end of the channel bandwidth, most of the available spectrum is used by the upper sideband modulation components. Only modulating frequencies within 0.75 MHz of the carrier frequency are transmitted double sideband and the lower sideband is truncated by at least -20 dB compared to the peak carrier level by the time the lower channel edge is reached. This is referred to as Vestigial Sideband (VSB) modulation and since most R.F. modulators are double sideband, a VSB filter is used at the transmitter output. A filter is needed for each

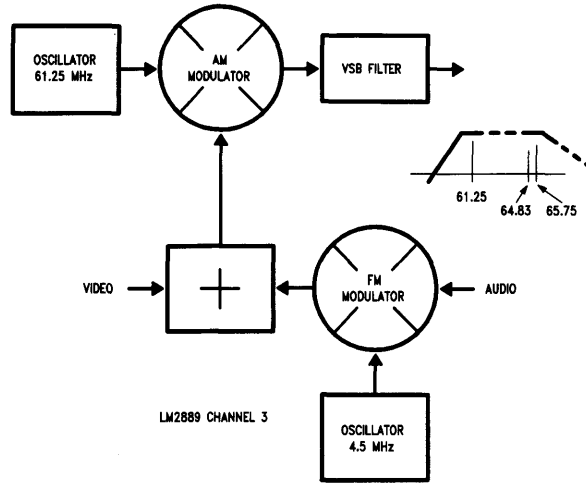
channel and consists of bandpass and harmonic filter sections. A broadcast transmitter uses a separate modulator for the sound carrier and this is added to the picture carrier via a diplexer before reaching the transmitting antenna. Close control is maintained on the picture and sound carrier frequencies to keep a 4.5 MHz spacing between them. This tight frequency control is used to advantage by the majority of television receivers which employ intercarrier sound circuits. The I.F. amplifier processes both the pix and sound I.F. carriers and detects the 4.5 MHz difference frequency at the video detector stage. This frequency modulated sound intercarrier is then stripped of amplitude modulation by a high gain limiter circuit and a quadrature demodulator recovers the audio.

The LM1889 and LM2889 use a slightly different modulation scheme to that described above for several reasons. For circuit economy L-C oscillators are used to generate the pix



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FIGURE 3. Broadcast Transmitter Block Diagram



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FIGURE 4. LM1889/2889 Sound and Video Modulation

carrier frequencies. The stability of such oscillators is good enough for the AFT circuits in modern receivers to maintain picture quality, but if a separate L-C sound carrier oscillator were used, the relative drift of the two carrier frequencies would be much too great for intercarrier sound receivers. For example, a typical television sound circuit tuned to 4.5 MHz will generate as much as 3% distortion if the difference between the R.F. carriers changes by 15 kHz. Apart from the difficulty of setting the initial frequency with sufficient accuracy, it is unlikely that two L-C oscillators could be kept within 15 kHz of each other at 60 MHz to 100 MHz operating frequencies. However, when the audio signal is modulated onto a 4.5 MHz intercarrier oscillator frequency and this carrier is used to modulate the picture carrier, we have only the 4.5 MHz oscillator drift to worry about.

A less obvious problem, but nevertheless significant if good audio quality is to be obtained, is incidental carrier phase modulation (ICPM). Even broadcast transmitters cannot maintain an invariant carrier phase as the modulation depth changes. Without feedback loops to control ICPM, a broadcast transmitter can produce from 3 degrees to as much as 30 degrees phase change as the carrier modulation decreases from sync tips to peak white. While the separate sound carrier is unaffected by this ICPM of the pix carrier, on reception in the intercarrier sound receiver the phase shift with picture information is transferred onto the 4.5 MHz sound intercarrier. This results in a phenomenon known as sound buzz. Even with exceptionally careful p.c.b. layout, an I/C modulator with L-C oscillators can expect the pix carrier frequency to change with modulation depth. Fortunately, by modulating the sound signal as a 4.5 MHz intercarrier onto the pix carrier, the ICPM occurs equally in both R.F. carriers and will not be detected by the intercarrier receiver.

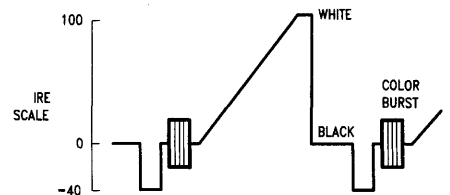
Video Modulation

The baseband input to the modulator is in an easily recognized composite format and this is a convenient point at which to introduce the I.R.E. scale. This is an oscilloscope scale divided into 140 units. The video portion of the signal representing the scene (picture) brightness levels will occupy the 0 to 100 I.R.E. portion of the scale, with 0 I.R.E. as black level and 100 I.R.E. as peak white level. From 0 to

-40 I.R.E. is the synchronization portion of the signal. The usefulness of this scale is that the standard composite video signal will always have a sync amplitude that can be normalized to 40 I.R.E. Similarly the color burst amplitude is always 40 I.R.E. For a 1V (p-p) video signal, an I.R.E. unit is equivalent to 7.5 mV.

Although the video is amplitude modulated on the carrier waveform, the carrier amplitude only decreases from the unmodulated level. This contrasts with standard AM where the carrier level alternately increases and decreases about the unmodulated level. For a television signal, the peak unmodulated level corresponds to sync tip level and increasing brightness levels cause decreasing carrier levels. To prevent complete suppression of the carrier (and consequent loss of the sound intercarrier in the receiver) the peak white signal is limited to a maximum modulation depth of 87.5% of the peak carrier. Returning to our I.R.E. scale we can see that from peak carrier to zero carrier is equivalent to 160 I.R.E. ($140/0.875 = 160$). One obvious consequence of this modulation scheme is that the video signal MUST BE dc coupled to the modulator. AC coupling will cause the peak carrier level to change with modulation scene brightness (standard AM) and the sync modulation amplitude will change. This spells trouble for the receiver sync circuits and the changing R.F. carrier black level will cause errors in displayed brightness—the picture will “wash out” or disappear into black.

The LM2889 uses doubly balanced modulator circuits with an L-C oscillator switching the upper transistor pairs. The signal is applied across the lower transistor pairs. If the signal input pins 10 and 11 are at the same dc potential, the



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FIGURE 5. Video Modulating Signal (in terms of the I.R.E. Scale)

Video Modulation (Continued)

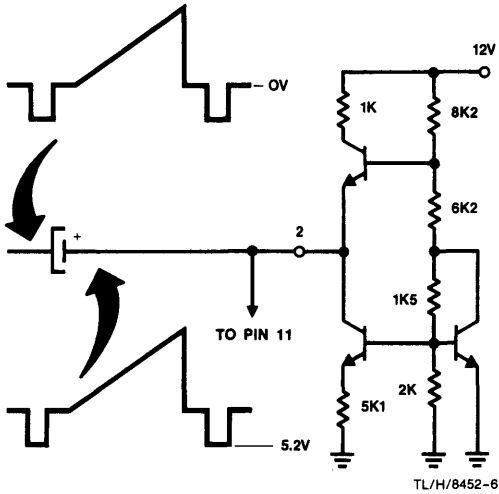


FIGURE 6. LM2889 Video DC Restoration Clamp

carrier is completely suppressed. As the offset voltage between pins 10 & 11 is increased, the carrier output level increases. With a 75Ω output load resistor, the conversion gain of the R.F. modulator is 20 mVrms/volt. A dc restoration circuit at pin 2 of the LM2889 allows the composite video to be ac coupled from the preceding stages, giving the designer flexibility in the video processing circuits (unless an LM1886 is being used as a video source, it is unlikely that the composite video dc level will be correct, even with dc coupled video sources). On a 12V supply, pin 2 clamps the sync tip of the video waveform to 5.1 VDC. Therefore, if we have a 2V (p-p) signal, one I.R.E. is equivalent to 14.3 mV and 160 I.R.E. is 2.29V. This is the required offset across the modulator input pins and since pin 11 will

be clamped to 5.1 VDC by the dc restorer circuit, pin 10 should be biased at $5.1V + 2.29V = 7.4$ VDC. A look at the R.F. carrier output will confirm that now the syncs occupy from 100% to 75% of the peak carrier, and that white modulates the carrier down to $12\frac{1}{2}\%$ of the peak. To maintain the proper modulation depth the clamp at pin 2 will track with supply voltage changes, allowing the bias at pin 10 to be set with a resistive divider connected between the supply and ground.

If the video signal polarity is reversed with positive syncs, either a dc coupled signal or an external dc restorer should be used that places the signal sync tip voltage towards the upper end of the common-mode input range at pin 11, which is 9 VDC with a 12V supply. Pin 10 is then offset below pin 11 voltage by the required amount for proper modulation. An input level of 2V (p-p) is optimal. Signal amplitudes of less than 1V (p-p) are also useable but internal offset voltages and the potential for carrier feedthrough or leakage to the output stage may make it difficult to maintain good R.F. linearity at peak modulation depths. Signal swings larger than 3V (p-p) should be avoided since this will produce relatively large AC/DC current ratios in the modulator and the resulting modulator non-linearities can cause a 920 kHz beat between the chroma and sound carriers.

Although only one video input is required, the LM2889 has two balanced R.F. modulators and two R.F. carrier frequency oscillators. Selection of the carrier frequency is by dc switching the supply voltage to the relevant oscillator tuned circuit. This automatically shuts off the other oscillator and modulator circuits. For test purposes when an output R.F. VSB filter isn't used, or when only one carrier frequency is needed, the output pins 8 and 9 can be wired together with a common load resistor. Providing two channel operation with two independent oscillator/modulator circuits is much superior to using a single modulator and attempting to change carrier frequency by switching the tuning components of a single L-C oscillator. The latter method involves

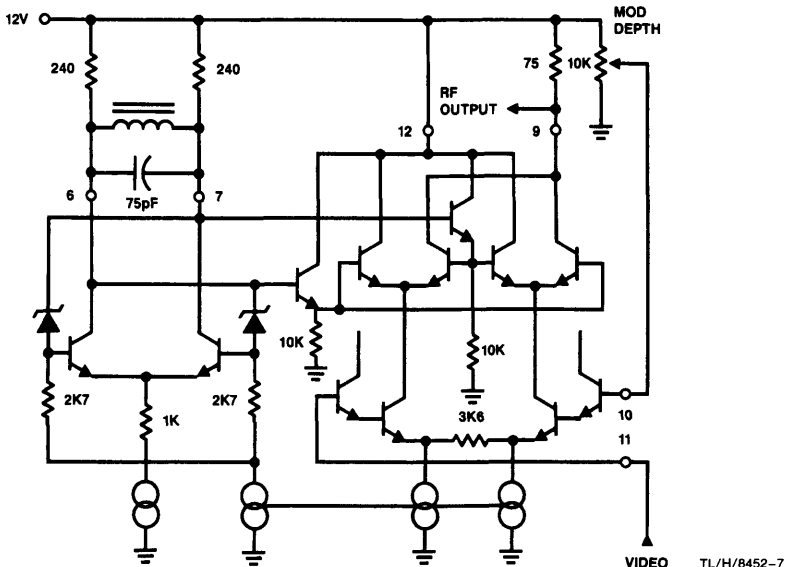


FIGURE 7. LM2889 R.F. Modulator and Oscillator (one channel)

Video Modulation (Continued)

use of isolating diodes (if unbalanced operation with attendant feed through problems is to be avoided) and expensive trimmer capacitors for tuning the second carrier frequency. A further disadvantage is the need to switch the VSB filter at the R.F. output.

The LM2889 oscillator configuration is the familiar cross coupled differential amplifier type, with level shifting zener diodes used to prevent the transistors from saturating with large oscillator output swings. The oscillator frequency is set by the tuned circuit components ($f = 1/2\pi\sqrt{LC}$), and the load resistors connected to the supply will set the oscillation amplitude and drive level to the modulators as well as determining the circuit working Q.

As might be expected, there are conflicting requirements on the practical range of working Q's. A high Q is desirable from the viewpoint of stability, but higher working Q's (set mainly by larger load resistors) increase the drive level to the modulator. Above 350 mV (p-p) the modulator will have attained full conversion gain and the R.F. output level will be determined by the amplitude of the video input signal. Unfortunately increased drive levels will also increase the carrier frequency second harmonic output from the modulator. Although a fully balanced design is used, parasitic capacitances on the emitters of the switching transistor pairs will rectify the oscillator waveform and this produces high levels of second harmonic. Load resistors much larger than 240Ω can produce a level of second harmonic matching the fundamental. Since relatively small load resistors are required (much smaller than the tuned circuit dynamic resistance) the working Q will be dominated by these resistors.

The acceptable degree of frequency stability will depend on the intended application, but L-C oscillators have proven to be adequate for most purposes. We can gain an idea of the frequency stability that is possible by considering the frequency drift produced by changes in the oscillator internal phase. A change in internal phase shift can be caused either by temperature or supply voltage changes but, as the LM2889 data sheet shows, the supply voltage dependency is low. Between 12V and 15V the frequency is essentially constant and changes by less than 30 kHz over the entire supply voltage range. With temperature, the internal oscillator phase shift changes by about 2 degrees over a 50 degree Celsius temperature range. If the tuned circuit Q is 15, then at 61.25 MHz (Ch 3 pix carrier) the oscillator frequency must change by -92 kHz to produce a compensating 2 degree phase shift. If the Q is 30, then the frequency would change by less than -45 kHz etc.

For high circuit Q, a large capacitance is desirable, but the inductor cannot be made too small if it is to remain the tuning element. This keeps the practical range of capacitance values to between 50 pF and 100 pF. Using a 75 pF capacitance, at 67.25 MHz the required inductance is just under 0.08 μH and the working Q is 15 with 240Ω resistors connected on either side of the tuned circuit to the supply voltage. Depending on the coil type, the number of turns for this inductance will be from 1½ to 3½ giving over 10 MHz tuning range. This is more than enough to compensate for component tolerance and variations in overall internal phase lag from I/C to I/C.

If better frequency stability of the carrier frequency over that provided by an L/C circuit is needed, then crystal control of the oscillators can be used. It is necessary to retain the inductor, since a dc short is required across the oscillator pins to avoid a collector current imbalance off-setting the

oscillator differential pair and preventing start-up. The inductor value is chosen to resonate with the capacitor in series with the crystal at slightly less than the desired operating frequency. About 20% less will allow the inductor to be fixed tuned. Close to its series resonant frequency (normally the 3rd overtone) the crystal will provide the additional inductive reactance necessary for the circuit to oscillate. The equivalent resistance of the crystal at the operating frequency will affect the tuned circuit Q and hence the peak-to-peak drive to the modulator circuit. Smaller capacitors in series with the crystal (with corresponding changes in the inductor value) will push the operating frequency closer to anti-resonance and produce large equivalent resistances dropping the oscillator drive level. Larger capacitance values cause the operating frequency to approach series resonance and a lower equivalent resistance (approaching R_S for the crystal, which is of the order of 40Ω to 100Ω at 60 MHz). This can produce higher drive levels but risks operation at the lower overtones. To prevent lower frequency oscillation a resistor can be connected across the crystal. Also a small resistor in series with one of the collector leads will form a low pass filter with the output capacitance and suppress spurious oscillations at higher frequencies. If this is needed, resistor values less than 30Ω should be used, so that dc offsets will not prevent the oscillator from starting. For the circuit of Figure 8, capacitor values between 20 pF and 56 pF, with the appropriate inductor value, work well with only slightly reduced oscillator drive compared to the conventional L/C circuit.

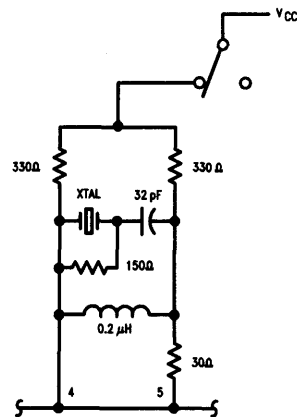


FIGURE 8. R.F. Crystal Oscillator Circuit
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The Sound Carrier Oscillator

Before moving to the R.F. output and the VSB requirements, we need to look at another signal that will be added to the baseband video—the aural intercarrier. Both the LM1889 and the LM2889 have L-C sound carrier oscillators operating at 4.5 MHz. Frequency modulation of the LM1889 sound oscillator is achieved by an external varactor diode which alters the tuning capacitance in response to the amplitude of the audio signal. The LM2889 has a similar tuned L-C oscillator but the frequency deviation is obtained by internally phase shifting the oscillator current. This is done by a low pass filter connected to the oscillator which provides a lagging phase voltage component of the oscillator waveform at the input to a differential amplifier. The current output from

Sound Modulation

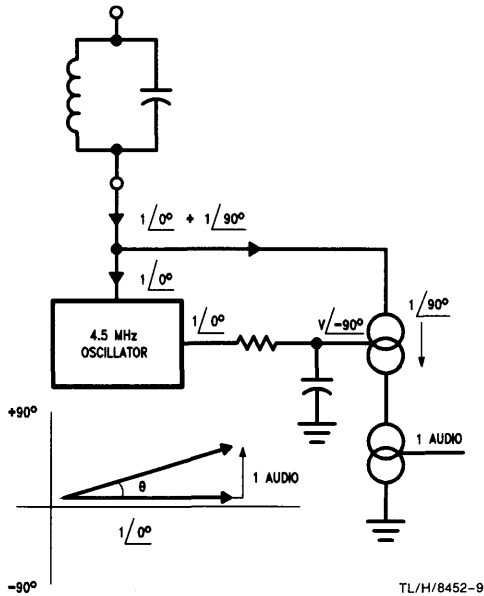


FIGURE 9. LM2889 Sound Carrier FM Modulator

this amplifier is controlled by the audio signal amplitude so that more or less of the current (now in quadrature to the original oscillator current) is added back to the tuned circuit producing the desired shift in the output frequency. Phase offsets of up to +12 degrees with increasing audio input levels will yield very low audio distortion (less than 0.2%). Also the use of a lagging oscillator waveform component reduces harmonic levels within the oscillator and a reduced possibility for undesired signals contaminating the R.F. waveform.

The tuned circuit operating Q is important in two respects. Similar to the R.F. oscillator tuned circuits, the 4.5 MHz tuned circuit should have a high loaded Q for stability, but the circuit bandwidth must also be wide enough to accommodate the FM sidebands produced by the audio modulation. For a maximum frequency deviation (Δf) and maximum modulating frequency f, the minimum bandwidth is given by Equation (1).

$$B-W \geq \Delta f (2.5 + 4f/\Delta f) \tag{1}$$

The other requirement is that the maximum phase deviation of the oscillator current is able to produce the maximum frequency deviation (Δf) of the carrier. This is given by Equation (2).

$$\Delta f = 4.5 \times 10^6 \times 0.12 / Q \tag{2}$$

Table I summarizes the results of calculating the maximum circuit Q that satisfies Equations (1) and (2) for the various monaural sound modulating standards used in the U.S. and Europe.

TABLE I

System	Δf	Modulation Bandwidth	Q_{max}		
			Modulation	Deviation	
USA	Mono	25 kHz	125 kHz	≤ 36	≤ 21
	Stereo	73 kHz	400 kHz	≤ 12	≤ 7.4
UK	50 kHz	200 kHz	≤ 30	≤ 15	
Continental Europe	30 kHz	150 kHz	≤ 36	≤ 22	

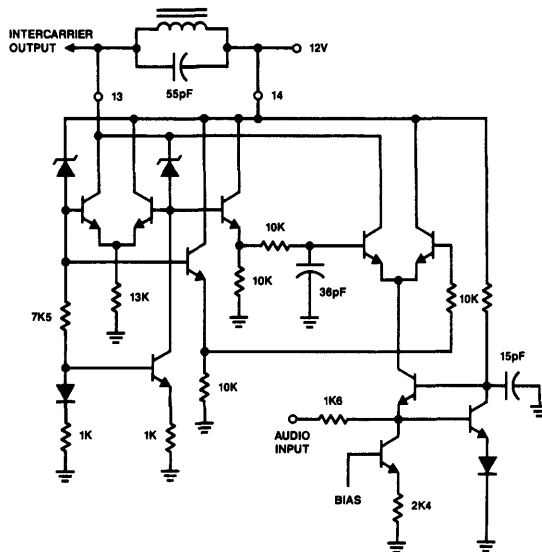


FIGURE 10. LM2889 4.5 MHz Sound Oscillator and Modulator

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Sound Modulation (Continued)

Clearly the deviation phase offset dominates the circuit Q requirement.

If we choose a Q of around 10 then the oscillator drift with temperature (assuming a 2 degree phase change in oscillator current with a 50 degree rise in temperature) is of the order of -9 kHz. A typical receiver will generate less than 3% distortion at peak deviations with this much frequency drift but if better performance is required, then the circuit Q can be raised. High modulation linearity will still be retained with a Q of 20 and the oscillator maximum frequency drift will be halved. Alternatively temperature compensated tuning capacitors can be used (between N20 and N75). When higher circuit Q's than 20 are employed, increased audio input levels will produce the desired peak frequency deviations but with the possibility of increased modulation distortion. The actual operating parameters that are selected can be balanced between distortion as a result of modulation, and distortion in the receiver circuits as a result of oscillator frequency drift.

To ensure that we have a sufficient 4.5 MHz oscillator level to provide enough drive to the internal phase shift circuit, the load impedance at pin 13 should be greater than 3.5 k Ω . A second requirement is that we have enough oscillator level to generate the desired aural carrier amplitude when modulated on the picture carrier. This means that load impedances greater than 6 k Ω are desirable. At 4.5 MHz, a typical oscillator coil of 23 μ H will have an unloaded Q of 55 and tune with 55 pF. For a working Q of 10, the external damping resistor is 7.5 k Ω .

Stereo Sound

The introduction in the U.S. of a multiplex stereo sound system (the BTSC system combining the Zenith MCS proposal with dbx noise reduction in the stereo difference channel) with peak carrier deviations in excess of 73 kHz puts even larger constraints on the tank circuit Q. Following the same rules as before, the maximum allowable Q for low distortion is now less than 7.4 —with a loaded Q of 5 being likely. With this loaded Q, maintaining a carrier center frequency accuracy better than 5 kHz with an L/C circuit becomes impractical

and other methods to set the oscillator frequency must be used. Since a crystal will provide the necessary temperature and voltage stable reference frequency a PLL is a useful solution (see Figure 11). Either the widely available 3.58 MHz crystals or a 4.5 MHz crystal can be used, but in either case, the L/C tank circuit frequency must be divided down before application to the phase detector. This is because frequency modulation of the sound carrier will produce many radians of phase deviation at the phase detector input—for a modulation frequency of 100 Hz and a peak deviation of 73 kHz the carrier phase change is given by Equation 3.

$$\theta = \Delta f/fm = 73 \times 10^3/100 = 730 \text{ rads} \quad (3)$$

Since the linear input range of most phase detectors is less than 2π radians, the modulated carrier input must be divided down by at least 233 to keep the phase deviation within this linear range. For a 4.5 MHz crystal, the reference frequency divider M and the sound oscillator divider N are the same. Available ripple counters such as the 74HC4040 and 74HC4060 can easily divide by 128 (for monaural) or by 256 for stereo. If a 3.58 MHz crystal is used the M:N divider ratio is 35:44 requiring substantially more packages, and the odd numbered divider must be followed by an even divide of 2 or 4 to "square up" the input waveform to the phase detector. Also, since the video will include a chroma subcarrier, good isolation is needed to prevent the reference oscillator beating with the chroma sidebands.

A suitable phase detector is the 74C932 Exclusive-Or type with a sensitivity of 1.6 volts/radian. The filter at the detector output prevents the input modulation from reaching the varactor diode and distorting the audio. Even so, the loop filter must have some ac bandwidth for a reasonable acquisition time and other dynamic characteristics. The components shown in Figure 11 have been chosen such that with a varactor sensitivity of 100 kHz/volt the loop has a hold-in range of over ± 150 kHz, with a lock-up time of less than 0.5 seconds. The T.H.D. is less than 1% for a 400 Hz modulating frequency producing 25 kHz deviation of the carrier. The accuracy of the sound carrier frequency is, of course, that of the crystal used for the reference oscillator.

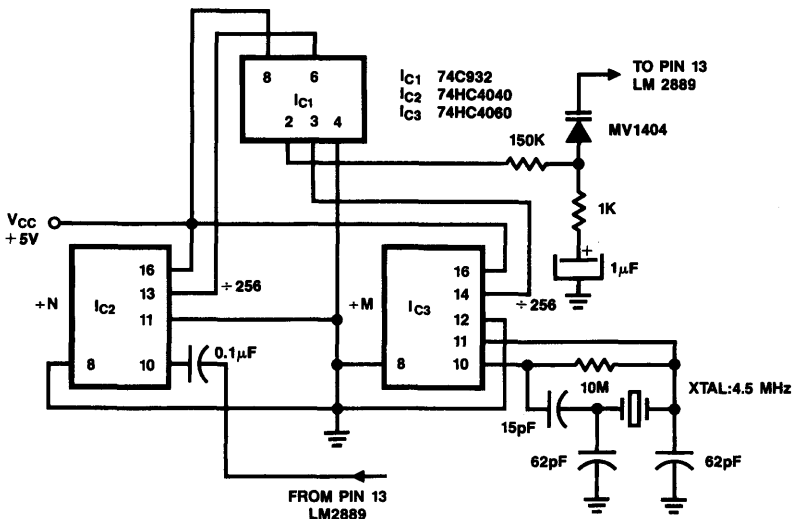


FIGURE 11. 4.5 MHz Crystal Reference Circuit

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Audio Processing For Sound Carrier Modulation

With the proper tuned circuit Q (see Table I), a linear increase in the amplitude of the audio signal will produce a correspondingly linear increase in the frequency deviation. Television receiver sound circuits in the U.S. have a 75 μ s de-emphasis and in Europe frequencies above 3.2 kHz (50 μ s) are de-emphasized at a 6 dB/octave rate. This is done to help improve the S/N ratio of FM reception and the transmitter incorporates the complementary pre-emphasis characteristic—above 2.1 kHz the audio frequencies are boosted at a 6 dB/octave rate. The consequence of this modulation scheme is that if a 0 dB peak signal amplitude at 15 kHz is capable of producing a 25 kHz deviation than a similar amplitude signal at 400 Hz will produce a peak deviation of only 3 kHz—a loss of some 18 dB in S/N ratio for the midband frequencies. Broadcasters usually employ compressors to enable high modulation levels to be obtained at mid-band frequencies without overmodulating high frequencies. If the audio input to the LM2889 is being sourced from an original broadcast (a scrambled signal decoder output for example) than this audio—without de-emphasis—can be directly applied to pin 1 of the LM2889, and the overall input level is adjusted so that the modulation limits are not exceeded except for brief intervals (less than 10 instances per minute). When the audio has not already been processed a different set of conditions will apply and an audio pre-emphasis network is required at pin 1.

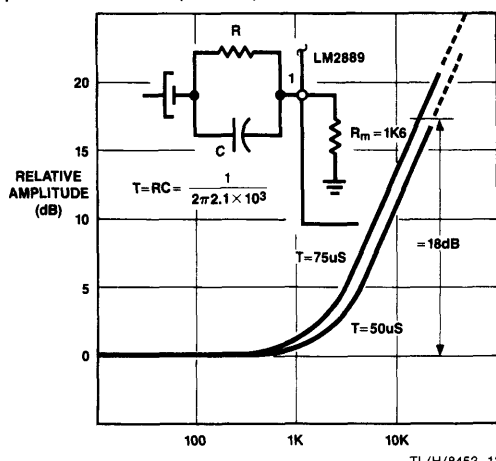


FIGURE 12. Audio Pre-emphasis

Since the audio source is likely to be at a relatively low impedance (a pre-amplifier output), the pre-emphasis network will also be used to attenuate the level of the average audio input to the LM2889 as well as providing a relative boost to the higher frequencies. The input sensitivity of the audio modulator is 150 Hz/mV which means that 118 mVrms will give a peak deviation of 25 kHz.

Next we have to decide what signal frequency and amplitude to use in calibrating the audio input. Unfortunately the 75 μ s time constant for FM broadcasting was chosen at a time when equipment limitations meant there was relatively low spectral energy at higher frequencies. Today, modern audio material is not well suited to boosting above 2.1 kHz since energy peaks at only -6 dB can be obtained at 10 kHz. A further complication is the ability of the audio level meter to predict high energy peaks. If a conventional VU

meter is used, peak levels of +10 dB are possible while the meter is indicating OVU. Obviously without processing the audio to keep it within predetermined limits, the input level calibration will be somewhat empirical in nature.

If we assume the decrease in spectral energy above 10 kHz is such that overmodulation peaks above this frequency are unlikely to occur, then we can allow a signal at 10 kHz to produce full modulation deviation. Since the amplitude of most audio signals at 10 kHz is at least 6 dB below the midband frequency level, we can calibrate the audio input with a -6 dB amplitude, 10 kHz tone to produce 100% deviation. As we shall see later, a frequency close to 10 kHz will make the measurement of actual peak deviations very easy indeed. With the standard pre-emphasis network, at signal frequencies less than 2 kHz, the modulating signal amplitude at pin 1 will be -8 dB below the anticipated peak 10 kHz level producing 100% modulation. This corresponds to a modulator input level of $118/2.2 = 45.4$ mVrms. The

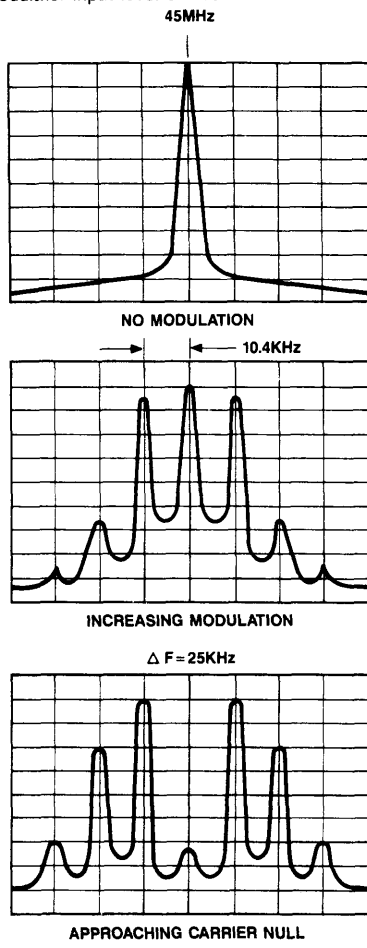


FIGURE 13. FM Spectrum with Increasing Audio Amplitude (f mod = 10.4 kHz) 4.5 MHz Sound Carrier Level

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Audio Processing For Sound Carrier Modulation (Continued)

input resistance at pin 1 is 1.5 k Ω so $R_1 = 30$ k Ω , if we assume an input source level of 1 Vrms at 400 Hz. For a 2.1 kHz breakpoint, $C = 0.0027$ μ F.

Anyone who has observed the output from an FM circuit with a spectrum analyzer will know that for a fixed modulating frequency the output spectrum will consist of the carrier frequency component and sidebands spaced by the modulating frequency from the carrier. As the modulation amplitude is increased (the modulation index m becomes larger), the carrier decreases to a null and then increases again. The modulation indices for which carrier nulls occur can be calculated and for our purposes it is important to know that the first carrier null occurs at $m = 2.4048$. For a system maximum deviation of 25 kHz the modulating frequency f is given by:

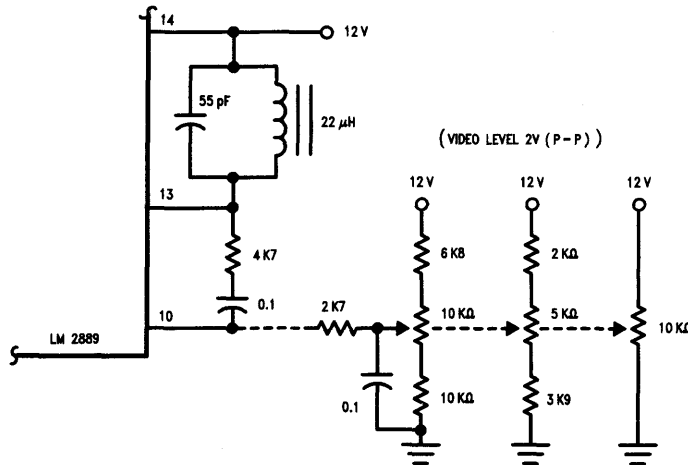
$$f = 25 \times 10^3 / 2.4048 = 10.4 \text{ kHz} \quad (4)$$

Therefore, if we use an input frequency of 10.4 kHz, as the input amplitude is increased the first carrier null will indicate peak deviation. If we continue with our assumption of a -6 dB level at 10 kHz, calibration consists of adjusting the audio input so that a -6 dB, 10.4 kHz signal causes the first carrier null. With the above pre-emphasis network, this should correspond to 500 mVrms at 10.4 kHz.

We have already looked at the tuned circuit parameters at pin 13 in terms of deviation linearity and oscillator stability. With a working Q of 10, the effective load at pin 13 is 6.2 k Ω . The oscillator current is 0.45 mA so that the output amplitude at 4.5 MHz is 3.6V (p-p). Some portion of this oscillator signal level is coupled over to pin 10 to set the sound carrier level and this can be done by splitting the external 7.5 k Ω damping resistor into two parts. The picture carrier level is set by the offset voltage between pins 10 and 11 as described earlier. For a 2V (p-p) video signal this offset is 2.3V. Since the 4.5 MHz signal will be ac coupled over to the bias pin 10, it will amplitude modulate the picture RF

carrier. This is conventional AM and a 4.6V (p-p) signal will yield sound carrier sidebands at -6 dB relative to the picture carrier. If we require a sound carrier amplitude at -17 dB, the signal coupled to pin 10 must be 11 dB below 4.6V (p-p), or 1.3V (p-p). This is obtained by using a 4.7 k Ω resistor coupled through a 0.1 μ F capacitor to pin 10, and a second 2.7 k Ω resistor connected to the wiper arm of the potentiometer used to set the video modulation depth. The effect of the potentiometer setting on the aural carrier level is eliminated by a 0.1 μ F capacitor connected from the wiper arm to ground. However, since the impedance presented by the potentiometer will, for all practical purposes, be relatively constant, the capacitor could be removed and the parallel resistance of the upper and lower arms of the potentiometer network used to provide the second resistor of 2.7 k Ω . If the video input level is well controlled, it may be possible to replace the potentiometer with a fixed divider.

The final part of the design concerns the output stage, and involves meeting the constraints applied by any regulatory agency. In the U.S., apart from the need to restrict the peak carrier output level to less than 3 mVrms in 75 Ω , we have two signals present in the output whose level will exceed the spurious emission limit of -30 dB with respect to the peak carrier level. One of these signals is the result of amplitude modulating the 4.5 MHz intercarrier audio on the picture carrier. Apart from the desired -17 dB sound carrier amplitude (upper sideband) an equal amplitude lower sideband will be present. For channel 3 this is at a frequency of 56.75 MHz—which is 250 kHz outside our channel lower limit. Therefore we need to provide at least 13 dB more attenuation at this frequency in the output filter. The second unwanted emission (or emissions) is the result of carrier frequency harmonics—specifically the 2nd harmonic level produced by high modulator drive. To suppress this, from -18 dB to -30 dB attenuation at 123 MHz is required.



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FIGURE 14. Audio Intercarrier Coupling to the Video Modulator R.F. Output and V.S.B. Filter

Audio Processing For Sound Carrier Modulation (Continued)

With a properly constituted baseband signal modulating the carrier, these are the only intrinsic unwanted emissions we are concerned with. Normal video modulation components appearing in the lower sideband will not have sufficient amplitude and do not extend beyond the lower channel limit. Even so, the filter requirements are not trivial.

If L-C filters are used, this can be done with three coils per channel but some alignment procedure will be required. Fortunately SAW filters are available from several sources which, although more expensive than the equivalent L-C filter, avoid the cost of production alignment. Usually the SAW filter will have a substantially greater insertion loss, but the LM2889 has enough output level to compensate for this. Both single channel and dual channel filters are available and in the latter case the LM2889 dual oscillator/modulator configuration enables easy dc switching between channels. A coil may be required, connected across the SAWF input, to tune out the SAWF input capacitance.

The load resistors connected to pins 8 and 9 will set the LM2889 conversion gain, which for 75Ω is typically 20 mVrms R.F. carrier per volt offset at the input pins 10 and 11. The actual load will include the input resistance of

the filter. Since the output of the filter will normally be terminated in 75Ω to match the cable (and provide triple transit echo suppression for a SAWF), the best way to choose the load resistor is to monitor the output to the cable and apply a dc offset between pin 10 and 11 that is equivalent to the expected video input. The resistor is then chosen to give the desired peak carrier level of 2.5 mVrms. The carrier should be unmodulated since downward modulation will reduce the mean carrier level by as much as 2–3 dB.

If the offset voltage between pin 10 and 11 is reduced, a check can be made on the residual carrier level at the output. This residual level is the result of oscillator feedthrough in the modulators and external coupling from the oscillator tuned circuits. The residual carrier level is normally better than -26 dB below the peak carrier level, ensuring good modulation linearity. High levels of residual carrier can be caused by coupling through ground or power supply leads. A good technique to minimize the effect of unwanted pick-up is to decouple the supply voltage to pin 8 and 9 load resistors over to the output connector shield ground. This removes at the output any carrier signal on the supply line to the load resistors.

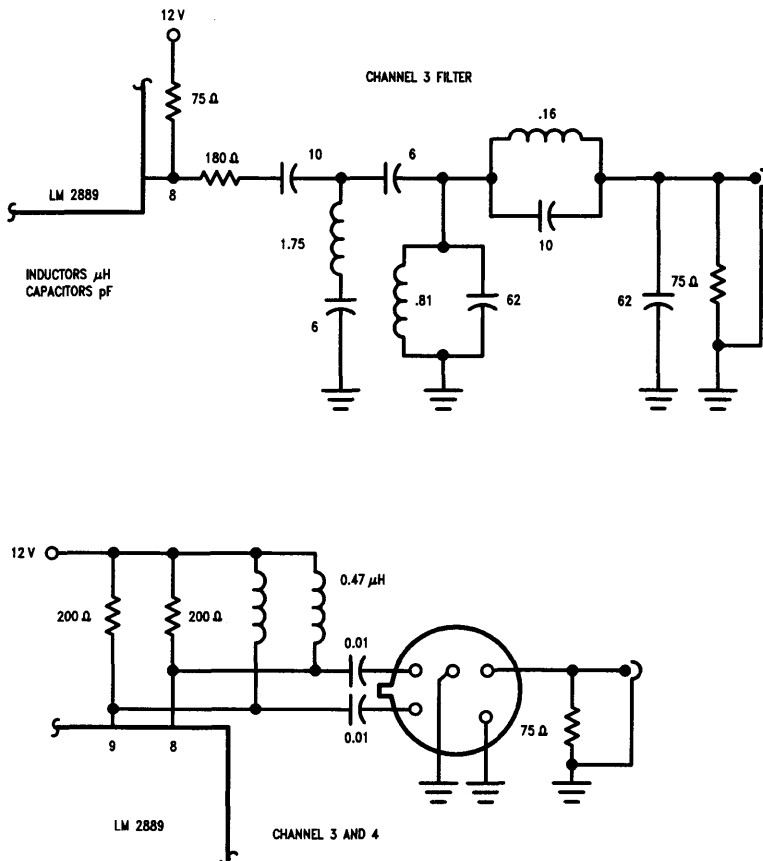


FIGURE 15. Vestigial Sideband Filters

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Audio Processing For Sound Carrier Modulation (Continued)

Sources: SAWFs

Crystal Technology, Inc.
1035 E. Meadow Circle
Palo Alto, CA 94303

Kyocera International, Inc.
8611 Balboa Ave.
San Diego, CA 92123

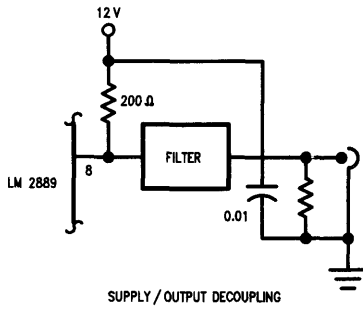
MuRata Corp. of America
1148 Franklin Rd. S.E.
Marietta, GA 30067

CRYSTALS

Saronix
4010 Transport at San Antonio Rd.
Palo Alto, CA 94303

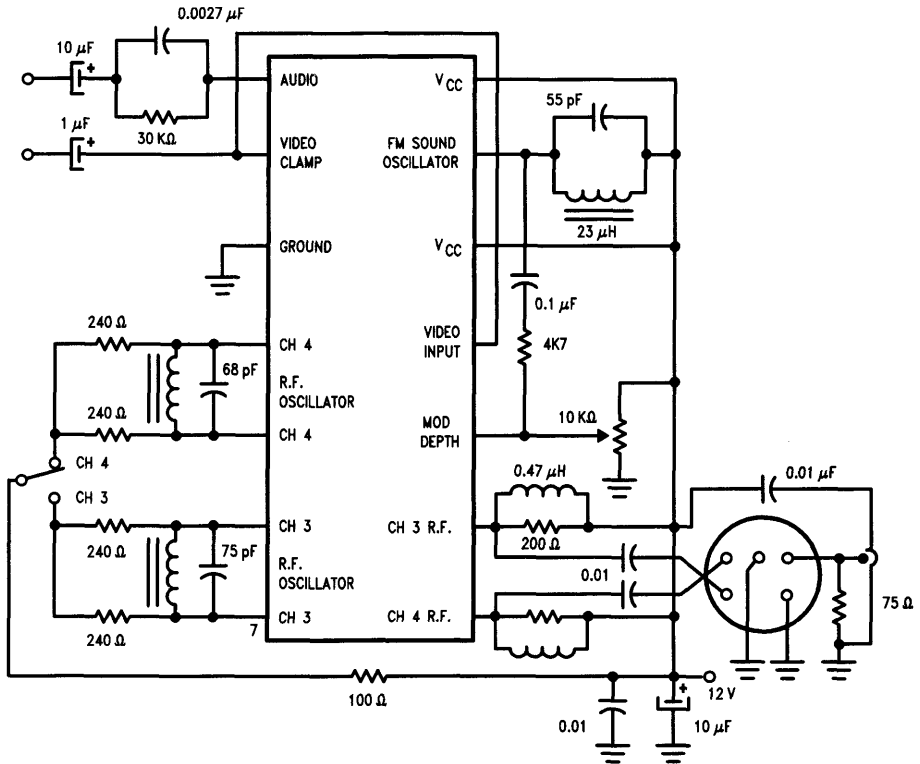
COILS

Toko America, Inc.
5520 W. Touhy Ave.
Skokie, Ill. 60077



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FIGURE 16. R.F. Decoupling at the Output



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FIGURE 17. Complete R.F. Modulator External Circuit